Solving (2) and (3) for h^2 and $s_{1,2}^2$ leads to

$$h^2 = \epsilon_1 \pm \sqrt{\epsilon_1(\epsilon_1 - 1)}, \tag{4}$$

and

$$s_{1,2}^2 = \frac{1}{2\epsilon_1}\left[(\epsilon_1 + \epsilon_3)(h^2 - \epsilon_1) + \epsilon_2^2 \pm \epsilon_2\sqrt{2(\epsilon_1 + \epsilon_3)h^2 - \epsilon_1 + \epsilon_3}\right]. \tag{5}$$

For ω smaller than $\sqrt{\omega_p^2+\omega_H^2}$, h^2 given by (4) is always real, and s_1s_2 is also real from (2) and (3). For this case the signs ($\pm$) in (4) corresponds to $\epsilon_1\epsilon_3 \lesseqgtr 0$ if $s_1s_2>0$, and $\epsilon_1\epsilon_3 \gtreqless 0$ if $s_1s_2<0$, respectively.

When s_1s_2 is negative, real part of either of s_1 and s_2 is negative in general, therefore the fields increase to infinity at $x\to\infty$. In Figs. 1 and 2 the frequency characteristics of the relative propagation constants for these improper modes are plotted by the dotted lines.

When s_1s_2 is positive, and in the frequency range of

$$\omega_p < \omega < \sqrt{\omega_p^2 + \omega_H^2}, \tag{6}$$

either of s_1 and s_2 is a positive imaginary and the other is a negative imaginary. This implies that the fields are merely the superposition of two nonattenuating plane waves, a left-handed and a right-handed plane wave, one being incident on the conducting plane and the other reflecting from the plane. The angle of incidence is different from that of reflection since $|s_1| \neq |s_2|$. The frequency characteristics of h are plotted by dashed lines in Figs. 1 and 2.

In the frequency range of

$$\omega < \text{Min}(\omega_H, \omega_p), \tag{7}$$

the electromagnetic wave is found to be trapped along the conducting plane. The relative propagation constant h is given by $[\epsilon_1+\sqrt{\epsilon_1(\epsilon_1-1)}]^{1/2}$ and is plotted by the solid lines in Figs. 1 and 2. For this case s_1 and s_2 are both positive real numbers or conjugate complex numbers with positive reals, therefore the equiphase surface is vertical to the guiding plane. The group velocity v_g of this trapped wave is obtained as

$$c/v_g = h\left[1 + \left(1 - \frac{1}{2h^2}\right) \cdot \frac{\omega_p\omega^2}{(\omega_H^2 - \omega^2)(\omega_p^2 + \omega_H^2 - \omega^2)^{1/2}}\right], \tag{8}$$

where c is the velocity of light.

Fig. 3 illustrates the relations between the propagation constant h discussed above and the permissive z-directed propagation constant of nonattenuating left-handed and right-handed plane waves in a free magnetoplasma. It is seen from this figure that there can exist no right-handed and left-handed plane waves propagating with the axial phase velocity equal to that of the trapped wave in the region of $\omega < \text{Min}(\omega_p, \omega_H)$. The incident and reflecting angles of the left handed and right-handed waves take imaginary values in the trapped-wave region.

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A Multistub Coaxial Line Tuner

Introduction

To date, one of the major difficulties in precision coaxial line reflectometer work has been the lack of suitable coaxial line tuners. The tuner described here was designed specifically for reflectometer work and overcomes the major problems of mechanical instability, coarse tuning adjustments, and leakage prevalent in presently available coaxial line tuners.

Discussion of Tuner

The geometry of the tuner is essentially that of a parallel plane transmission line[1] where the ends of the parallel plates have been closed at a distance sufficiently far from the center of the line so as not to alter appreciably the characteristics of the parallel plane line.

The electric field configuration in this type line is shown in Fig. 1. The field is very strong in the narrow gap between the center conductor and the side wall and is very weak in the large region between the center conductor and the top. This condition is represented by the high concentration of lines in the narrow gap and low concentration of lines in the large region. To get the desired tuning operation, one set of tuning stubs has been placed in the side wall directly across from the center conductor for operation in the region of the concentrated electric field and another set of tuning stubs has been placed at the top of the tuner for operation in the region of the weak field. The location of these tuning stubs is shown in Fig. 2. The stubs in the region of the concentrated electric field give a coarse tuning operation, and the stubs in the region of the weak field give a fine tuning operation.

The tuning stubs are threaded through a collet-type lock to obtain smooth stub adjustment and to prevent leakage to the outside of the line. Also, a groove has been cut completely around the side plate and a woven gasket has been placed in this groove to prevent any leakage from around the side plate. This groove is shown in Fig. 3.

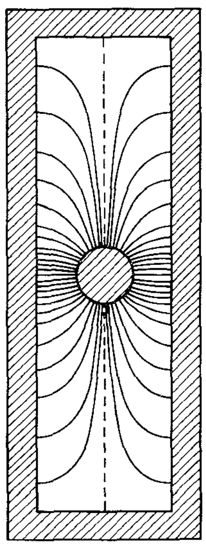


Fig. 1—Electric field configuration in the enclosed parallel plane line.

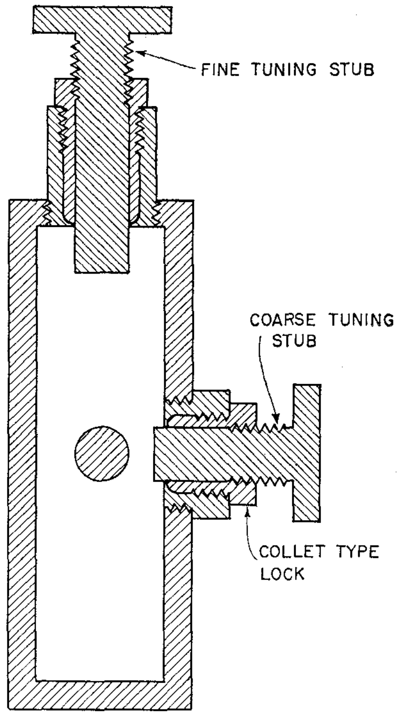
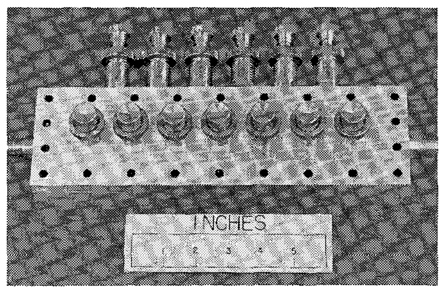
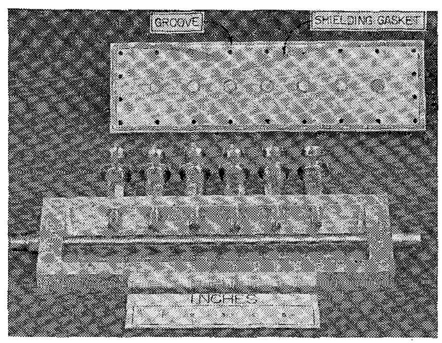


Fig. 2—Cross section of tuner showing location of tuning stubs and collet-type locks.



(a)



(b)

Fig. 3—(a) Photograph of tuner. (b) Photograph of tuner.

Results

Fig. 3 shows a photograph of a tuner designed to tune out a reflection coefficient of 0.33 (VSWR of 2 to 1) over the frequency

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[1] W. B. Wholey and W. N. Eldred, "A new type of slotted line section," Proc. IRE, vol. 38, pp. 244–249; March, 1950.

range of 1 to 4 GHz. The tuner was checked for tuning range with a slotted line and actually has the capacity to tune out a 2 to 1 VSWR down to 0.5 GHz. The tuner was tested for reflectometer applications in a combined waveguide-coaxial reflectometer system. The directional coupler of the reflectometer could easily be tuned for directivities of greater than 60 db.

The leakage from the tuner was checked with a sensitive receiver and the leakage to the outside was down by more than 85 db from the coaxial line power level.

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Orientation of YIG Spheres for Minimum Temperature Dependence

Due to magnetocrystalline anisotropy energy, YIG spheres display a resonant frequency shift with temperature variation. By orienting the sphere along certain directions relative to the applied dc magnetic field, this frequency shift with temperature can be minimized. Clark, Brown, and Tribby [1] have analyzed this problem, and resorted to X-ray alignment using the Laue back reflection pattern to obtain the temperature-stable orientation. They have pointed out the difficulty in maintaining the accuracy of alignment in transferring the sphere from the X-ray apparatus to the RF structure.

It is possible to do away with the costly and time consuming X-ray orientation method, and align the YIG spheres directly in the RF structure intended for use. Between the limits of 1680 and 3360 Mc, a YIG sphere operates in the coincidence limiting region and saturates at input power levels in excess of −10 dbm [2]. If the garnet sphere is not temperature aligned, applied power levels above the limiting threshold will result in heating of the sphere and hence in a resonant frequency shift.

Fig. 1 shows the laboratory setup. The sweep frequency generator is adjusted to cover a range of a few hundred megacycles above and below the resonant frequency, at a power level well below the limiting threshold. The high power CW signal generator is adjusted for the resonant frequency. A coaxial switch allows for rapid connection of the signal generator or sweep-frequency generator to the garnet sphere.

Initially, the sweep-frequency generator is connected and the resonant frequency f_0 is noted on the oscilloscope. Then the high power signal is applied for a few seconds and the sweep generator is reconnected. If the sphere is not oriented for minimum temperature dependence, the resonant point will appear displaced, and then will slowly drift back to its original position. The entire procedure can be repeated rapidly, since only the sphere is heated, and its mass is small. By rotating the post to which the garnet sphere is attached, an orientation can be located such that the resonant frequency is not displaced after high powers are applied. Because of the anisotropy, the magnetic field must be adjusted to establish resonance at f_0 after any change in orientation of the garnet. Otherwise, the high power generator will not be tuned to the resonant frequency of the garnet sphere.

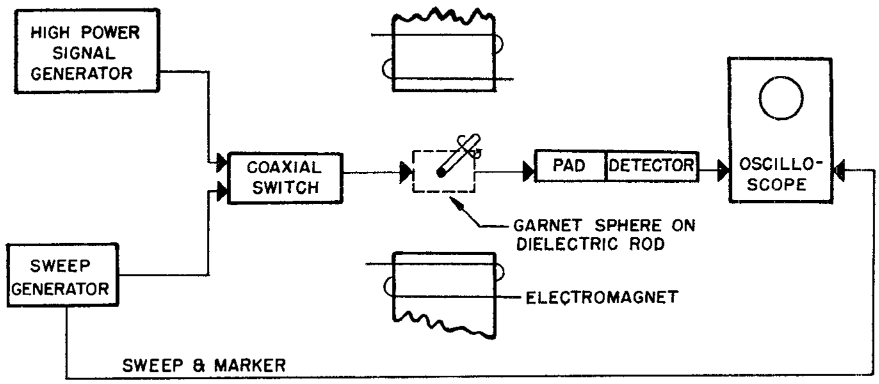

Fig. 1—Test setup.

The above technique can be extended to multiple-stage YIG devices. By means of an RF probe, the first stage is temperature aligned. The following stages are then synchronously tuned with the first.

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[2] B. Lax and K. J. Button, "Microwave Ferrites and Ferrimagnetics," McGraw-Hill Book Co., Inc., New York, N. Y., p. 679; 1962.

An Empirical Formula for the Design of Radial Line Filters

The present wide interest in varactor parametric amplifiers and frequency multipliers has led to renewed interest in bandpass and band-stop filters suitable for microwave frequencies. In negative resistance amplifiers it is essential to prevent leakage of pump and idler frequencies into the signal circuit and in multipliers it is necessary to prevent outputs at unwanted higher harmonic frequencies. In many such instances simple radial-line band rejection filters are adequate, combining high reflection within the rejection band with low insertion loss at other frequencies.

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A great deal of useful information for the design of such filters is given by de Loach[1] and the present note reports a simple extension of his work. All measurements were carried out using standard $\frac{3}{8}$-inch air-filled coaxial line (50 ohms) with dimensions

Inner Conductor O/diameter
0.120 inch ($=2a$)
Outer Conductor I/diameter
0.276 inch ($=2b$).

A cross section through a typical radial line cavity is shown in Fig. 1. From de Loach[1] (and intuitively), it is expected that for a fixed cavity height h the cavity diameter d will be inversely proportional to frequency. Measurements were carried out on cavities of fixed height (0.125 inch), but varying diameters, having rejection frequencies in the K-band region (12–18 Gc), and the results were graphed against reciprocal/linear scales. As expected, the points were scattered about a straight line, and the line of best fit was obtained by the method of least squares. The equation of this line, relating cavity diameter to resonant (rejection) frequency, is

$$d = 0.18\left(1 + \frac{46.51}{f_0}\right). \qquad (1)$$

where d is the cavity diameter in inches and f_0 is the rejection frequency in Gc.

It must be emphasized that this is a purely empirical relationship and as such is valid only for air-filled cavities of height 0.125 inch, with input lines as specified above. Further work is being carried out to determine the relationship between h and for fixed d.

Eq. (1) has been tested at frequencies below K band, and is accurate, giving diameters to within ±0.003 inch down to 8 Gc. With less accuracy (±0.006 inch) it is useable down to 5 Gc: below this frequency air-filled cavities become inconveniently large, and dielectric filling, as described by de Loach, should be adopted.

At high frequencies, application of (1) is limited by the onset of multimode propagation, which for the coaxial line dimensions quoted occurs above 19 Gc: up to this limit (1) may be used with confidence.

All cavities used in these experiments

[1] B. C. de Loach, "Radial line coaxial filters in the microwave region," IEEE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-11, pp. 50–55; January, 1963.