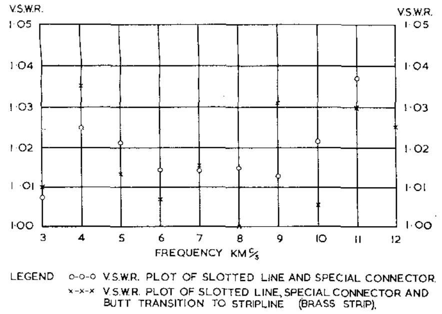


Fig. 4—VSWR plot from 3 to 12 kMc of the compensated butt coaxial-to-stripline transition.

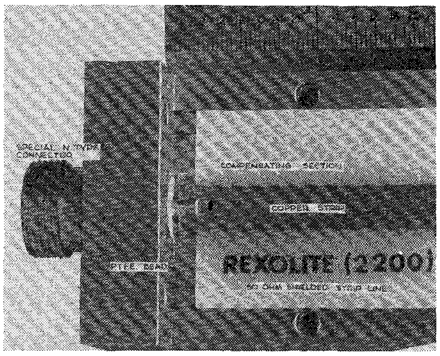


Fig. 5—Compensated coaxial-to-stripline butt transition.

With careful manufacture and use of the special N-type connector junction it has been found that results consistent with those given in Fig. 4 may be obtained to within 1.005 VSWR.

The compensated butt junction described above is easily made and lends itself particularly to interdigital stripline filters.[3]

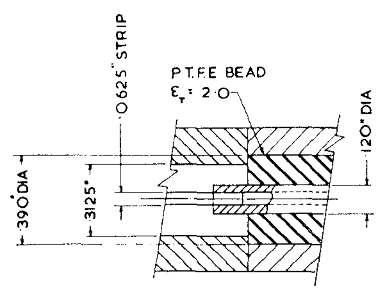
Stripline with Copper Clad Rexolite Card

This type of stripline [Fig. 1(b)] consists of a copper clad rexolite card between two parallel flat brass plates. The copper is removed to leave two central strips slightly narrower than the solid brass strip discussed previously. The characteristic impedance of this line is also 50 ohms.

The compensated butt junction is shown in Fig. 5; details of the junction region appear in Fig. 6. The connection of this type of junction to the coaxial slotted line is similar to that for the first type of transition (Fig. 3).

Since the rexolite card is not as rigid as a solid brass strip it was necessary to slot the inner conductor of the coaxial line to receive the card. This was found to provide a rugged connection leading to repeatable experimental results. In the region of the overlap the capacity per unit length of line is increased and it was found necessary to remove the rexolite almost to the copper strip in compensation.

The optimum performance was obtained when the inner conductor of the coaxial



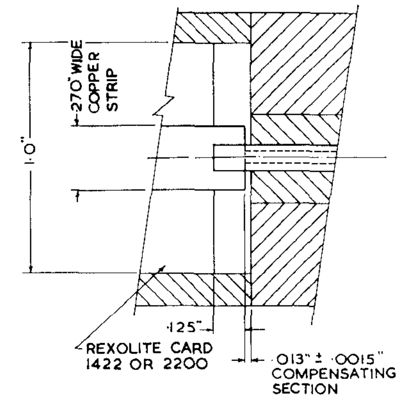


Fig. 6—The compensating section for the transition to rexolite supported stripline.

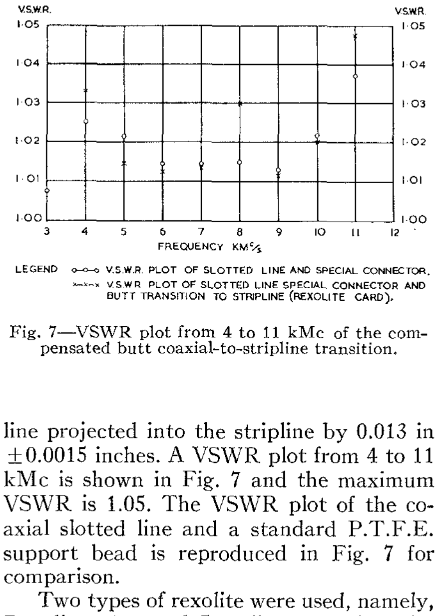


Fig. 7—VSWR plot from 4 to 11 kMc of the compensated butt coaxial-to-stripline transition.

line projected into the stripline by 0.013 in ± 0.0015 inches. A VSWR plot from 4 to 11 kMc is shown in Fig. 7 and the maximum VSWR is 1.05. The VSWR plot of the coaxial slotted line and a standard P.T.F.E. support bead is reproduced in Fig. 7 for comparison.

Two types of rexolite were used, namely, Rexolite 1422 and Rexolite 2200. Samples of each from different batches were compared and the maximum VSWR deviation at any frequency was 1.015. Reasoning as for the first type of transition we may conclude from Fig. 7 that the maximum VSWR of the coaxial-to-stripline transition alone is probably better than 1.025.

This is another type of easily manufactured butt transition and would be useful for direct coupled resonator filters[4] using capacitive gaps in the copper strip.

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Experimental Verification of the Phase Relationships in Parametric Amplifiers

The purpose of this communication is to present experimental evidence to show that the phase of the amplified signal in a three-frequency parametric amplifier is independent of the phase of the pump, while the phase of the idler varies directly with the phase of the pump.

The frequency relationship, $f_p = f_s + f_i$ [1] and the phase constant relationship for traveling-wave structures $\beta_p = \beta_s \pm \beta_i$ [2] are the well-known conservation of energy and momentum equations, respectively, which govern the behavior of parametric amplifiers.

However, after the discovery of the parametric amplifier [3], it was found that the degenerate parametric amplifier was unique in that there was an additional relationship required concerning the phase of the signal frequency to the phase of the pump frequency [4]–[7]. In this case maximum gain occurred when $\Theta_s = \Theta_p/2$, minimum gain occurred when $\Theta_s = \Theta_p/2 + \pi/2$, (where Θ_s, Θ_p are the signal and pump phases, respectively) and intermediate gains occurred at intermediate phases [8]. The nondegenerate parametric amplifier was found apparently to amplify signals independent of the pump phase.

Analyses by Heffner and Wade [9], by Tien [10], and by others have shown that the amplified signal output is independent of the pump phase, but that the idler current (output) is dependent on the pump phase. Thus, it has been stated that $\Theta_p = \Theta_s + \Theta_i$, where Θ_i is the phase of the idler current (or voltage). The following results confirm this prediction.

A parametric amplifier was used which could function both as a one-port reflection-type parametric amplifier and as a two-port frequency converter parametric amplifier. The amplifier used was originally developed by Bossard and Pettai [11]. When operated in the reflection mode, it gave the following results: signal band, 2000 to 3000 Mc; gain, 9.5 db; 3-db bandwidth, 830 Mc; noise figure, 1.8 db; pump frequency, 12,250 Mc [12].

In order to measure the phase of the amplifier signal with respect to the phase of

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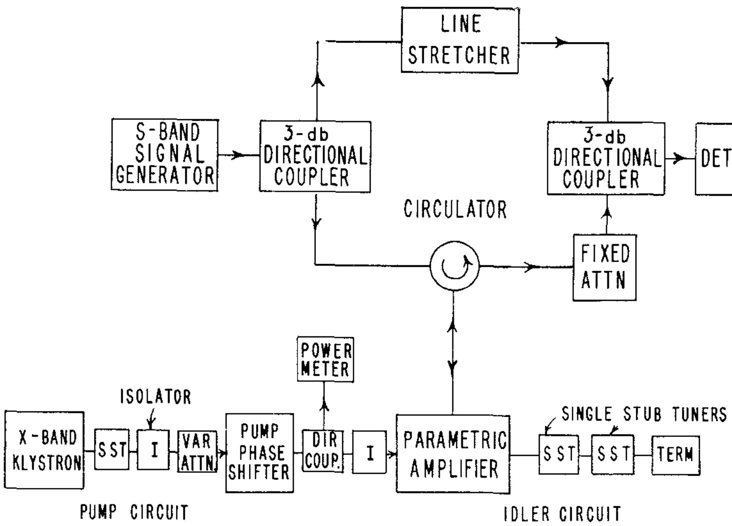

Fig. 1—Block diagram of microwave bridge circuit used to detect any signal phase change as a function of pump phase change.

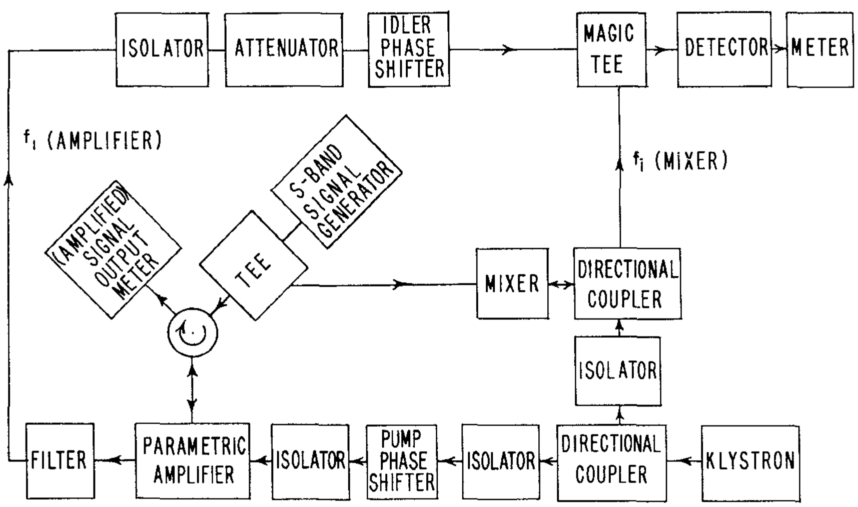

Fig. 2—Simplified block diagram of bridge circuit used to measure the variation of the parametric amplifier idler phase as a function of pump phase.

the input signal, a bridge circuit was set up as shown in Fig. 1. The input signal was initially amplified by the parametric amplifier with approximately 6-db gain which was measured to be at least 3 db below the threshold of oscillation. This amplified signal was then balanced out on the bridge using the line stretcher (for phase variation in the input line) and a fixed 6-db attenuator and a null reading was obtained using a Rubicon galvanometer with a sensitivity of 0.01 μa/mm. The pump phase was varied using the phase shifter and the galvanometer reading was noted for 10° pump phase shift intervals. No change was noted in the galvanometer, thus requiring no phase adjustment of the input signal circuit which indicated that the amplified signal was independent of changes in the pump phase, within the sensitivity of this experiment.

In order to measure the idler current output and to determine how this varied with pump phase, another bridge circuit was set up as shown in Fig. 2. However, in order to balance out the idler current from the parametric amplifier, an additional

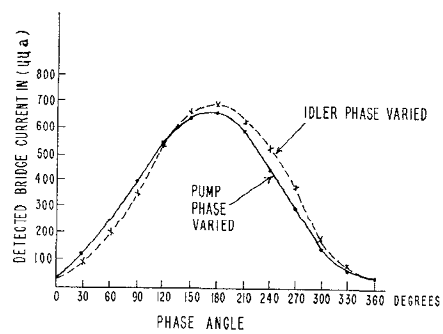

Fig. 3—Graphs of unbalanced bridge current as a function of (idler or pump) phase angle.

"input" or "standard" idler current was required. Since the input was at the signal frequency and not at the idler frequency, a "standard" idler current was generated by mixing part of the input signal frequency and the pump frequency in a microwave crystal rectifier. The input signal level to the mixer was reduced to the point where its fourth harmonic being generated in the mixer resulted in no detected (fourth harmonic) current in the bridge circuit.

This "standard" idler frequency was assumed to be a faithful reproduction of the difference frequency between the pump and signal frequencies. This "standard" idler frequency was then used to balance out the "amplified" idler due to the parametric amplifier. As confirmed by the use of a spectrum analyzer, the narrow-band filter allowed only the "amplified" idler to be applied to one port of the bridge circuit. The pump phase was then changed so that (only the pump power) to the amplifier was varied in phase. The bridge output meter was used to record the variation in the idler current output due to the pump phase change.

The idler phase shifter was varied to obtain a minimum bridge reading (4.8×10^{-11} amps), and this setting (40°) was assumed to be the zero reading. After the bridge was balanced, the pump phase was varied and the detected bridge current noted for 30° intervals of pump phase change. Then, with the pump phase returned to its original setting and the bridge still balanced, the idler phase was varied and the resultant detected current measured. The resultant data are plotted in Fig. 3.

From the curves given in Fig. 3 it can be seen that the variation of the pump phase, with the signal phase constant, causes the bridge to be unbalanced which is an indication of a change in phase in the parametric amplifier idler phase angle.

When the bridge was balanced and the idler phase varied, the output current also changed in a manner analogous to the previous experiment. These two curves were not identical 1) because of measurement errors resulting mainly from setting the phase angles and 2) because the two phase shifters were not identical in their characteristics.

Thus within the experimental accuracy of the measurements, one may conclude that the idler phase varies directly with the phase of the pump, and that the expression $\Theta_i=\Theta_p-\Theta_s$ is valid for a three-frequency parametric amplifier [13]. In addition, within the sensitivity of the experiment, the phase of the output signal does not vary with any change in the pump phase.

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Inline Waveguide Attenuator

A new inline attenuator fabricated in the manner of the common multihole directional coupler has qualities for use as an interlaboratory standard. Calibrations of one model of this attenuator and experiments with conventional waveguide components indicate improvements in calibration results with simplified calibration proceeure.

The "inline" waveguide attenuator constructed with two sections of waveguide coupled together in the manner of the common multihole directional coupler possesses very desirable qualities for use as an interlaboratory attenuation standard. This type of construction allows good properties of stability and very low reflection at each port. However, most "inline" waveguide attenuators of this construction do not have the two ports of the attenuator aligned with the same axial reference. Usually one port is displaced in a transverse direction from the reference by a distance equal to one of the transverse dimensions of the waveguide. This displacement of one port of the attenuator makes it more difficult to perform an accurate calibration of the attenuator because the attenuation calibration system must accommodate not only the axial distance represented by the spacing between the two attenuator ports but also a small transverse displacement from the axis. This condition is sometimes difficult to accommodate when one is attempting to perform a very accurate attenuation calibration. The attenuation calibration system usually can provide more accurate measurements with a true inline attenuator.

A configuration for a true inline attenuator is shown in Fig. 1. This configuration is suggested for attenuation values of 3 db or less. The section of waveguide to which a part of the input energy is coupled is made shorter than the main section of waveguide and is terminated at each end with a matched load. Two variations of the configuration of a true inline attenuator having attenuation values greater than 3 db are shown in Fig. 2(a) and (b). In this configura-

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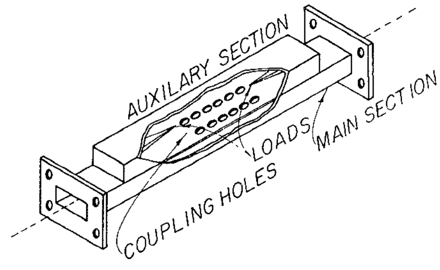

Fig. 1—Inline attenuator for values of attenuation 3 db or less.

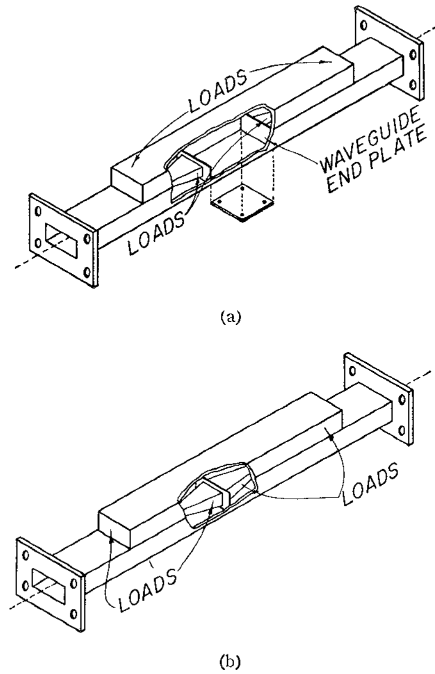

Fig. 2—Inline attenuators with four loads for values of attenuation greater than 3 db.

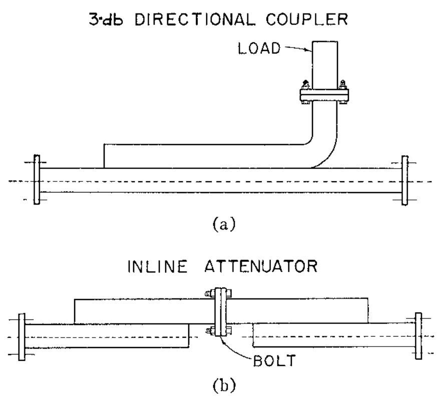

Fig. 3—Inline attenuators made with waveguide components.

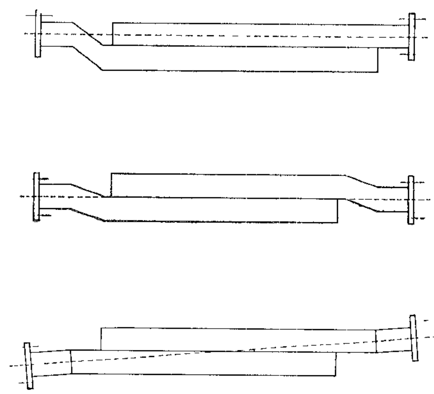

Fig. 4—Inline attenuators with two loads for values of attenuation greater than 3 db.

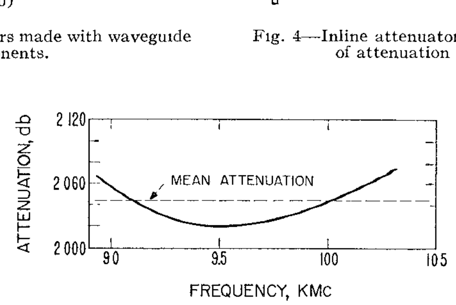

Fig. 5—Attenuation characteristics of Model 1 inline attenuator.