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RADIO AND ELECTRICAL ENGINEERING DIVISION

A SIMPLIFIED RADAR MIXER

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ABSTRACT

A radar mixer for operation in the three-centimeter band is described. A type-2K25 local oscillator mounted in a manner electrically equivalent to the standard JAN mount, provides two well decoupled outputs to feed the AFC and signal crystals. This mount minimizes mode discontinuities and other difficulties peculiar to this type of klystron. The principle is applicable to mixers using any klystrons in which the power output is obtained from an antenna projecting into a wave guide.

CONTENTS

FIGURES

A SIMPLIFIED RADAR MIXER

Introduction

A radar mixer has been designed particularly for use with klystrons of the 2K25 series. The output from these klystrons is obtained from an antenna which projects into a wave guide. Mode discontinuities and other difficulties occur with this tube unless its mount is electrically equivalent to the standard JAN mount for the tube. The mixer to be described has been designed to minimize these difficulties, and simplicity of manufacture has been kept in mind.

Requirements

The following specifications, typical of current radar practice, applied to this mixer:

(a) Two crystals are used — one as a signal mixer, the other as automatic-frequency-control mixer. Both are driven by the same local oscillator.

(b) The mixer is preplumbed; that is, there are no matching adjustments other than the T/R switch tuning adjustment.

(c) A klystron of the 2K25 series is used as a local oscillator. (This group includes the 2K22, 25, 26, 29, 45, 54 and 55, in which the r-f output antenna replaces one of the pins in an otherwise standard octal base.)

The following additional features were required by the particular radar which was available for the final trials of this mixer:

(a) The radio-frequency band of the mixer is 9,345 to 9,405 megacycles per second.

(b) The crystals used are type-1N23B.

(c) The local oscillator is type-2K25.

(d) The intermediate frequency is 30 megacycles per second, with the local oscillator operating at a lower frequency than the magnetron.

(e) Location of the magnetron input and receiver outputs was approximately predetermined by the over-all layout of the radar.

The following characteristics are desirable in a radar mixer of this type:

(a) There should be no mode discontinuities of the local oscillator during the automatic-frequency-control sweep.

(b) The local oscillator should deliver its rated output power throughout the complete operating bandwidth.

(c) The signal and automatic-frequency-control crystal controls should operate smoothly and should not interact appreciably.

(d) The signal energy should be well matched to the average crystal, with a minimum of signal energy wasted in the local oscillator coupling circuit.

(e) The structure should be as simple as possible, for ease of manufacture, with few critical tolerances.

(f) The undesirable coupling of T/R leakage through the local oscillator line to the automatic-frequency-control crystal should be minimized.

Design of the Mixer

Fig.1 is a diagram of a klystron of the 2K25 series. The output is coupled to the wave guide through a long transmission line (A). The electrical length of this line varies among different tubes of the same type (see Ref.(1), p.146).

For reliable operation it is necessary to operate the tube in a mount that is electrically equivalent to the standard mount used for production tests in the tube factory. The essential dimensions of the standard mount are indicated in Fig.1. This mount has only one radio-frequency output, and one of the requirements of this mixer is that two outputs be available. It is also necessary that the coupling between these outputs be as low as possible.

The method devised to obtain two outputs from a mount which is essentially equivalent to the standard mount is indicated in Fig.2. Y is a large susceptance which approximates a short circuit.

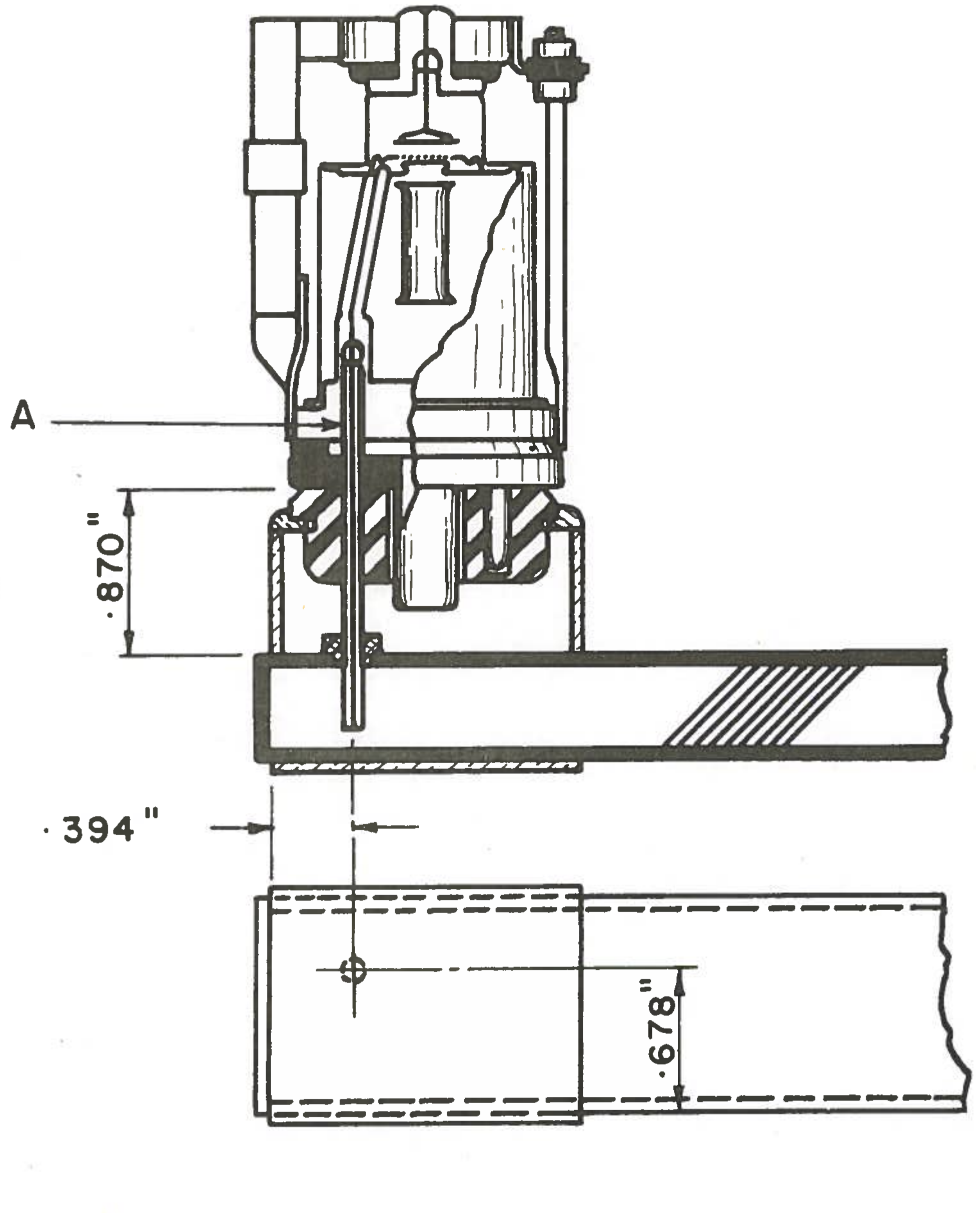

FIG. I

TYPE-2K25 KLYSTRON IN STANDARD MOUNT

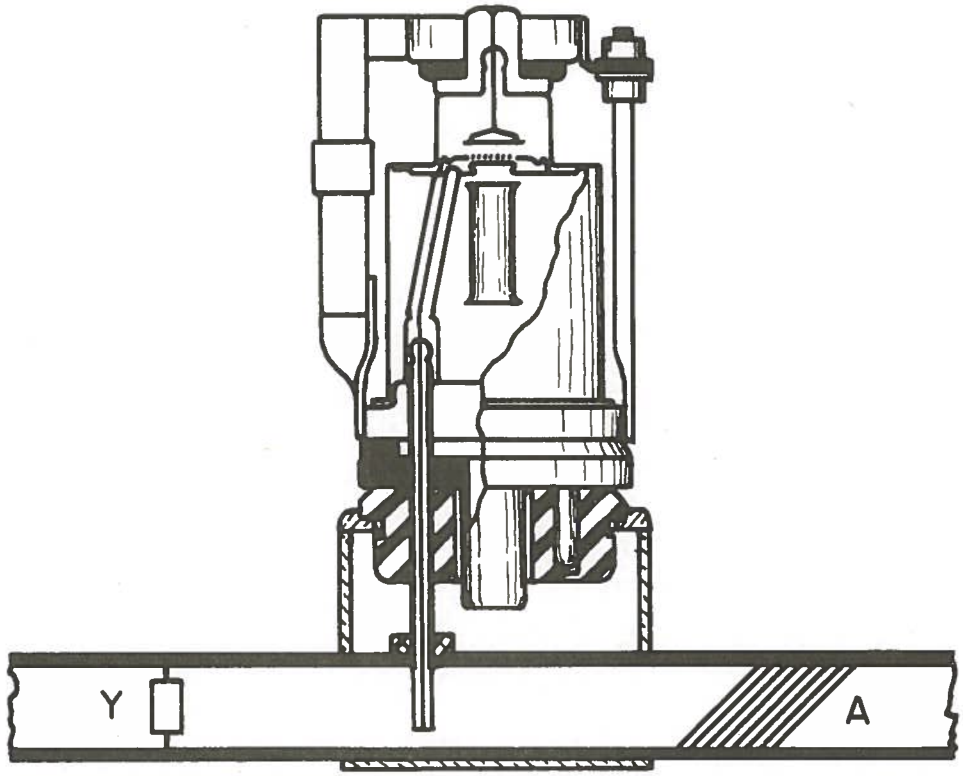

FIG. 2
BASIC MODIFICATION OF STANDARD MOUNT

A is a matched attenuator feeding one output; the other output comprises the power which passes the susceptance, Y.

It is usual to provide for adjustment of the local oscillator drive to each crystal because of large variations in conversion gain and impedance from one crystal to another, on which are superimposed similar large variations in local oscillator power output. The most straightforward method of incorporating these adjustments is to allow the susceptance, Y, to be variable, and to let the attenuator, A, be a conventional flap attenuator. Construction similar to this has, in fact, been used in an application(2) where cost was not an important factor.

A variable attenuator, however, is a relatively expensive device, and in the present design it was found satisfactory to make A a fixed attenuator, followed by a variable susceptance for AFC crystal-current control. This method of crystal-current control is discussed in Appendix I.

An inductive post is used for the variable susceptance. Its construction is shown in Fig.3 and is similar to that suggested by Mr. H.C. Aubrey for use on another project (see Ref.(2), p.2). A is a beryllium-copper sleeve, having slots, B, and arranged to make positive contact at C around the probe D. The expense and difficulty of a choke joint may be avoided with this construction. The beryllium-copper sleeve serves another function in that, if the mixer should ever be operated accidentally with the signal-crystal current-adjusting screw completely removed, this sleeve will provide an adequate susceptance to prevent crystal burn-out from excessive local oscillator drive.

The length of the post was chosen so that its susceptance is inductive. This allows greater length than the equivalent capacitive post, and has another minor advantage in that clockwise rotation of the right-hand threaded screw increases crystal current, which is the normal sense for a control.

The signal and AFC crystals are mounted in the same wave guide which carries the local oscillator. The mixer is completed by the addition of a type-1B24 T/R switch in front of the signal crystal, and an attenuator ahead of the AFC crystal, in the conventional manner.

The complete basic mixer is shown in Fig.4. All components are mounted on a single section of 1 by $\frac{1}{2}$ by .050 inch wall, brass wave guide. In order that the mixer may have a convenient physical shape it will be desirable in many cases to add two right-angle bends,

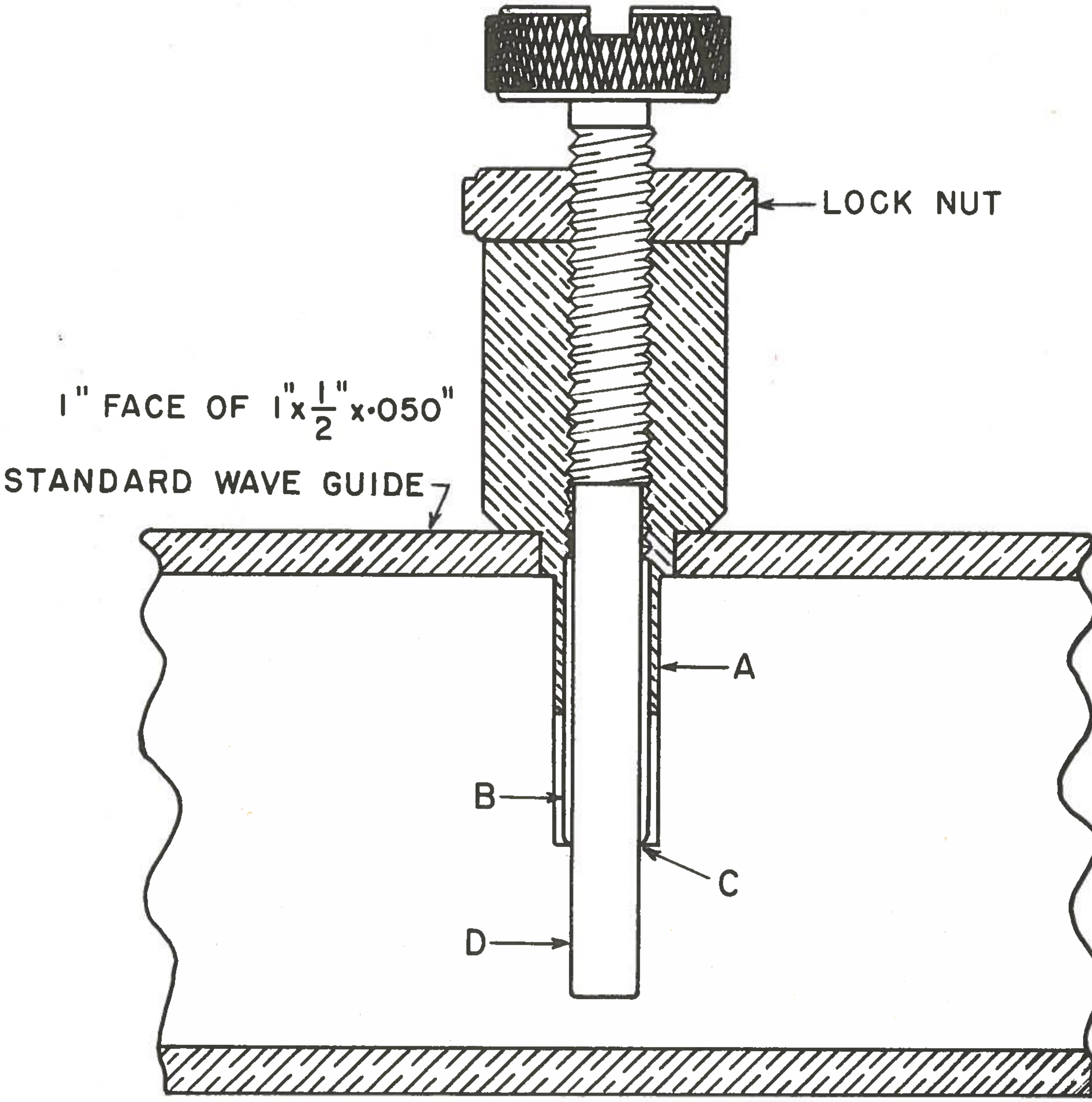

FIG. 3
CRYSTAL-CURRENT ADJUSTING POST

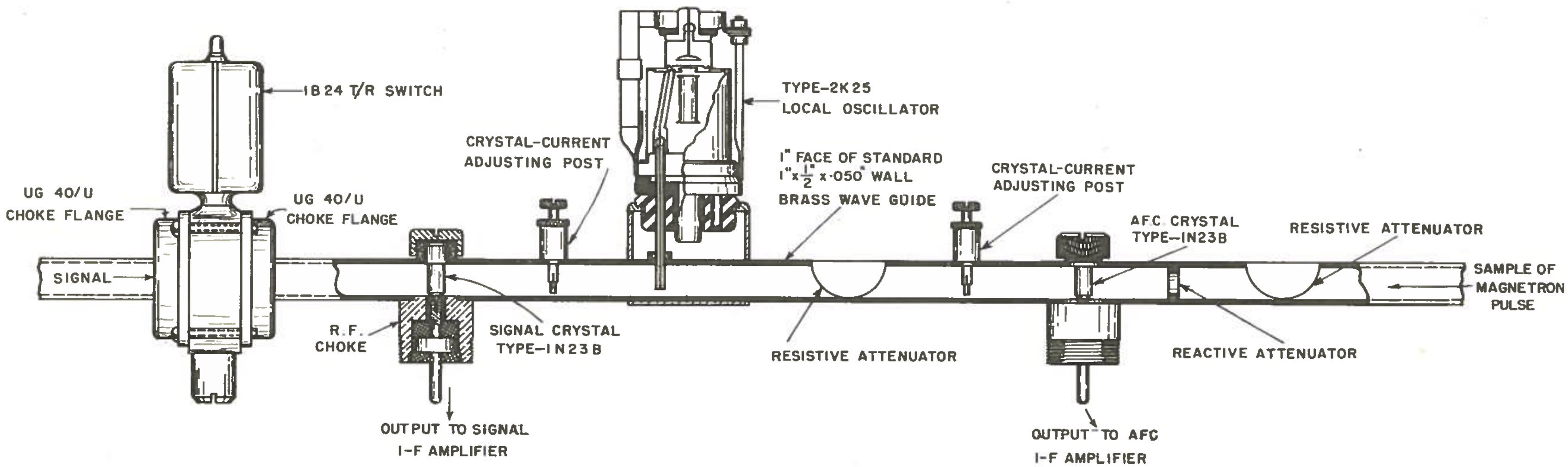

FIG. 4
COMPLETE BASIC MIXER

either in the E-plane or H-plane, thus producing a U-shape for the complete device. Matched corners could, of course, be used for these wave-guide bends. However, production costs can be reduced by using an unmatched corner, and mounting the inductive post at the corner. Fig.5 shows the resulting configuration when the corner is made in the H-plane. The combination of corner and post may be considered as a unit for calculations, by determining its equivalent circuit. This is discussed in Appendix II.

Fig.6 shows the complete mixer in diagrammatic form. In order to establish dimension "a"(Fig.6) for the local oscillator mount, it was necessary to establish the phase of the reflected voltage when local oscillator energy is fed in at A (Fig.5). The impedance at B due to the signal crystal and other components was not yet established. However, this impedance will certainly not have a reflection coefficient greater than 0.4, and in this case the reflected phase at A will be almost independent of the impedance at B. The impedance at B was approximated by a load having a reflection coefficient of 0.4, adjustable through all possible phases, and a signal at 9,345 megacycles per second fed in at A. The dimension b (Fig.5) defining the phase at A was found to vary slightly about the value .094". Thus dimension "a"(Fig.6) is established as

$$\frac{n\,\lambda_g}{2} + .394'' + .094'' .$$

The 0.394" dimension is taken from the standard mount (Fig.1); n is an integer which is arbitrary, except that it cannot be too low because of the interference of higher-order modes of the output antenna and the corner. A value of n higher than necessary will needlessly restrict the bandwidth of the mixer. The minimum safe value of n may be established by constructing a mount with n = 2, and reducing n until a difference in performance is observed. In this manner a value of 1 was chosen for n.

The equivalent short-circuit position, when looking in at B (Fig.5), at a frequency of 9,375 megacycles per second, was found to be

$$\frac{n\,\lambda_g}{2} + .043''$$

in front of the center line of the post. This measurement was made with a klystron located correctly and a matched load behind the

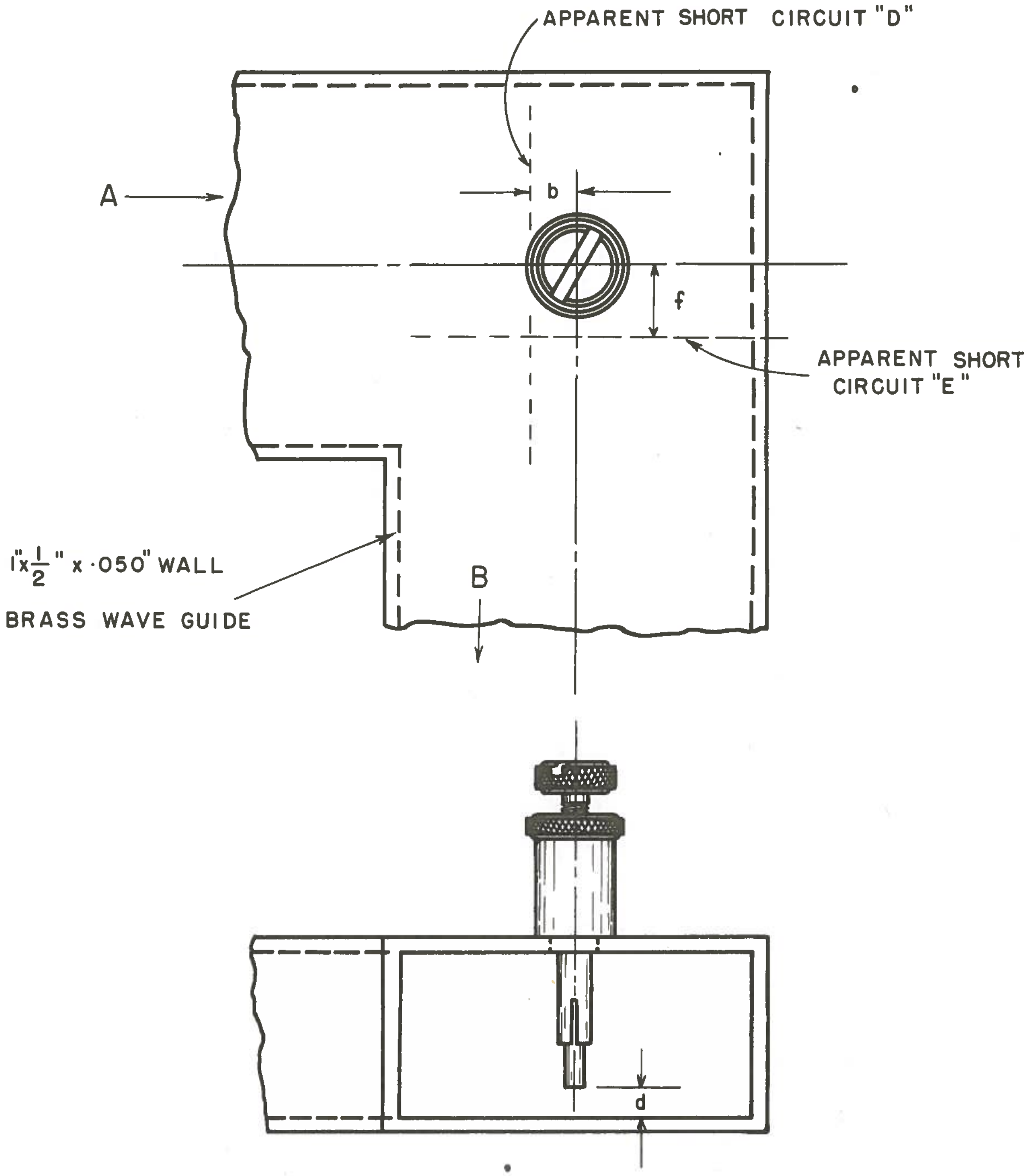

FIG. 5
CRYSTAL CURRENT ADJUSTING POST
MOUNTED IN A WAVE GUIDE CORNER

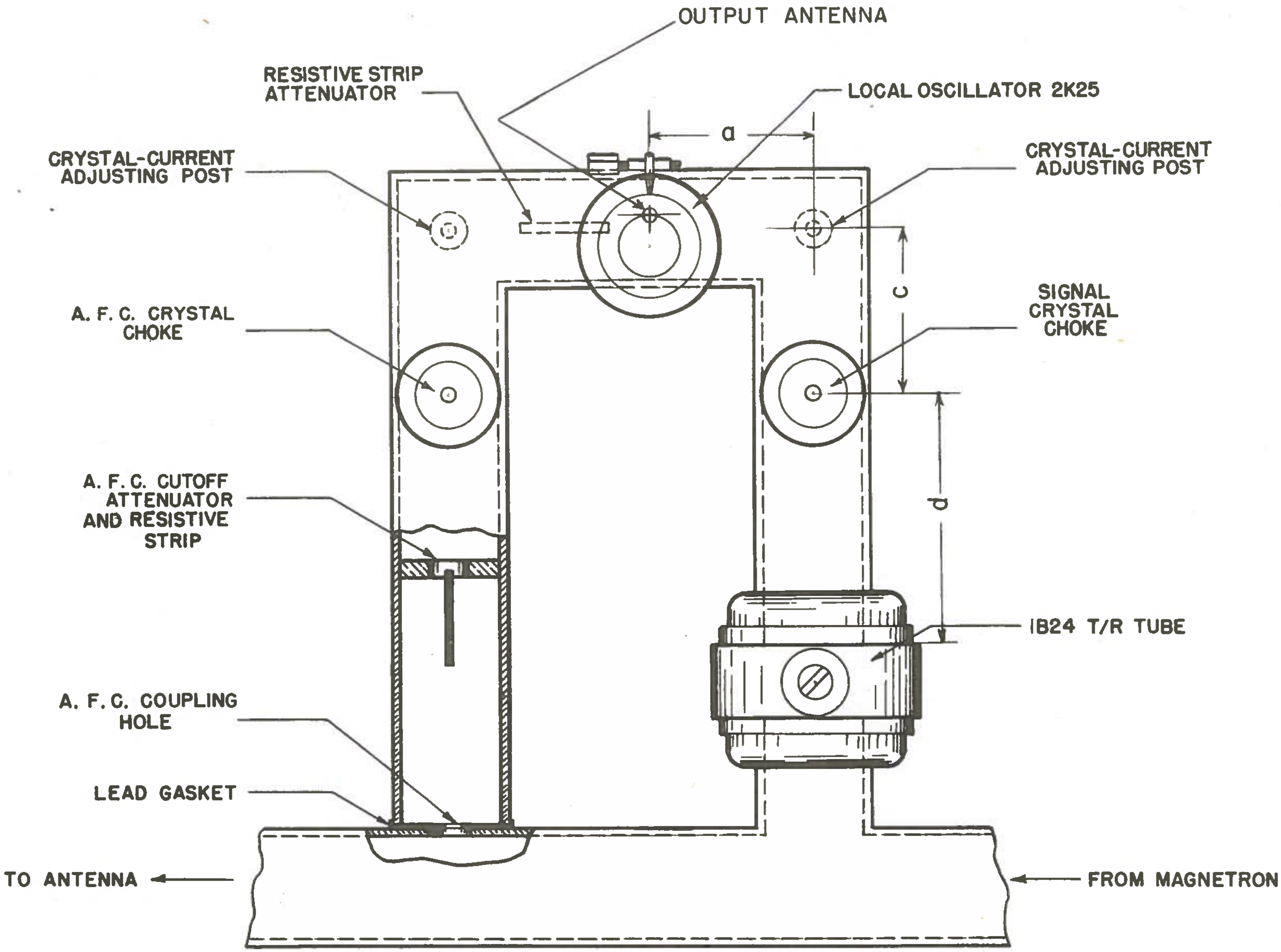

FIG. 6
PLAN VIEW OF COMPLETE MIXER

klystron. Dimension c, Fig.6, is determined by the requirement that the apparent short circuit be the optimum distance behind the signal crystal for best match of signal energy to the crystal. Unfortunately, this distance varies considerably between crystals.

Fig.7 indicates the power delivered to a number of typical 1N23B crystals from a constant matched source at 9,375 megacycles per second, as a function of the distance from the center line of the crystal to the face of a short-circuiting plunger. It can be seen that a distance of about 0.215 λ_g is optimum*. It can also be seen that use of this value will permit as much as 1.7 decibels loss for extreme crystals, because of the preplumbing. There will be a further loss of from 0 to $\frac{1}{2}$ decibel due to the fact that the peak of each curve does not represent a perfect match. The crystals used for these measurements were manufactured a number of years ago. It may be that currently manufactured crystals have a more closely controlled radio-frequency impedance. This matter is under investigation. In practice the tuning of the T/R cell will reduce this loss to some extent.

The crystal mount and associated radio-frequency choke may be seen in Fig.4. The radio-frequency choke consists of alternate quarter-wavelength sections of low and high impedance coaxial line, resulting in a very low series impedance at the tip of the crystal. The performance is similar to that obtained with a standard choke joint. The output capacity, including the crystal, is 9 micromicrofarads.

Location of T/R Switch

The distance d (Fig.6) from the T/R switch to the crystal, is governed by three mutually contradictory criteria, as follows:

(1) d should be chosen for best coupling of local oscillator energy.

(2) d should be chosen so that the match of the greatest number of crystals is improved by the T/R tuning.

(3) d should be chosen for self-protection against burn-out (see Ref.(1), page 173).

Because of the nature of the admittance spread of the crystals used, the second criterion was unimportant. Experience had shown no difficulty with crystal burn-out for the radar used. Thus the third requirement was ignored.

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*This value was established from measurements taken on a larger group of crystals.

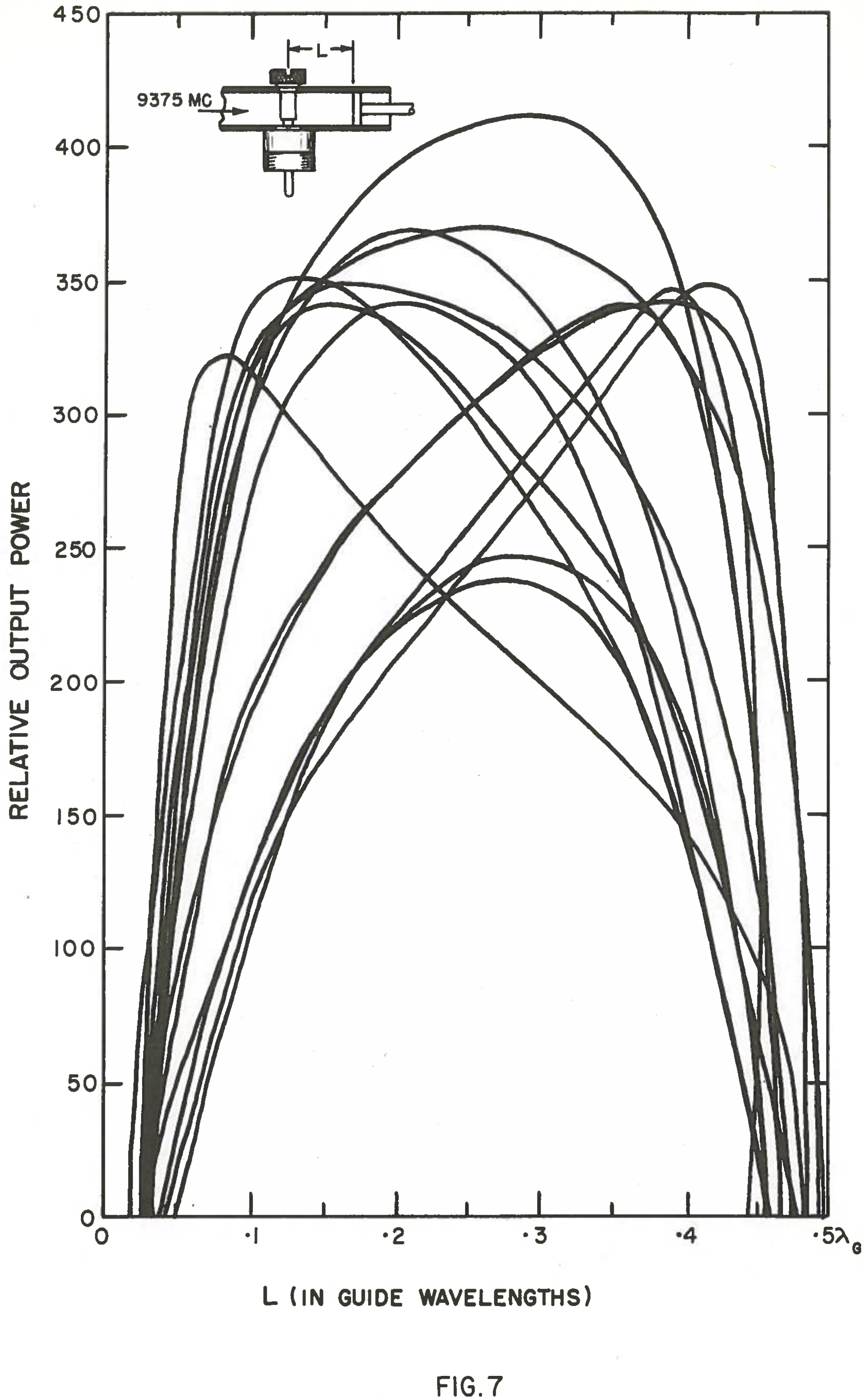

FIG. 7
VARIATIONS IN CRYSTAL MATCH

The crystal mount was added to the mixer and a movable plunger mounted in place of the T/R switch. The best position of the plunger for coupling of local oscillator energy to the crystal was established at $\frac{n\lambda_g}{2}$ + 0.185 inches behind the crystal. This condition will allow the crystal adjusting post to be set at the largest possible susceptance; thus it will be most effective as a short circuit in its function of reflecting signal energy back to the crystal.

The type-1B24 T/R switch has a loaded Q of about 300. Thus, when tuned to 9,375 megacycles per second, it will present a large mismatch at 9,345 megacycles per second. This permits locating the T/R switch in such a position that it approximates the plunger described in the previous paragraph as far as local oscillator energy is concerned. Experiments on the T/R switch showed that the apparent reflection occurred 0.100 inches inside the face of the switch. This establishes dimension d (Fig.6).

Design procedure for the AFC side of the mixer is similar to that for the signal side. In the AFC case, the cutoff attenuator replaces the T/R switch as an efficient reflector for the local oscillator energy.

The magnetron used was type-2J42, operated at a peak power level of six kilowatts. The desired level of magnetron sample reaching the AFC crystal is about 1.5 milliwatts. Thus an attenuation of 66 decibels must be provided by the AFC coupling hole, the cutoff attenuator, and the resistive strip attenuator. This attenuation can be obtained easily, and the substitution method can be used to check the attenuation of the final design.

The mixer was designed to fit the existing tee-section in the radar. This section can be seen in Fig.6 (not to scale).

It was determined experimentally that d (Fig.5) for the signal crystal-current control would range between 0.110" and 0.150" in operation, due to differences between crystals and other variables. It was necessary to determine the extent to which the voltage minima D and E (Fig.5) deviated from their design position, as the crystal-current post was varied over this range. Deviations of D will interfere with the "standard" nature of the local oscillator mount, while variations in E will mismatch the signal crystal. In Appendix III it is shown that the shift in apparent minimum position as the crystal-current screw is varied over this range is about ±3° (λ_g = 360°). This was confirmed by experiment, and is a negligible variation.

The attenuator which provides the matched load for the klystron may be built in any convenient manner. A value of about 7 decibels was found suitable. For simplicity of construction, a disk of 1/16" Synthane*, Grade L-564, is inserted through a slot in the wide face of the wave guide.

Test Data and Conclusions

Decoupling of the two crystals is a problem which must be considered in a mixer which uses two crystals fed from the same local oscillator. A maximum of decoupling is desired, in order to prevent spurious signals from the T/R spike, or from adjacent radars, from reaching the AFC crystal at levels comparable with the normal magnetron sample.

The decoupling in this mixer was measured and found to be about 35 decibels**, or greater, with some variation due to lack of uniformity of crystal impedance. This amount of decoupling is adequate.

Interdependence of the signal and AFC crystal-current controls is a property of some mixers which, while not serious, is a nuisance when adjusting the radar. To investigate this, the mixer was installed in a complete operating radar and the setting of the crystal-current adjusting screws was varied as indicated in Figs. 8 and 9. It may be seen from these diagrams that the interdependence is negligible. It should be appreciated that the automatic-frequency-control was in operation during this measurement, and thus some variation in local oscillator power output could be expected during the test of Fig. 9, since there was considerable variation in the strength of the signal reaching the discriminator because of variation in conversion gain of the AFC crystal with crystal current.

A good over-all test of the AFC operation is to disturb the system deliberately by tuning the klystron mechanically, and observing the range over which the reflector voltage will vary automatically before the automatic-frequency-control drops out. This range was found to be about 25 volts on a typical klystron, which is satisfactory.

Brazing of the wave-guide corners and subsequent cleaning of the inner surfaces would normally be a difficult manufacturing operation. However, fillets and projections of solder up to 1/8 inch

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*Manufactured by the Synthane Corporation, Oaks, Penn., U.S.A.

**A signal through the T/R cell driving the signal crystal at a level P_o will drive the AFC crystal at a level 35 db less than P_o.

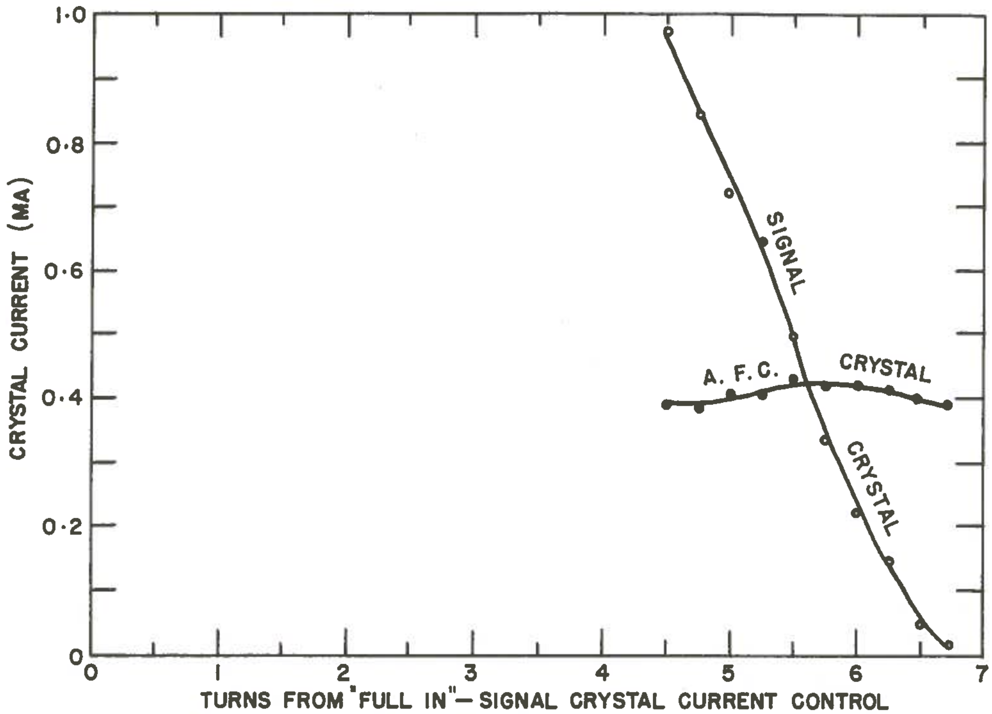

FIG. 8

OPERATION OF SIGNAL CRYSTAL-CURRENT CONTROL

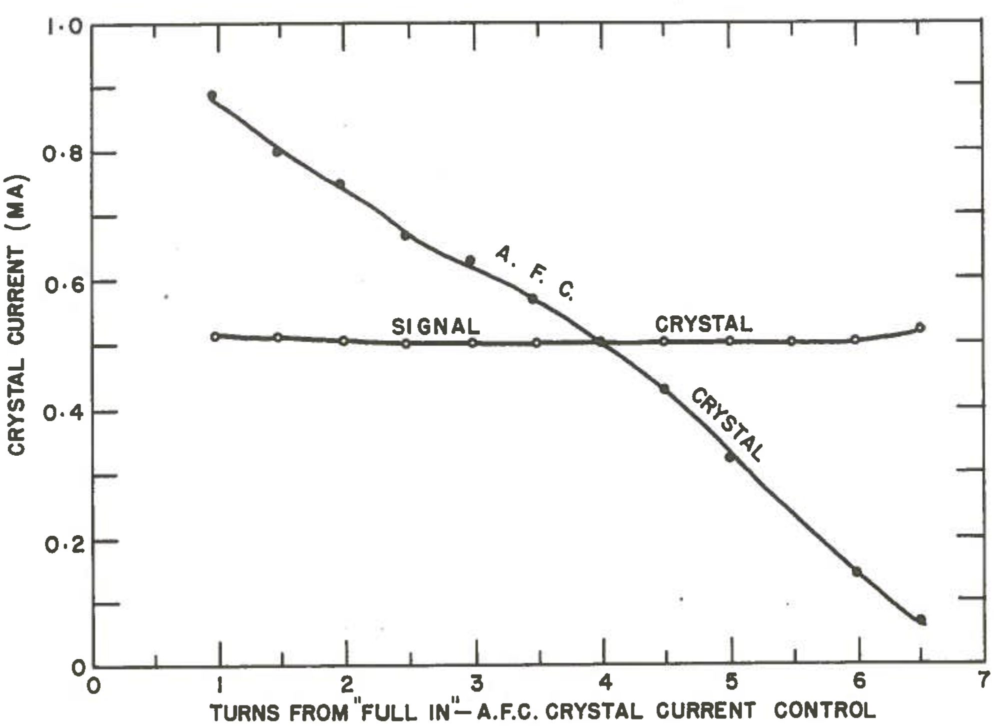

FIG. 9

OPERATION OF A-F-C CRYSTAL-CURRENT CONTROL were simulated at these corners, and produced no effect on performance. Hence it will not be necessary to clean the inside of these joints. The first order effect of projections of solder is merely to add susceptance in parallel with the existing susceptance. This is, of course, not objectionable.

MIT[3][4] has reported that noise from a type-2K25 klystron is very dependent on the impedance presented by the tube mount. If it is desired to make use of the MIT data or to investigate it further as applied to a particular radar receiver, the impedance presented to the klystron may be adjusted to any desired value by means of a post in the wave guide between the klystron antenna and the flat load.

Photographs of the completed experimental mixer appear in Figs. 10 and 11.

This mixer has been installed in an operating radar in order that field experience may be acquired (see Fig.12). It is felt that present tests are adequate to prove the feasibility of the mixer; however, only a few experimental models have been built, and some redesign would be desirable for a production unit.

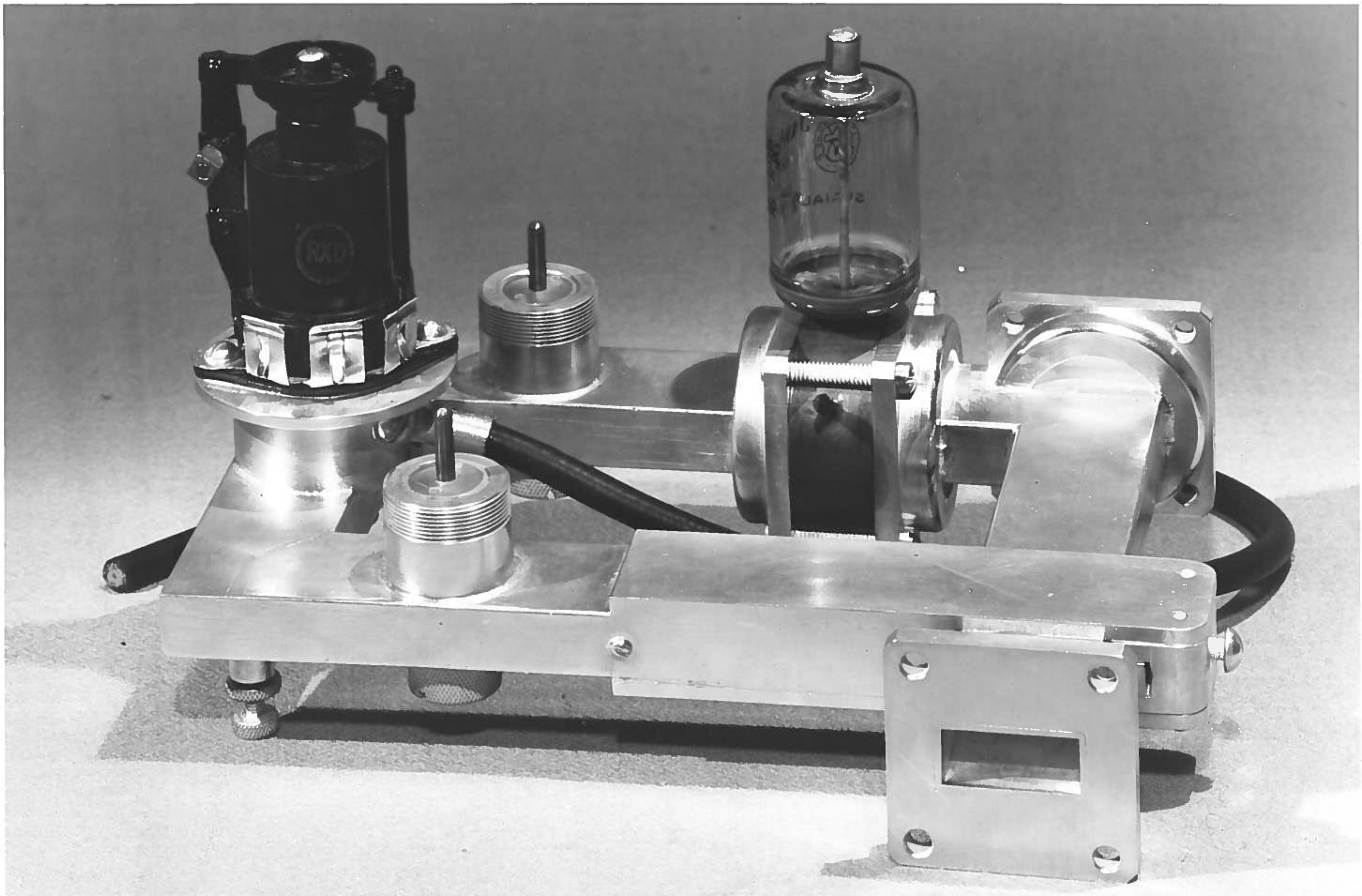

FIG. 10
SINGLE WAVE GUIDE MIXER

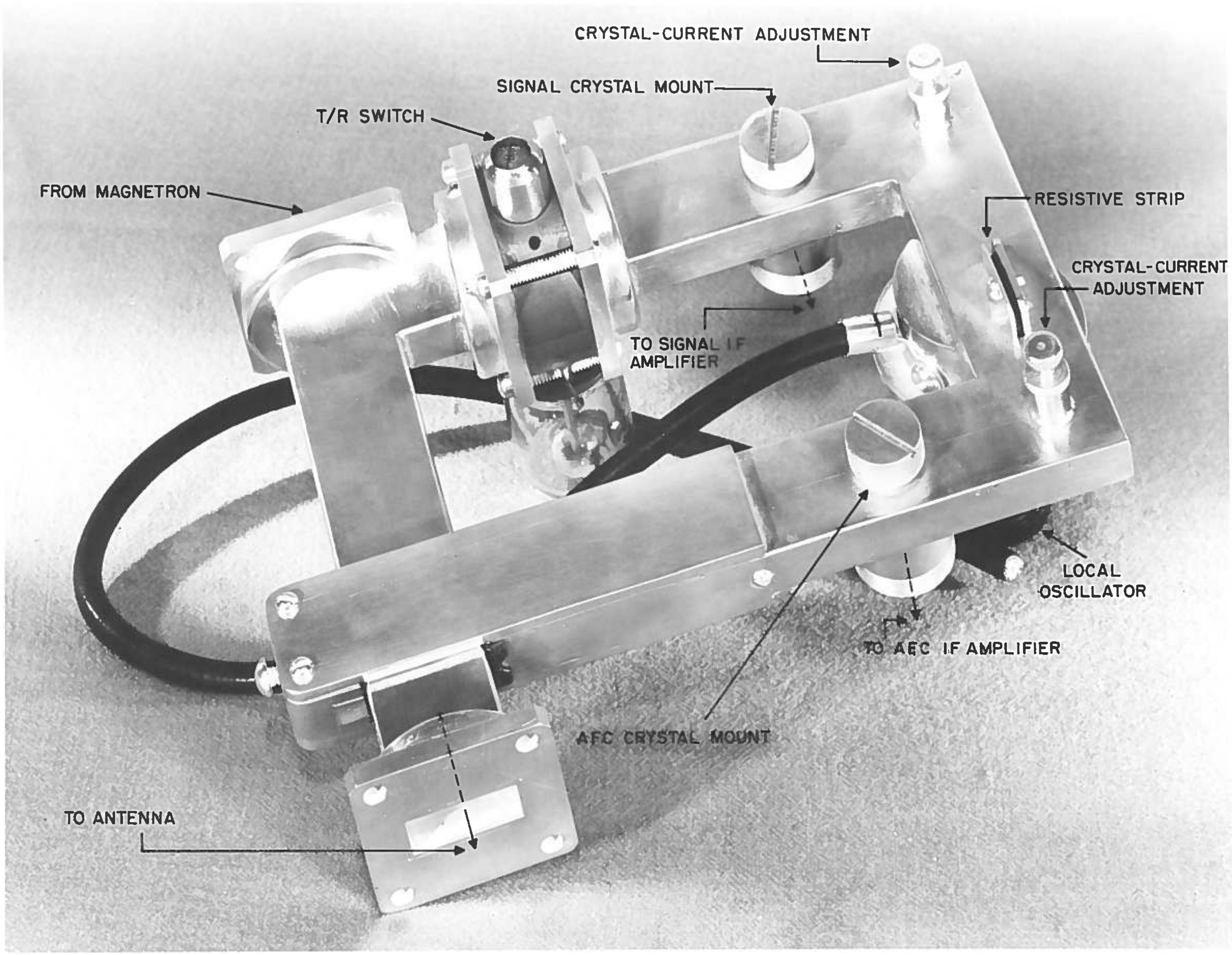

FIG. 11
SINGLE WAVE GUIDE MIXER
(UNDERSIDE VIEW)

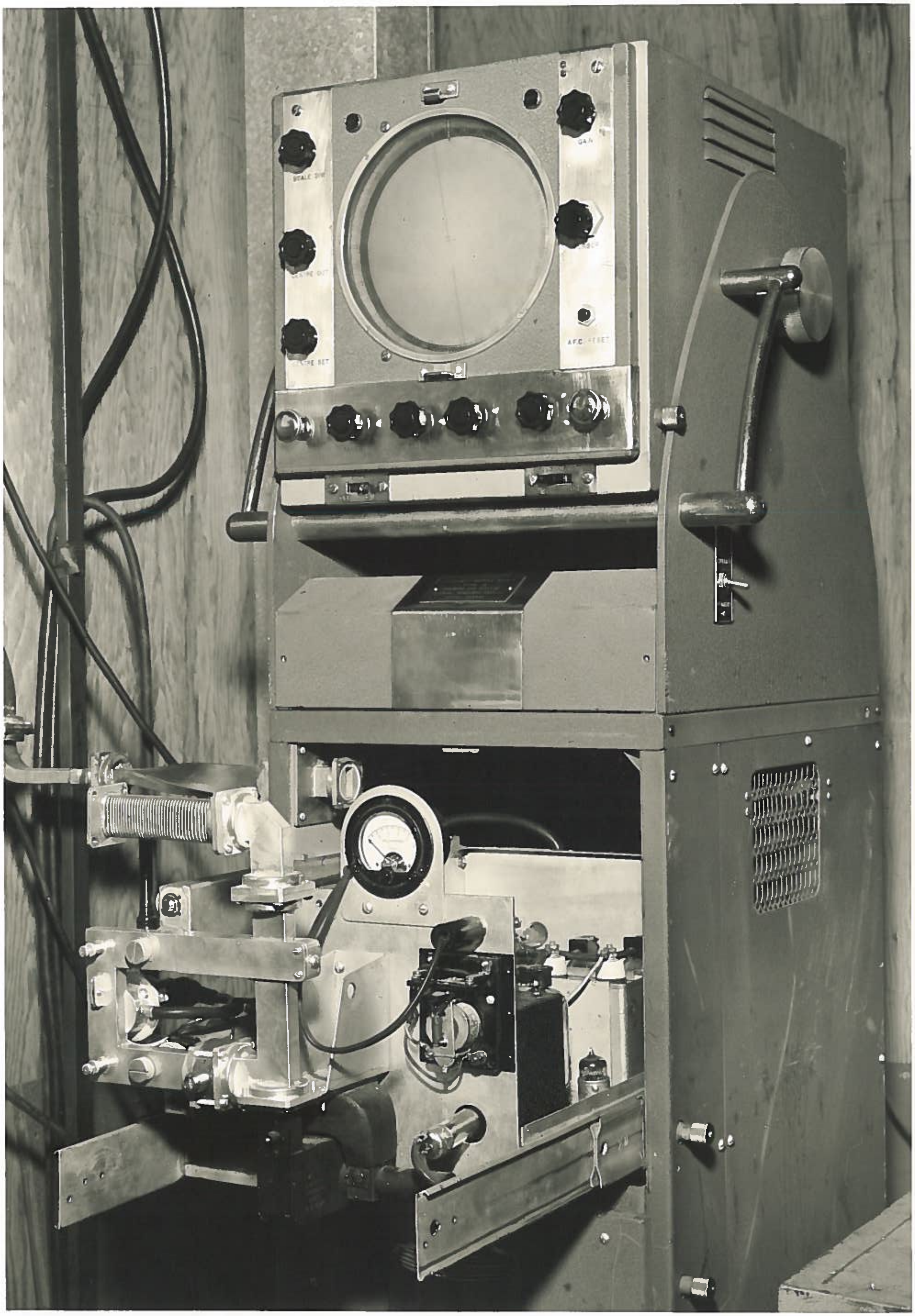

FIG. 12

MIXER MOUNTED IN "MERCHANT MARINE RADAR"

APPENDIX I

DEPENDENCE OF IMPEDANCE MATCH OF CRYSTAL ON CRYSTAL-CURRENT ADJUSTMENT

The method used for crystal-current control will have little effect on the AFC drive. This is illustrated in the admittance chart, Fig. 13. Let AB be the locus of admittance at M as the depth of the crystal current post is varied; M is the equivalent plane at which the corner and post appear as a pure shunt susceptance.

Referred to plane M' before the crystal is added, this locus becomes A'B'. Adding the crystal conductance, the locus will assume some such position as A"B", or A'''B'''.

It can now be seen that radio frequency entering at N will see an impedance which varies little with the position of the crystal-current adjusting post, and hence the magnetron power reaching the crystal will not vary appreciably.

Similar reasoning applies to the signal side of the mixer.

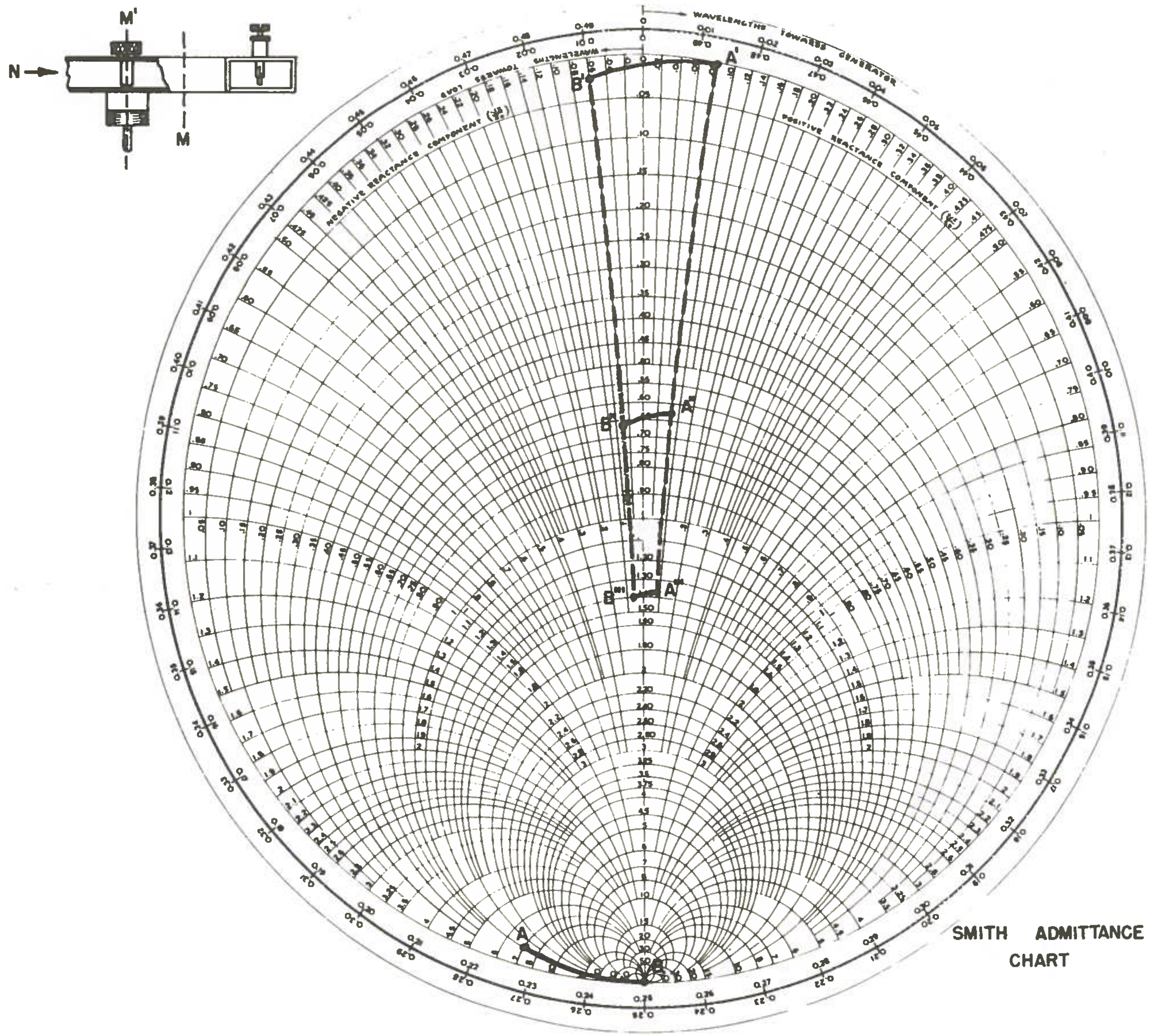

FIG. 13

VARIATION IN MATCH CAUSED BY ADJUSTMENT TO CRYSTAL-CURRENT

APPENDIX II

DETERMINATION OF EQUIVALENT CIRCUIT OF POST AND CORNER COMBINATION

An adjustable post mounted in a wave guide corner, as in Fig. 5, was constructed, with the dimension "a", Fig. 14, from the center line of the post to the face of each choke flange arbitrarily set at 1.760 inches. The distance b from the standing wave minimum to the reference plane A was measured in guide wavelengths with a standing wave probe. Two readings were taken, with a short circuit and an open circuit at output B. The short circuit was a brass plate across the choke flange; the open circuit, a quarter-wavelength of short-circuited wave guide. The equivalent circuit of the device with reference to the two arbitrarily chosen planes was determined from the usual network formulae:

$$X_s = X_{oc} \mp \sqrt{X_{oc}\,(X_{oc} - X_{sc})}, \text{ and} \qquad (1)$$

$$X_p = \pm\sqrt{X_{oc}\,(X_{oc} - X_{sc})}, \qquad (2)$$

where X_{oc} and X_{sc} are open- and short-circuited input reactances.

This has been done for various positions of the adjustable screw and the results are plotted in Fig. 15. The sign of the root has been chosen arbitrarily since limited equivalence is adequate for the equivalent circuit.

The reference planes are arbitrary, therefore they may be located to simplify the equivalent circuit. Reference 5, and page 109 of Reference 6, give a method for establishing these simple reference planes. The method described below which starts only from a measurement of X_{oc} and X_{sc}, may also be used.

The corner and post assembly as constructed are represented by the equivalent T shown in Fig. 16. Fig. 17 represents the same corner with the addition of two equal lengths of wave guide of arbitrary length L. It is desired to find the value of L which will make the entire network appear as a single shunt element. It is noted that the shunt element of the original equivalent circuit has been replaced by two elements in parallel.

A four-terminal network may be characterized by a matrix

$$\begin{Vmatrix} a & b \\ c & d \end{Vmatrix}$$

where a, b, c, d, are defined by the following equations:

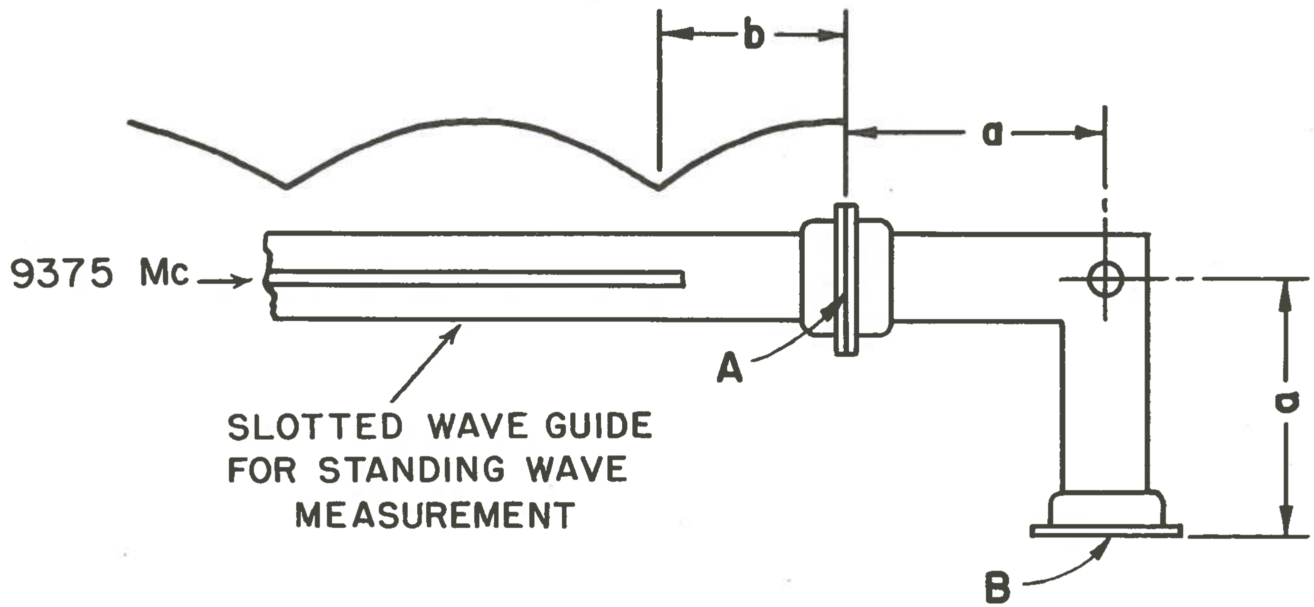

FIG. 14
MEASUREMENT OF REFLECTION FROM CRYSTAL-CURRENT ADJUSTING SCREW

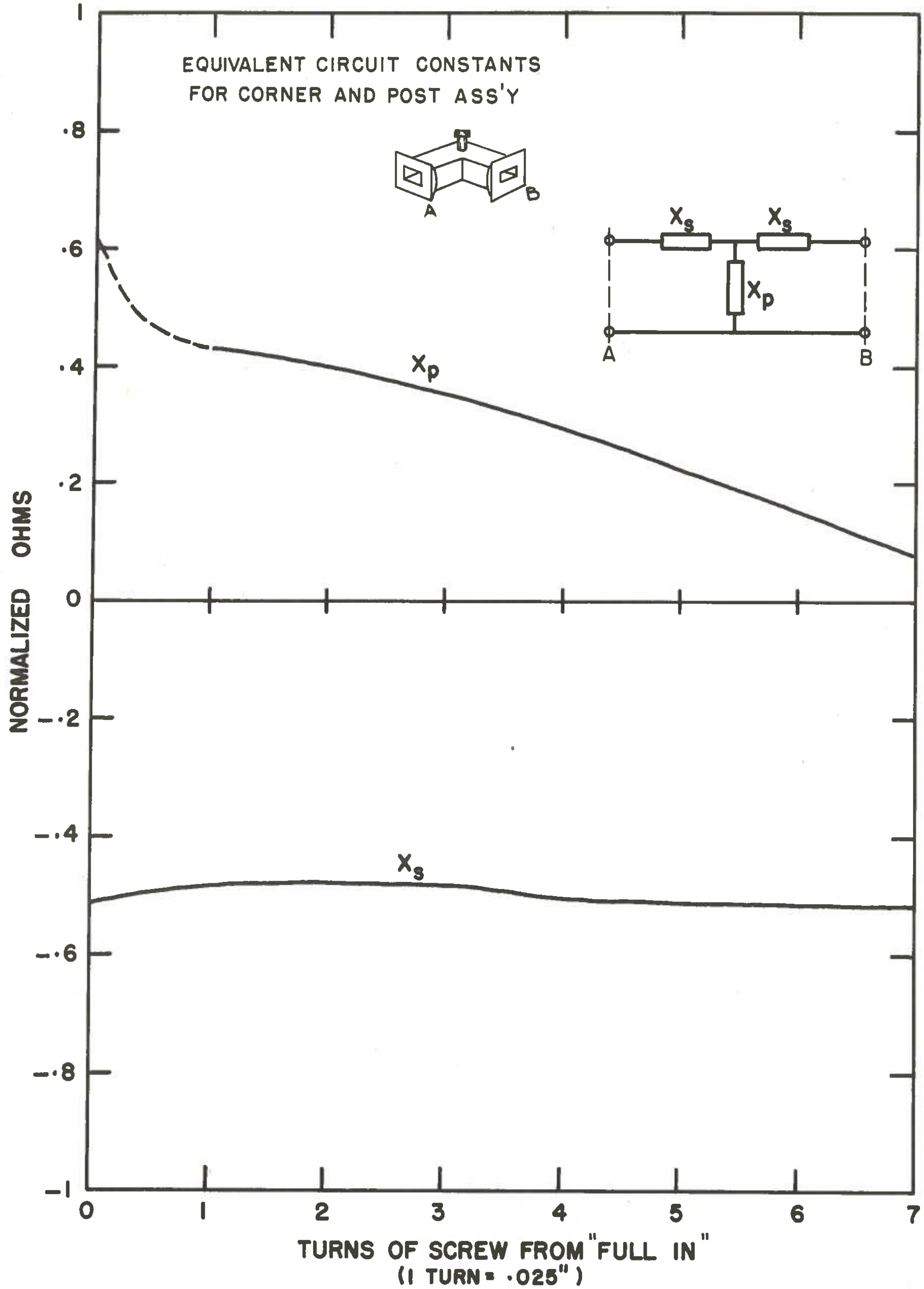

FIG. 15
MEASURED CONSTANTS OF EQUIVALENT CIRCUIT OF CORNER

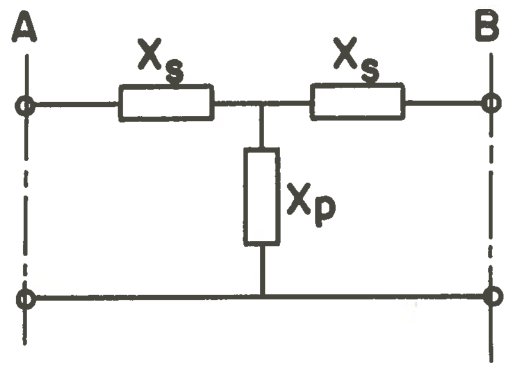

FIG. 16
EQUIVALENT CIRCUIT OF CORNER

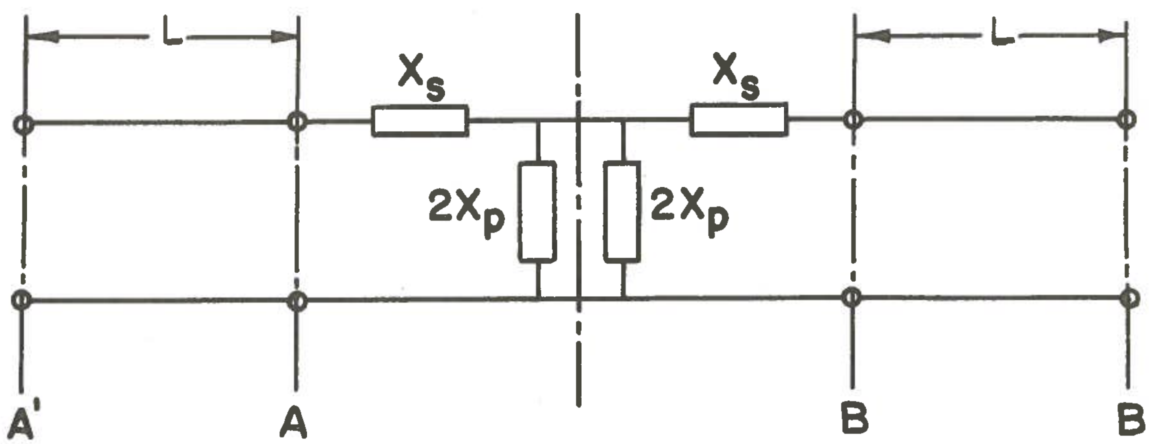

FIG 17
EQUIVALENT CIRCUIT OF CORNER WITH ADDED TRANSMISSION LINE

$$E_1 = aE_o + bI_o, \text{ and}$$

$$I_1 = cE_o + dI_o ,$$

E_1, I_1, E_o, I_o, being the input and output voltages and currents.

The determinant (ad - bc) of this matrix equals unity.

The normalized matrix of a section of lossless transmission line is:

$$\begin{Vmatrix} \cos\theta & j\sin\theta \\ j\sin\theta & \cos\theta \end{Vmatrix} \qquad M_1$$

where $\theta = -j\gamma L$,
γ = propagation constant of the line, and
L = length of the line.

Also the matrix of the appropriate L-section (Fig. 17) is:

$$\begin{Vmatrix} (1 + \frac{X_s}{2X_p}) & jX_s \\ \frac{-j}{2X_p} & 1 \end{Vmatrix} \qquad M_2$$

Thus the matrix of the portion of Fig. 17 to the left of the center line is M_1M_2.

Applying the usual rule for matrix multiplication, the matrix becomes:

$$\begin{Vmatrix} (1 + \frac{X_s}{2X_p}) \cos\theta + \frac{1}{2X_p} \sin\theta & j(X_s \cos\theta + \sin\theta) \\ j(1 + \frac{X_s}{2X_p}) \sin\theta - \frac{1}{2X_p} \cos\theta & -X_s \sin\theta + \cos\theta \end{Vmatrix}$$

$$= \begin{Vmatrix} a & b \\ c & d \end{Vmatrix} \qquad M_3$$

The matrix of this network with the input and output reversed is:

$$\begin{Vmatrix} d & b \\ c & a \end{Vmatrix} \qquad M_4$$

Multiplying these matrices gives the matrix for the entire network of Fig. 17,

$$\begin{Vmatrix} (ad + bc) & (2ab) \\ (2dc) & (ad + bc) \end{Vmatrix} \qquad M_5$$

The matrix of a reactive shunt element is:

$$\begin{Vmatrix} 1 & 0 \\ \frac{-j}{X'_p} & 1 \end{Vmatrix} \qquad M_6$$

If Fig. 17 is to be equivalent to a pure shunt element, then:

$$\begin{aligned} ad + bc &= 1, \\ 2ab &= 0, \\ 2cd &= -\frac{j}{X'_p}, \end{aligned}$$

and also, $ad - bc = 1,$

from which, $ab = bc = 0.$

Thus, either $a = c = 0,$ or $b = 0.$

It is clear that $a = c = 0$ is trivial, thus $b = 0$; hence:

$$X_s \cos\theta + \sin\theta = 0,$$

from which $\tan\theta = -X_s$ (3)

This establishes the length of line which will make the network a shunt element. The admittance of the shunt element is:

$$Y = 2dc$$

$$= 2j\,(-X_s \sin\theta + \cos\theta)\left[\left(1 + \frac{X_s}{2X_p}\right)\sin\theta - \frac{1}{2X_p}\cos\theta\right].$$

Dividing through by $\cos^2\theta$ and replacing $\tan\theta$ by $-X_s$:

$$Y = -2j\cos^2\theta\,(X_s^2 + 1)\,\frac{(2X_sX_p + X_s^2 + 1)}{2X_p}.$$

Replace $\cos^2\theta$ by its value $\dfrac{1}{1 + \tan^2\theta} = \dfrac{1}{(X_s^2 + 1)}$,

$$X_p' = \frac{X_p}{2X_sX_p + X_s^2 + 1} \qquad (4)$$

It may be seen from Fig. 15 that the value of X_s is approximately constant at 0.5 normalized ohms. The choice of reference planes to make the corner appear as a shunt element is then essentially independent of the depth of the adjustable post, since equation (3) states that the location of these planes is a function of X_s only, and not X_p.

Assuming these planes to be established, formula (4) may be used to obtain the data of Fig. 18 which shows the equivalent shunt reactance that appears in parallel with the wave guide as a function of post penetration.

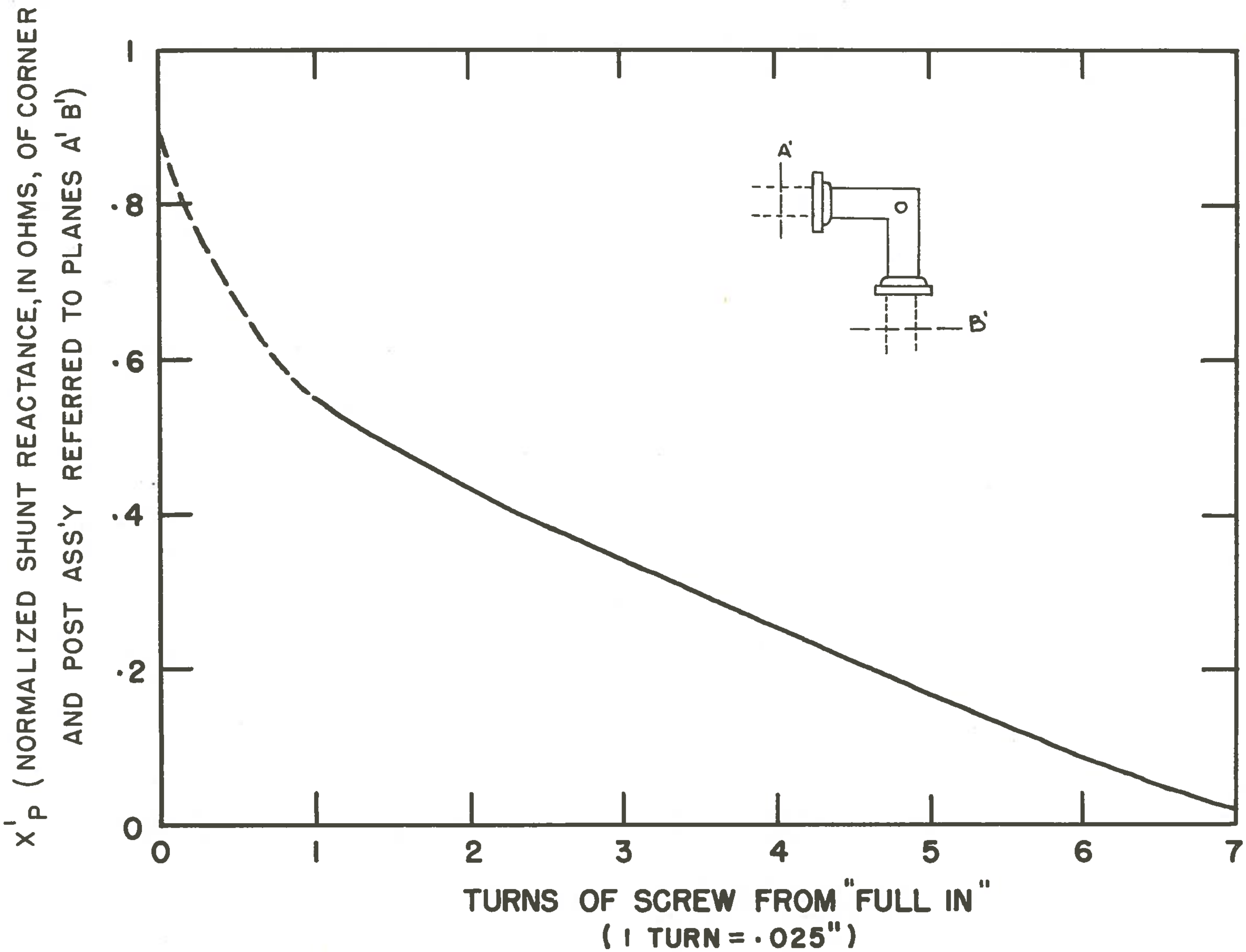

FIG. 18

X'_P VERSUS POST DEPTH

APPENDIX III

VARIATION IN PHASE OF REFLECTION FROM CRYSTAL CURRENT POST WITH POST DEPTH

Experience with the mixer described has shown that the operating position (d---Fig. 5) for the signal crystal-current adjusting post will usually fall within the range 0.110" to 0.150"; that is, with the screw between 4.5 to 6 turns from full in.

Reference to Fig. 18 shows that the equivalent normalized shunt reactance of the post will range from 0.2 to 0.1 ohms. Assuming this to be in parallel with a conductance of unity, the total admittance then ranges from 1-5j to 1-10j ohms.

Application of the formula

$$\rho = \frac{1 - Y_r}{1 + Y_r},$$

where ρ = reflection coefficient,
and Y_r = admittance at the line termination,

indicates that the phase angle ϕ of ρ varies from 158 to 169 degrees.

It is easily shown that the electrical distance, θ, from a termination to a voltage minimum is given by:

$$\theta = 90 + \frac{\phi}{2} .$$

Thus the location of the voltage minimum will vary over a range of less than ±3° as the post depth is altered. This amount of variation does not appreciably effect the crystal match.

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