

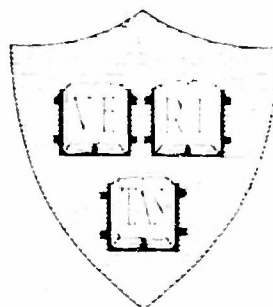
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IMAGE-PLANE AND COAXIAL LINE
MEASURING EQUIPMENT AT 600 MC



By

Howard W. Andrews

July 1, 1953

Technical Report No. 177

Craft Laboratory
Harvard University
Cambridge, Massachusetts

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Abstract

This report considers the construction of the measuring lines and the associated electronic devices for determining the driving point impedance and the current and charge distributions upon antennas. Configurations possessing a plane of symmetry are mounted over an image screen behind which is located the measuring equipment.

The measuring line connects into the image screen and its center conductor may be extended to form an antenna. This conductor is slotted and the currents and charges distributed on it may be determined by probes sliding within it. An unmodulated source of radio frequency energy is used and the modulation introduced in the superheterodyne receiver used to measure the signal level. The frequency of the source is monitored by comparison, in a spectrum analyser, with a crystal-controlled source of energy.

1. Introduction

Described in this report in considerable detail is an experimental setup for investigating the properties of antennas in the region of 600 MC. Physically it is capable of but one thing: presenting a meter reading proportional to the current in the center conductor of a specially constructed coaxial line. These readings, when properly interpreted, give two of three important properties of the antenna configuration, namely its current distribution and driving point impedance over a finite image screen.

The equipment is capable of investigating the above properties only of antenna configurations which possess a plane of symmetry. Such arrays may be mounted over an image screen and their characteristics measured by

personnel and equipment operating behind the shielding screen and out of the immediate field of the antenna structure. Searching devices may be projected through the antenna and possibly through the image screen to measure the desired properties. The antenna is driven by sources at the plane of the image screen or by generators on the antenna surface but excited by transmission lines leading back through the antenna to the generating equipment.

The physical configurations used represent as closely as possible the mathematical antenna model. The theoretical slice generator is replaced by a thick annular generator at the base of the antenna. This representation of the ideal generator is not perfect and fails principally in the immediate vicinity of the base of the antenna over a region approximately as large as the annulus. The theoretically infinite image screen is replaced by a finite one that is large compared to the wavelength and to the antenna structure being investigated. Finally, the ideal measuring line with its continuous parameters is replaced by one which is terminated by the discontinuous surface of the image screen. All these physical approximations are unavoidable in devising an experimental representation of the mathematically tractable theoretical model. The exact evaluation of the errors so arising is not simple. The problem of the finite ground screen and the effect of the discontinuous line have been considered. Through these techniques it is at least possible to estimate the resulting error and, to some extent, correct for its presence.

The general equipment arrangement is a vertically mounted, outdoor, image screen into which is connected a horizontal coaxial line. The line is excited by a transmitter and the detecting equipment is a superheterodyne receiver. The various components are described in the following pages. Immediately following is a block diagram of the experimental arrangement.

2 The Image Screen

The image screen is as large as is physically convenient considering that the antennas are to be mounted approximately in its center and that the antenna position is to be opposite a third floor window. The screen extends up to the roof of the building and below about the same distance.

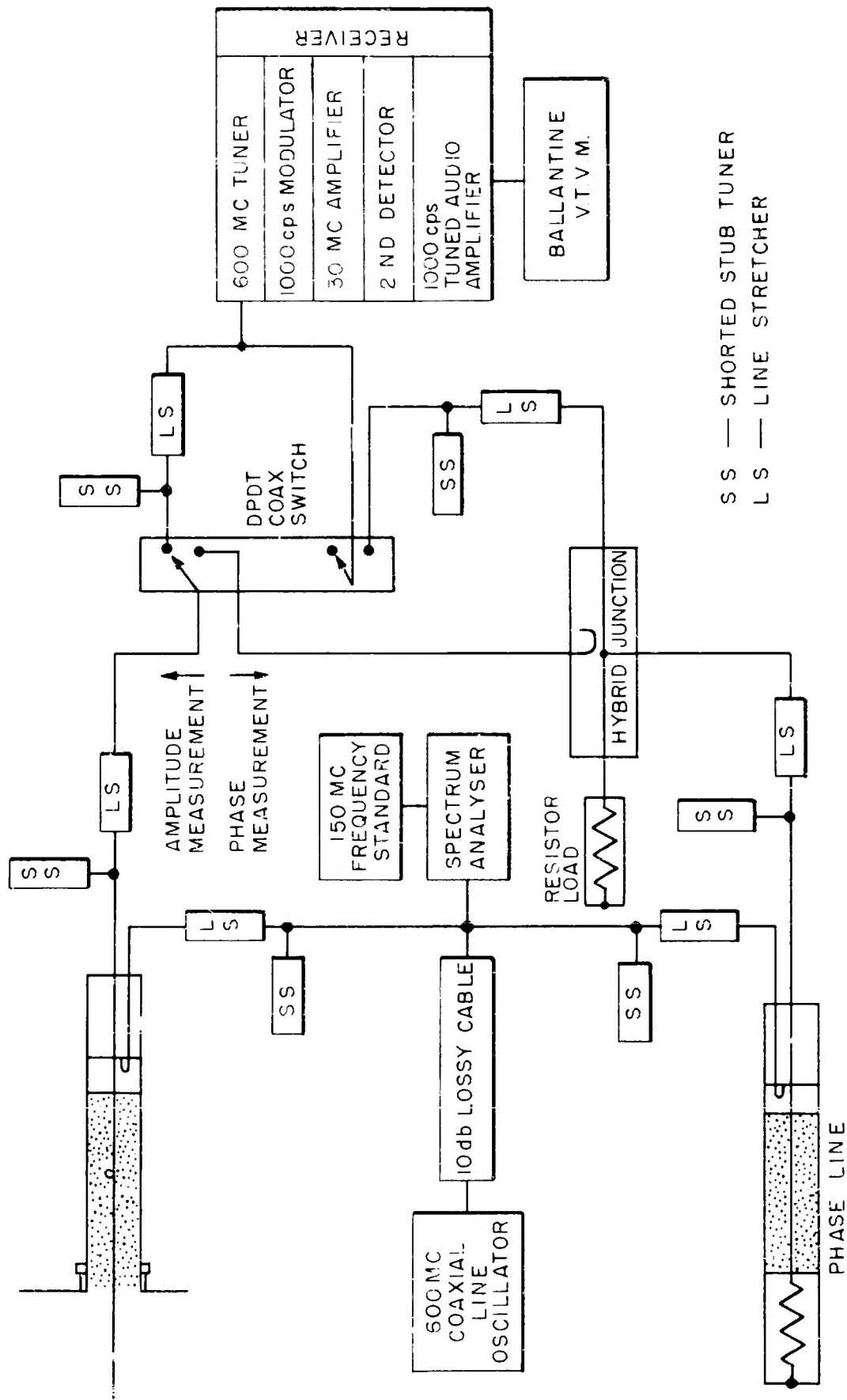


FIG. 1 BLOCK DIAGRAM OF EXPERIMENTAL EQUIPMENT

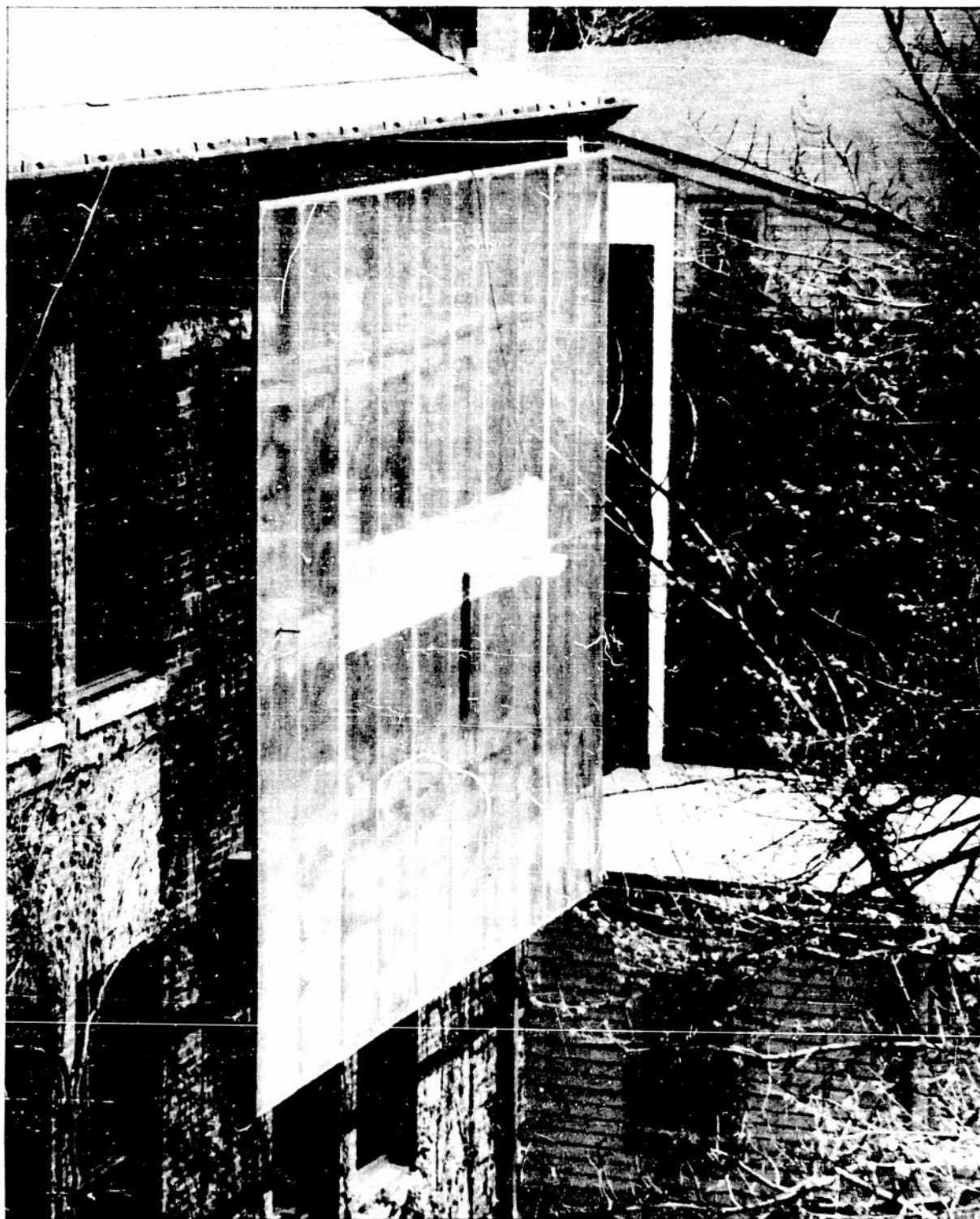


FIG. 2 IMAGE SCREEN AND ANTENNA MOUNT

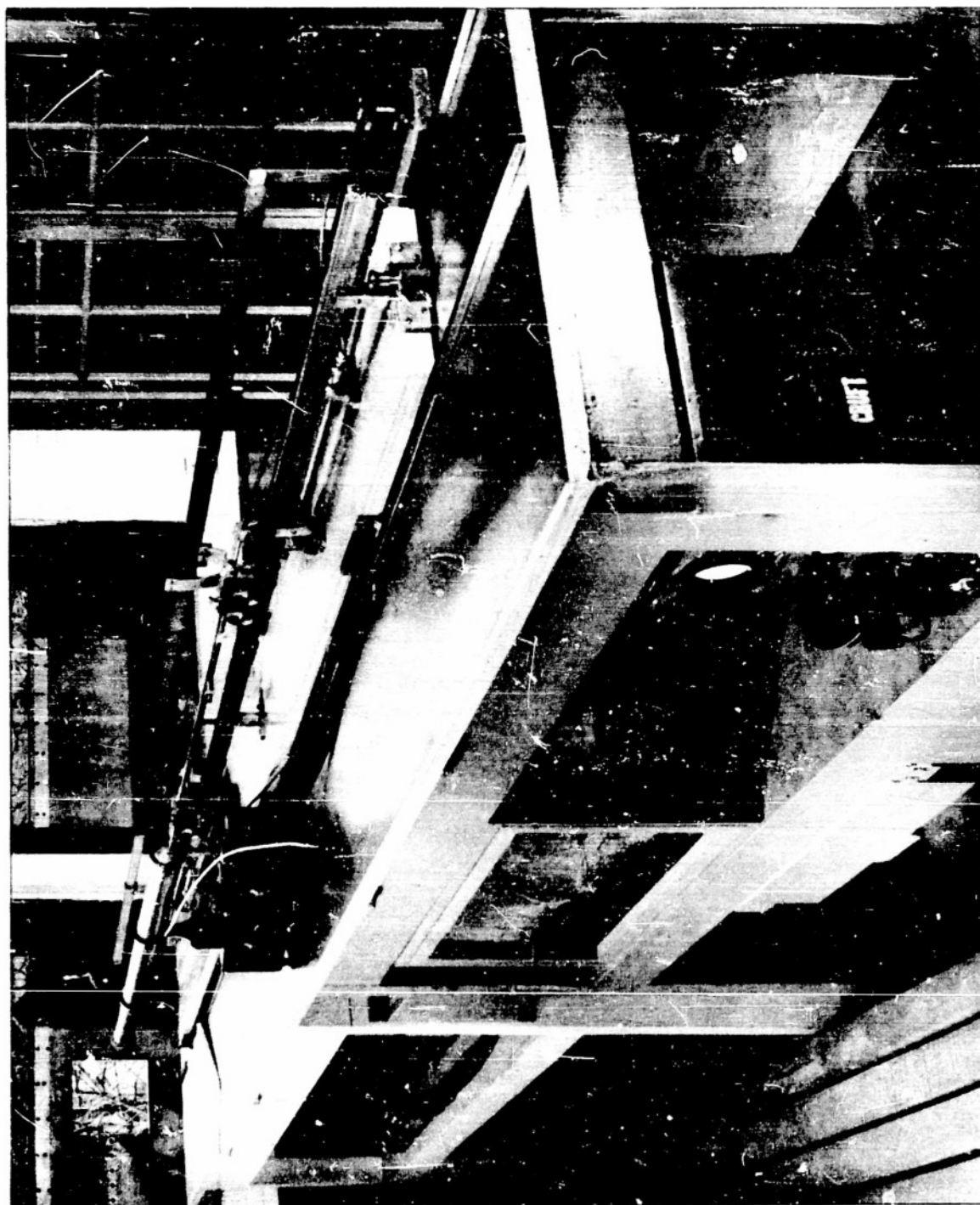


FIG. 3 OPERATING SIDE OF MEASURING LINE

It is slightly larger on one side, being seventeen by twenty feet overall.

The screen is constructed entirely of aluminum structural members. Right angle sections and rectangular box sections are heli-arc welded and bolted together to form the framework which is attached to the building with stainless steel bolts through the brick walls. Allowance in the design was made for a considerable ice load in addition to the weight of the screen itself.

To the framework are attached three one-eighth inch thick aluminum sheets to form the center section of the screen. This is the region where the antennas are to be mounted and, hence, where it is most important that the surface be smooth, rigid, and capable of being conveniently machined. They are removable from the back side in order that they may be machined and eventually replaced if the machine work becomes obsolete. The remainder of the surface of the screen is hardware cloth stretched and screwed to the framework of the screen at frequent intervals. Its wire spacing is one-quarter of an inch and each joint is soldered. The spacing is a very small fraction ($1/100$) of a wavelength at 600 MC and so provides a rugged electrically continuous surface. Although not as rigid as aluminum sheets, it is adequate for the purpose and much lighter and less wind resistant. (See Fig. 2 for the general arrangement.)

The center section's arrangement allows two experimental set-ups to operate in conjunction with each other, one through each of the windows behind the image screen. There is a platform behind the screen that allows one to work closer to it than is possible from the windows.

A small ten inch square beveled hole has been cut into one of the center sections, and into this may be set a window of similar size. Upon this is mounted whatever image screen equipment is used in a particular experimental set-up. It may be removed to bring it in out of the weather or interchanged with another for a different problem.

There is a socket in the window to align the main coaxial line with a hole in this window and the main line plugs into it when measurements are being made.

3 Choice of Frequency

The factors to be considered here are many and varied. Among them are the size of the image screen used, the antenna lengths of interest, the size of the room in which the equipment is to be located and the transmitting and detecting equipment available. All of these effect the final accuracy of the measurements and the convenience with which they are made.

Antenna lengths at least equal to one wavelength are of interest and it would be advantageous if those two wavelengths in half-height could be investigated. The antenna is extended horizontally out from the image screen. A probe slides inside it so that the smallest usable diameter is about one-quarter of an inch. A tube of this size can be no longer than one meter and be able to support itself. This indicates a wavelength of the order of fifty centimeters. On the basis of one-quarter inch diameter, a height h equal to a half wavelength and an $\Omega = 2 \pi (2h/a) = 10$, a wavelength of 100 centimeters is desirable. A wavelength of 50 centimeters would give $\Omega = 8.6$ for the same height antenna and $\Omega = 10$ for a half-height of a full wavelength.

The size of the image screen enters directly into the error inherently present in the measurements. Storer¹ has found that the error in the measured impedance of an antenna mounted in the center of a circular image screen varies inversely as the diameter of the screen. His expression for the center-driven antenna is

$$Z - Z_0 = \frac{9.55}{\pi(d/\lambda)} e^{2\pi i(d/\lambda)} \left| \frac{2\pi}{\lambda} \int_0^h \frac{I(z)}{I(0)} dz \right|^2$$

where $I(0)$ is the driving-point current, d is the diameter of the screen, λ is the free space wavelength, Z_0 is the impedance for an infinite screen, and Z the measured impedance. In some unpublished work he has extended his theory to find that the error of an antenna in the center of a square screen varies inversely as the $3/2$ power of the screen size. If the antenna is placed off center in a square screen, the error varies inversely as the square of the screen size. For a circular screen 10 wavelengths in diameter, he found an error of about $2\frac{1}{2}$ percent in the impedance of a half-wave dipole. The error is $1\frac{1}{4}$ percent for a screen 20 wavelengths in diameter. A square

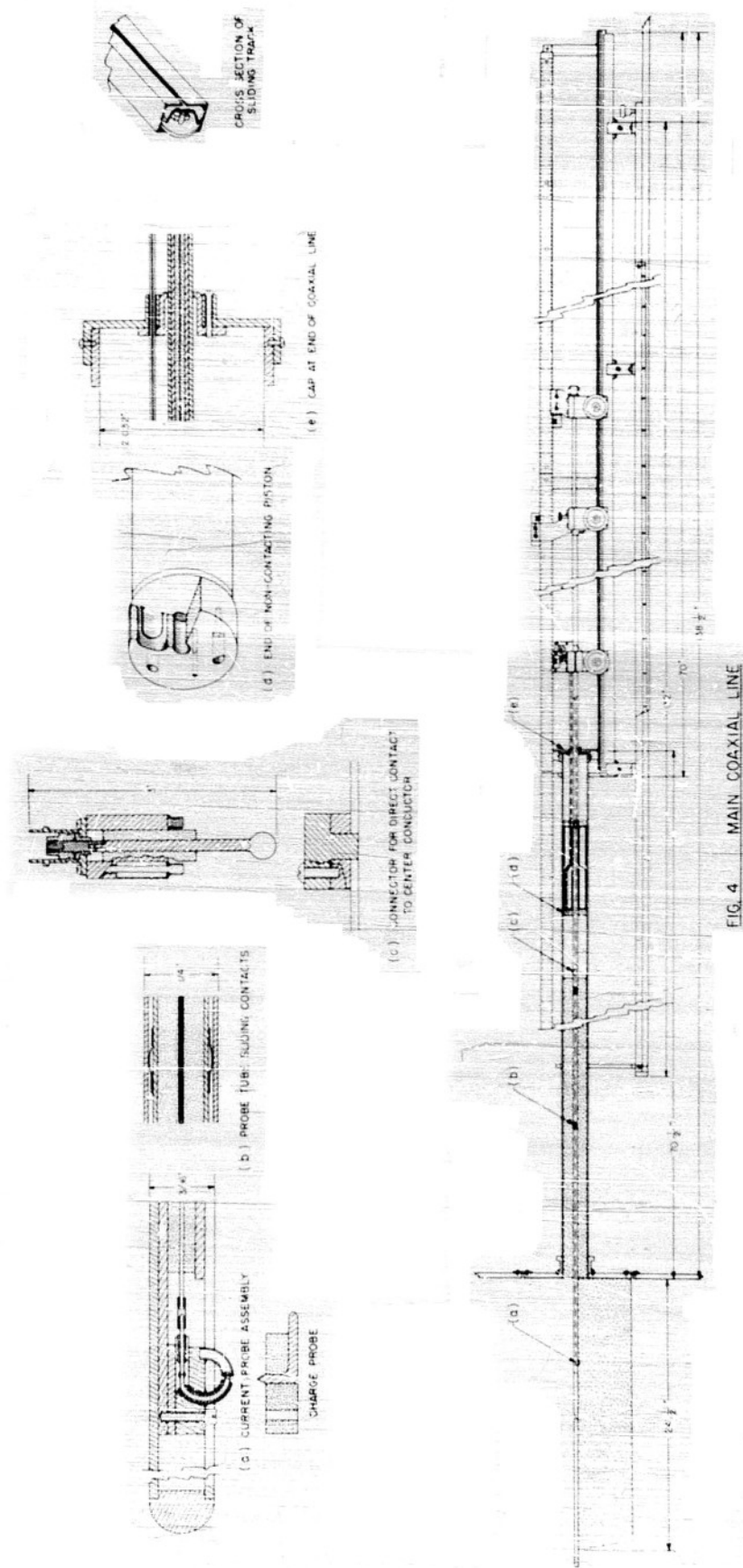


FIG. 4 MAIN COAXIAL LINE

screen 10 wavelengths on a side with the antenna mounted off center would give an error of about one-half percent. This is certainly quite satisfactory if it is assumed that the error of any impedance measurement on the screen is of about the same magnitude as for a halfwave dipole. This assumption seems reasonable.

The screen is 5.2 by 6.1 meters. By taking 5.5 meters as the size of an equivalent square screen, it is found that if the screen is to be at least ten wavelengths square, the wavelength must be 55 centimeters or less.

The portions of the main coaxial measuring line, and the lengths of the associated scales and the rack and pinion drive for the probes and the antenna are directly related to the wavelength to be used. Of some consideration is the measuring technique, for example, the main coaxial line must be considerably larger if current or voltage distribution techniques as well as resonance techniques are to be employed. The operating position of the finished line is shown in Fig. 3 to illustrate the following discussion.

The line and probe scales used in the measurement of distributions in the line take up a considerable amount of room. To measure an impedance using the standing-wave-ratio technique requires at least a half wavelength and a full wavelength is better. Considering that field-distortion effects at each termination of the line require about a half wavelength to be attenuated completely, approximately two wavelengths of line are needed to make use of standing-wave ratio and distribution measuring techniques. The resonance technique uses a movable short circuit at the end of the line away from the load and requires a minimum of a half wavelength of piston travel in order to measure an impedance. A non-contacting piston is used and is slightly less than a half wavelength long. The resonance method then requires about one wavelength of distance in the pipe. The sum of these two is a total of three wavelengths for the overall length of the main coaxial measuring line.

A series of three scales is needed. Proceeding away from the main line they are used to measure the position of the short circuiting piston, to measure the length of the antenna beyond the image screen, and to measure the position of the probe in the antenna and center conductor relevant to the plane of the image screen. The piston-position scale is the half wavelength

required for the resonance technique. An antenna of length equal to two wavelengths beyond the image screen has already been considered desirable, and the measuring section in the main line requires about one and one-half wavelengths of scale length.

The pipe length plus scale lengths come to a complete assembly length of seven wavelengths. This is to fit comfortably inside the room behind the image screen. The maximum length available is about six meters of which only two thirds should be used for equipment. This results in a wavelength of sixty centimeters being appropriate.

With all of these criteria in mind, especially the size of the allowable image screen error and the space available for the equipment, a wavelength of fifty centimeters (corresponding to a frequency of 600 MC) is reasonable. The overall equipment length is then three and one-half meters. The Q -values of the antenna are a bit low, but not seriously so. These proportions are quite conservative and further consideration shows that the line may be used at frequencies as low as 100 MC. However, this is not convenient, and the error is at least ten to twenty percent at the low end. The maximum reasonable high frequency is limited by the onset of higher modes (at 3000 MC for this pipe) in the coaxial line and to some extent by the size of the probes and the holes in the ground screen. The limit is at least above 2000 MC.

4. The Transmitter

The high-frequency coaxial-line oscillator from a military APT-5 radar jamming transmitter is combined with a regulated power supply and a 1000 cps modulator to provide the transmitter for the equipment. It is capable of an output of 25-40 watts in the frequency range of 350-700 MC. This decreases gradually to 5 watts at 1400 MC. The oscillator tube is a disc-seal power triode of 125 watts plate dissipation. It is mounted in concentric coaxial cavities which tune the plate-grid and grid-cathode circuits separately. The circuit operates as a grounded-grid oscillator. Radio frequency power is coupled out of the plate cavity by means of a movable electric probe. Feedback between the cavities is through both loops and electric probe coupling in order to maintain oscillation over a wide frequency band. The oscillator assembly is mounted on a rugged casting and a blower supplies the

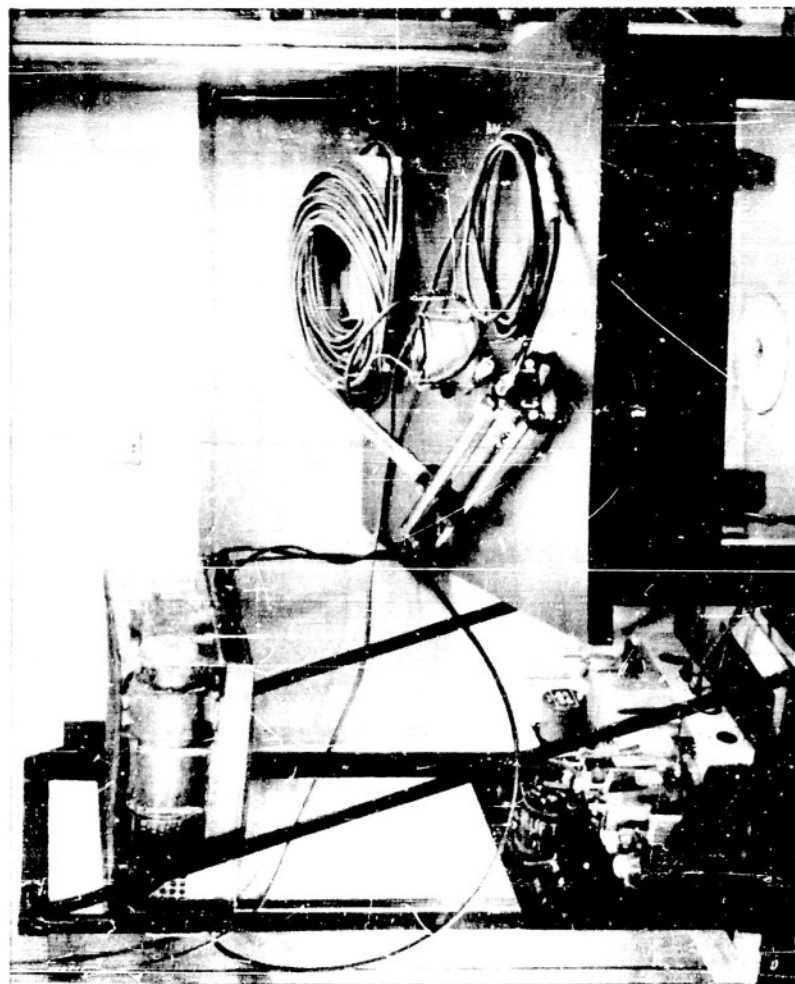
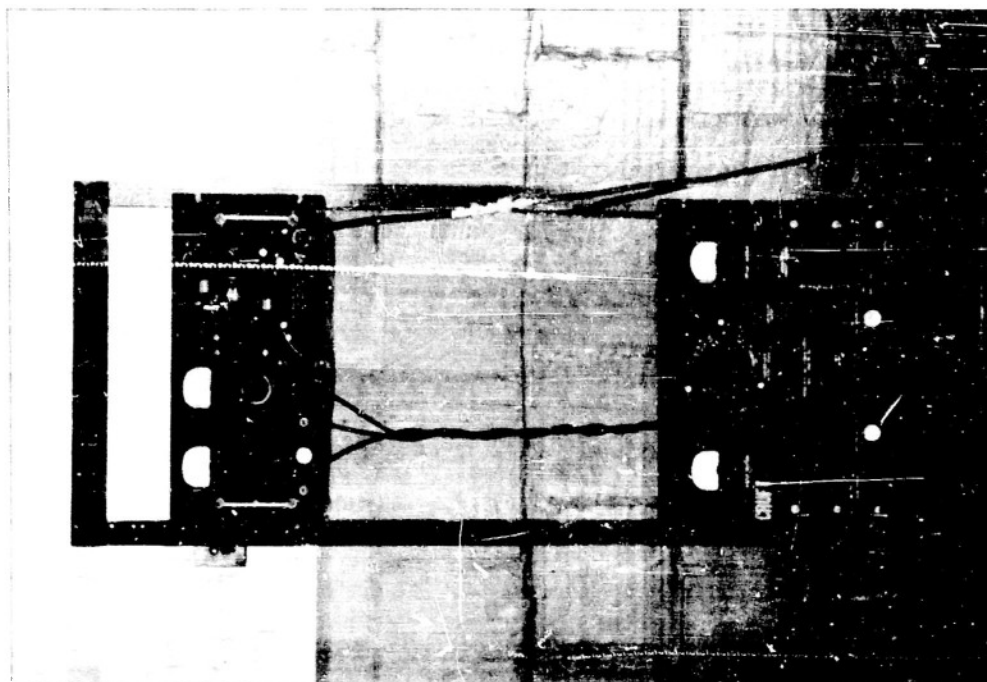


FIG 5 600 MC TRANSMITTER

forced air cooling needed for the tube. It was found necessary to shield the complete oscillator-modulator chassis to minimize the radio frequency power radiated directly by the cavity assembly. Copper window screening securely bonded to the chassis was found to be adequate. The power supply may be operated regulated or unregulated. When regulated it will supply 150 ma at 550-850 volts; unregulated, it is capable of supplying the same current at 0-950 volts.

5. The Main Coaxial-Line Assembly

The main coaxial line, the tubes for positioning the antenna and probes, and the input and output tuning stubs are mounted on an 11 foot long, $\frac{1}{2}$ inch thick plywood plank. The plank is mounted on rollers and a track on top of two 7 foot long tables. When making measurements, the complete assembly is rolled forward two feet through the window behind the image screen and the end of the main line plugs into the aligning socket mounted on the image screen. The end of the line has several one-inch long slots cut into it to assure a good sliding fit into the image-screen socket. The center conductor of the main coaxial line is extended beyond the surface of the image screen to form the antenna to be investigated.² The center conductor is positioned by means of a handwheel and rack and pinion mounted behind the main coaxial line. The distributions of charge and current on this extended center conductor and also on the portion of it inside the coaxial line are measured by means of probes sliding inside the center conductor and projecting through a $\frac{1}{16}$ -inch slot cut into the tube. The probe is positioned by a pinion on the same rack that is used to set the antenna position. The outer conductor of the coaxial line is heavy-walled brass tubing of 2.032 inches inside diameter. The inner conductor is $\frac{1}{4}$ -inch outside-diameter brass tubing with the probe slot running most of its length.

The main coaxial line is completely filled with polyfoam resulting in an electrically and physically continuous structure in which the current and charge distributions are measured in determining a load impedance. Filling the line completely with polyfoam eliminates the difficulties associated with a sagging center conductor or with discontinuous dielectric supports. For all practical purposes the polyfoam is lossless at 600 MC and is sufficiently

uniform to be considered continuous. It affects the electrical characteristics, of course, but the characteristic impedance of the line can be calculated readily after measuring the dielectric constant of the polyfoam.

The velocity of propagation on the line is determined by using a known frequency for the input energy. The wavelength in the line may be measured by obtaining the distance between the sharp minima resulting from the use of a badly mismatched load on the line. Knowing the frequency and the line wavelength, the velocity of propagation can be computed. The line velocity varies inversely as the square root of the dielectric constant, so that this is easily calculated. This is then used to compute the characteristic impedance of the line. The fact that the wavelength in the line is not the free-space wavelength does not complicate the computation of load impedances since the value of β used in the computational procedure is usually irrational for all except a wavelength chosen expressly to make it a more convenient number. The characteristic impedance of the line is 123.6 ohms. It was intentionally chosen this high from consideration of the values of the standing-wave ratio and of ρ that are usually encountered in antenna measurements. [$Z/Z_c = \coth(\rho + j\Phi)$ where Z is the load impedance, Z_c is the characteristic impedance of the line, and ρ and Φ are the variables needed to describe the voltage and current variations on the line using hyperbolic functions.] A value of Z_c in this region has much less-highly-peaked maxima in ρ for the usually encountered antenna impedances than do those resulting from use of the more common 50 to 75 ohm coaxial lines. This is of importance in smoothing data, since it is often of considerable advantage to plot ρ and Φ and smooth these curves before converting to the values for the resistance and reactance.

The pipe is filled with sections of polyfoam four inches long. They were first roughly cut on a bandsaw, drilled for the 1/4-inch inner conductor and then mounted on an arbor to have the larger diameter machined. The outside was thus turned concentric with the inner hole and was made two- to three-thousandths of an inch oversize to insure a snug fit in the pipe. Each section was checked to be sure that it was of reasonably uniform density and then carefully pressed into the pipe. The center conductor was inserted and finally a slot was cut into the polyfoam to make room for the probes to extend through the slot in the center conductor.

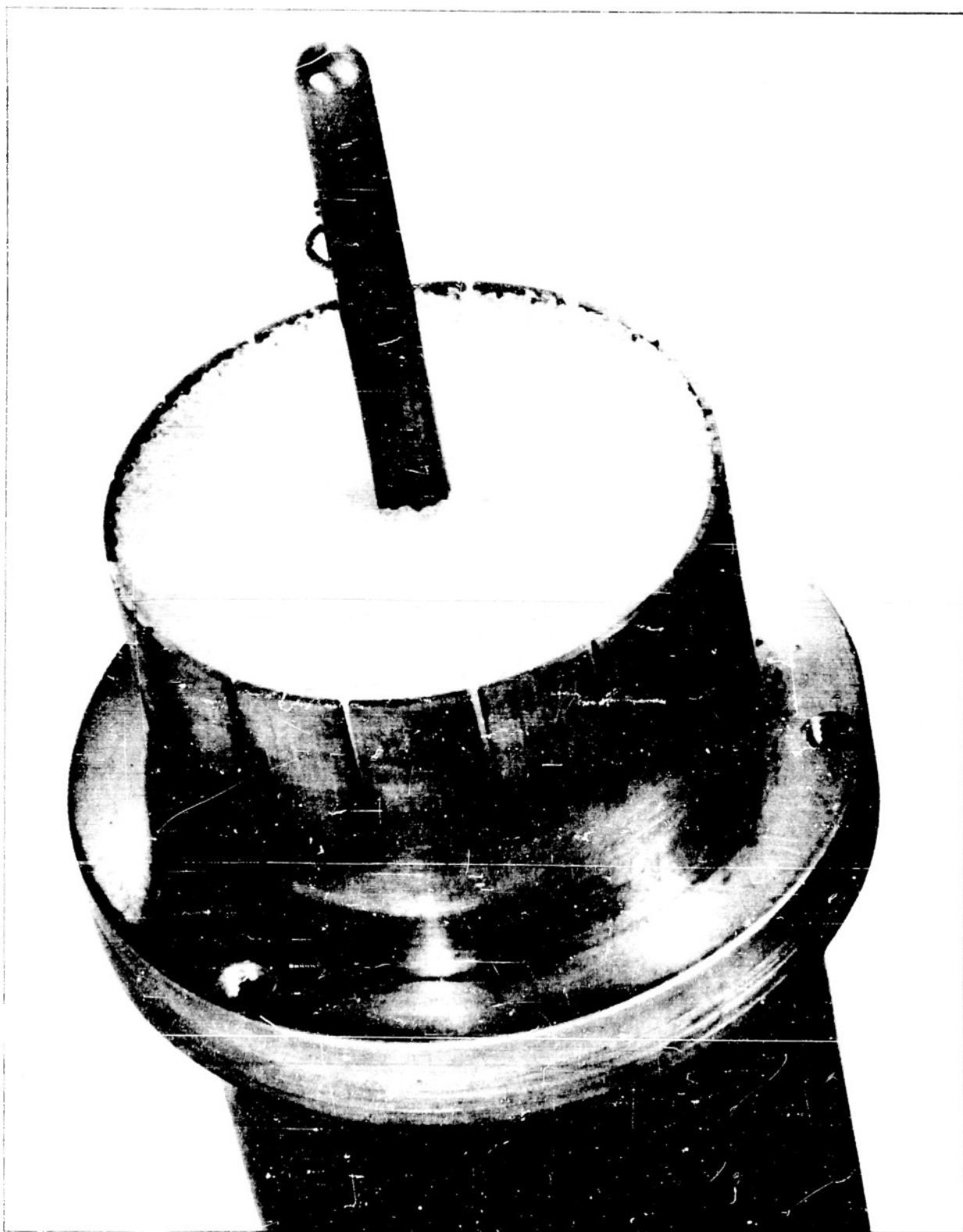


FIG. 7 END OF LINE WITH PROBE

The end of the main coaxial line away from the image screen is terminated by a movable non-contacting short circuit.³ The purposes of this movable piston are to tune the line to resonance for the various loads and to measure a connected load impedance using the resonance technique. This method, or a variation of it, is the only one possible when the antenna structure is being used as a receiving antenna.

The usual short-circuiting piston makes contact with the inner and outer conductor of the line by means of flexible metal fingers. This method has the inevitable difficulty of irregular contacts which are, at least, intermittent and noisy. The various sliding parts erode each other with the result that metal dust accumulates in the line and the contacts eventually wear out or at least become even more difficult to use. Actually, all that is needed is a termination that is a nearly perfectly constant reactance, a short circuit being only one possibility. Such a termination is the sliding choked S-section piston used in this line. The reactance that this non-contacting piston presents to the line need not be known. It is, however, nearly pure and it is as near a constant as can be measured with this equipment. The plane of the short circuit to which it is equivalent is unimportant, but it can be established by using a load of known Z . Such a load could be used in the resonance technique for measuring impedances in order to determine this plane. The immediate quantity usually determined, however, is a reference in the piston position scale corresponding to the known Z of the standard used for a load. This is sufficient information to measure an unknown impedance. The non-contacting piston is not a single frequency device but will operate quite satisfactorily over a $2\frac{1}{2}$ to 1 band. This one has been designed for use in the range from 425 MC to 1000 MC. It would probably operate over a still wider band, but the design should be further checked for parasitic resonances at high frequencies.

The design of the piston is a straightforward procedure.³ An alteration of the termination of the coaxial line behind the piston was found to be admissible and more convenient. Although the design is based upon the assumption that the termination is a matched line, this is difficult to achieve in a small space and over a broad band. A ring of resistors was constructed, for example, and it had as much or more capacitative susceptance than it had conductance. After a few attempts to obtain an untuned matched load, a

short circuit was tried. The feasibility of using this was checked in the following manner. A movable, contacting short-circuiting piston was placed in the other end of the line. Then, as this second piston was moved along the line to a series of positions, the line was tuned to resonance by means of the non-contacting short-circuit and the distance between the two pistons measured each time. This distance was found to be a constant independent of the position of the non-contacting piston. Therefore the reactance of the non-contacting piston was constant and independent of its distance from the short-circuit behind it. Hence, the use of a short circuited termination behind the non-contacting piston was electrically acceptable as well as being much more convenient mechanically.

The structure of the piston is apparent from the cross sectional drawing of Fig. 9a. It is machined of three brass pieces which are soldered together. A thick polystyrene disc is pinned and screwed to each end of the piston to support it away from the pipe and to keep the center coaxial conductor properly spaced so that it does not touch the piston. The piston is positioned in the line by means of two brass rods threaded into one end and extending through the end of the coaxial line to a handwheel and pinion. A small solid-outer conductor coaxial lead passes down through the piston to a 1/2-inch loop on the front face of the piston. This is used to excite the main coaxial line and provides the loose coupling necessary in the resonance technique for measuring impedances.

The load of known Z mentioned above as being the standard used in the resonance technique is a short-circuiting cap fitting over the load end of the line. The cap is designed to place the plane of the short circuit as accurately as possible at the end of the line. This is achieved at the outer conductor by the same means as is used to obtain a good contact at the ground screen socket for the line. The wiping contact for the inner conductor is best seen in the drawing and photograph of Fig. 9c.

Measuring impedances by the distribution methods (standing-wave ratio and width of the distribution minima) requires the use of the current or charge probes which slide inside the 1/4-inch center conductor of the coaxial line. The probes are detachable from the end of the 3/16-inch brass tubing which slides inside the line's center conductor. This 3/16-inch tubing is used to position

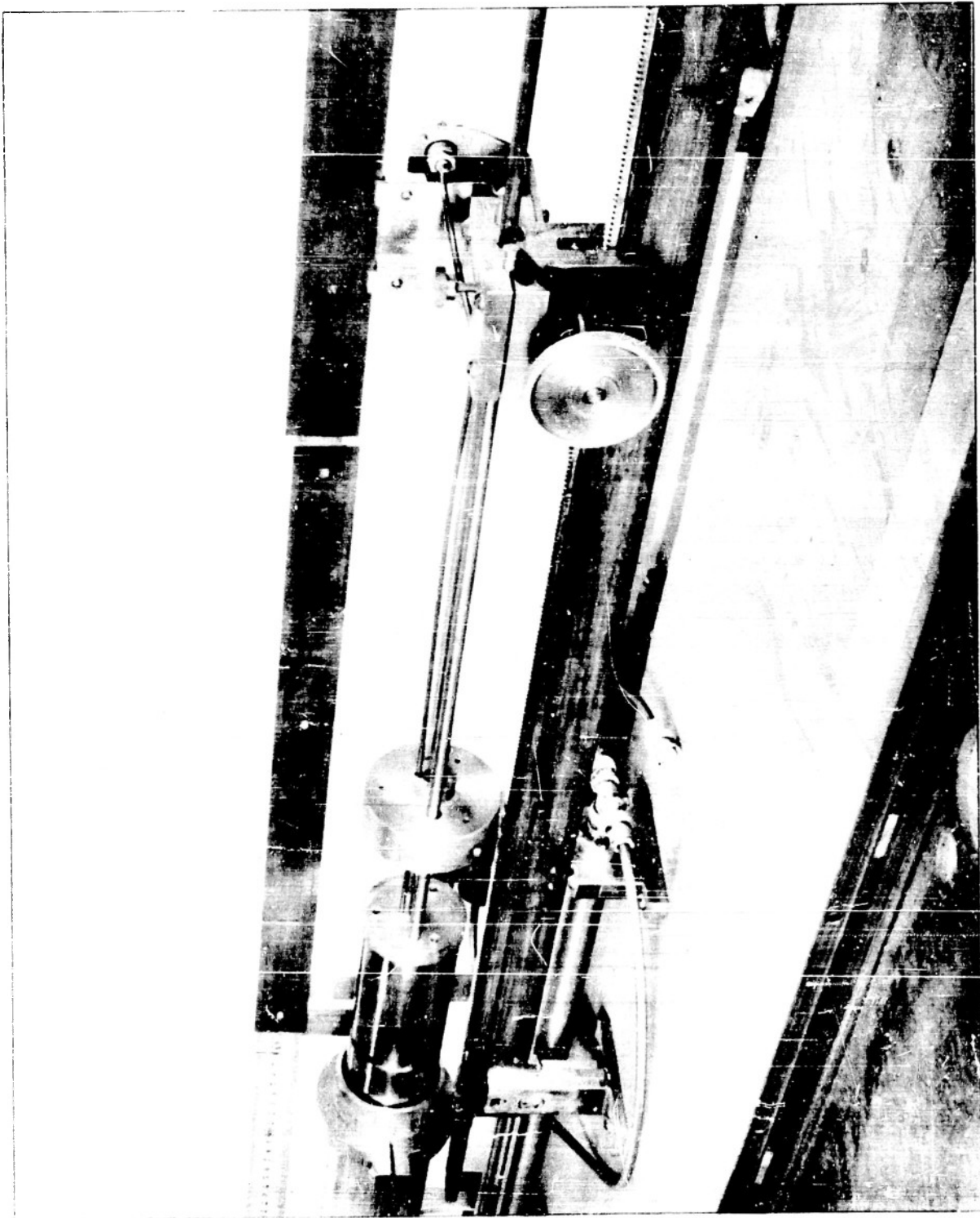


FIG. 8 PISTON ASSEMBLY AND END OF MEASURING LINE

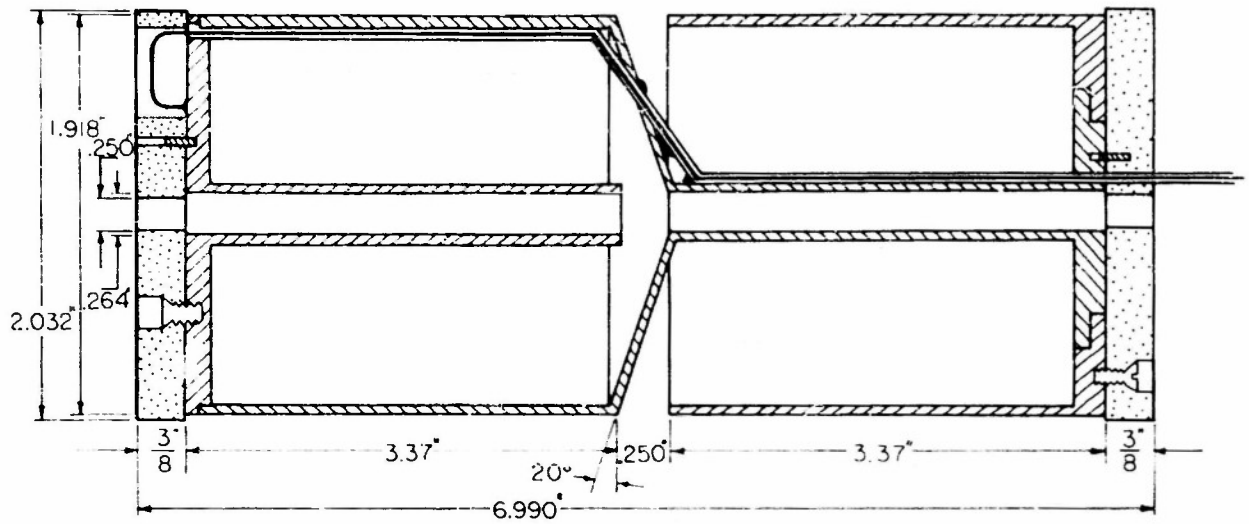


FIG.9-a NON CONTACTING SHORT CIRCUIT

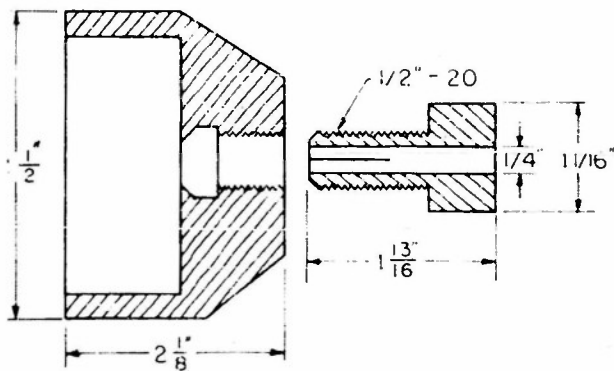


FIG.9-b SHORT CIRCUITING CAP



FIG.9-c SHORT CIRCUITING CAP

the probe and also is the outside conductor of a small coaxial line used to carry the energy from the probe back through the main line and out to the measuring equipment. The charge probe is simply a short piece of wire projecting through the slot into the electric field in the main coaxial line. The current probe is a much more complicated device, in that it is to respond to the current in the center conductor of the main line but not at all to the electric field in which it is immersed and which excites electric probes of the same dimensions quite effectively. The probe is built as a small shielded loop of about 1/10-inch diameter, and it projects into the line about this distance. See Fig. 10 for the constructional details. This type of probe and the effects of the slot on the field distribution have been discussed by Morita.⁴

Some difficulty was experienced with the probe tube as it slid inside the coaxial line's center conductor. There was a quite unpredictable sliding contact between these two tubes, and this caused some intermittent irregularities in the distributions measured in the line and in the antenna. This was especially troublesome when it occurred in the vicinity of a minimum in the line distribution. The solution to the difficulty was to install a series of thin tubular springs soldered to the outside of the 3/16-inch probe tube. These are spaced at 3/4 of a wavelength at 600 MC in order to minimize the possibility of any resonances occurring in the annular space between the conductors.

Steel metric scales are used to measure the probe position and the antenna length beyond the ground screen. These have been suitably located and the indicators used with each positioning handwheel are movable over a small distance so that the minima will read zero when the end of the antenna or the probe is at the surface of the image screen. The two distances are thus immediately available; in particular, in impedance computations, the position of the minimum relative to the end of the line is immediately available from the measurements without an adjustment for a scale reference position. The scales are calibrated in centimeters and millimeters and the verniers can be read to tenths of millimeters. This represents comfortably read distances with an accuracy of at least 0.0002 wavelengths or better than 1/10 of an electrical degree. Measurements of 2/100 degree accuracy are possible but not required in most of the measurements.

With respect to its uses in measuring impedances, the usual coaxial line

is almost completely described by its characteristic impedance and the velocity of propagation in it. The characteristic impedance must be computed and is as constant throughout the line as is the spacing and the dielectric constant of the material between the conductors. The inspection of the polyfoam as a check on its uniformity has already been mentioned and also the care needed in assembling the completed line so that there would be no compressed layers of polyfoam. It is possible to check the spacing of the line by using a charge probe and measuring the distributed capacitance along the line. The center conductor must be insulated from the outer conductor and then an audio oscillator is connected between them. The electric field appearing between the conductors is proportional at each point on the line to the distributed capacitance at that point. The pickup from the electric probe is then checked to be constant over the length of the line. Sorrows, et al,⁵ discuss the general problem of evaluating the errors resulting from imperfect slotted lines. They feel that it is impossible to predict the effect of constructional errors and, in addition, that measurements at audio frequencies, as above, are not a completely reliable indication of the value of a line used in the UHF region. Their method is to match the line as well as possible (less than 1.1) at the frequency to be used and plot the distribution. There will be irregularities in the distribution and the maximum deviation from a smoothed distribution curve is used to determine the possible errors in the measurement of an impedance.

It is of interest to determine the attenuation constant in the line. It is of some importance in measuring impedances, but only when measuring those that give a very high standing-wave ratio. For this line the computed value is .0027 nepers per meter. There are eight effects on the line due to this attenuation and upon any of these measurements of a may be based. As the distance from the load increases, the standing wave ratio decreases, the widths of the minima in the distribution pattern increases, the values of the maxima increase, and the heights of the minima increase. An exactly similar set of four conditions apply to the quantities measured in the resonance technique. The piston distance takes the place of the probe distance from the load, and the resonant maxima and minima interchange respectively with the distribution minima and maxima. Any of these eight effects will allow the determination of the value of a but some are more convenient than others

The use of the changing standing-wave ratio requires that some corrections be incorporated into the computation owing to attenuation in the quarter-wavelength of line between a maximum and a minimum. The increasing width of the minimum in the distribution was found convenient, and an approximate check on the α of the line was made using this method. Two sets of data were taken at two consecutive minima for a parameter varying in a particular antenna structure. The minimum further from the load was always wider, and the values for ρ computed for the further minimum were larger in proportion. The differences in ρ for the two minima were averaged for the data and an average change in ρ obtained. The change $\Delta\rho$ is directly proportional to α and the distance Δw between the minima. Then α is computed using $\alpha = \Delta\rho/\Delta w$. The value obtained is .0026 and is somewhat smaller than theory predicts. It seems reasonable, however, considering the inherent error in the measurement owing to the fact that successive widths were subtracted to give a small difference. The error in the result can then be quite large. It was thought that probe loading on the line was the cause of the low measured value of α but further consideration made this doubtful. The effect of the probe should be almost the same at two successive minima. Then, when the subtraction was made to get the difference between the minima, this loading should be cancelled out leaving no probe effect.

Probe loading effects on the line were checked by using the resonance technique and an arbitrarily chosen fairly reactive dipole as a load on the line. The probe was first placed at a maximum in the distribution pattern and then at a minimum and the impedance measured each time. These two situations should give the greatest variation in probe effects. Several maxima and minima are available in the line. The probe was placed at each one, and the resonant width and positions measured. A definite increase of less than one millimeter was found between the resonant width with the probe placed at a maximum. There was also a slight shift of less than $1/2$ mm in the position of resonance. The probe loading effect was slightly larger than the measured value of the attenuation for a half wavelength of line. The shift in the resonant position was about one-quarter of a degree. The standing-wave ratio for the load used was 5.6. This is larger than has

been encountered in most of the measurements made on this line. The effect of the probe was considered negligible because the loading was only slightly larger than the line attenuation. Taking the effect of the probe into account in an exact manner would considerably complicate the impedance computation and considering the inaccuracy of the experimental representation of the theoretical antenna model, it is not worth the additional inconvenience.

The measuring of the probe capacitance was attempted by using a short circuit as a load. The shift in line resonance was determined when the probe was moved from a charge minimum to a charge maximum point. With a short-circuited load and a standing wave ratio in the hundreds, the probe effect at a charge minimum should be quite negligible so that the measured shunting effect at the maximum would be quite precise. A resonant shift of about $1/4$ mm resulted, and from this a capacitance of $0.0064 \mu\mu f$ was obtained. This is small compared to the $0.27 \mu\mu f$ per centimeter distributed capacitance along the line.

A determination of the wavelength in the line, and hence of β , is necessary in order to compute the ρ and ϕ of a load. For this measurement a very poorly matched load is needed in order to get the sharp minima that may be accurately located. The short circuit previously described as the standard in the resonance technique may be used. Another possibility is the open circuit provided by retracting the center conductor of the coaxial line into the pipe a few tenths of a wavelength. This is not a perfect open circuit at the plane of the end of the center conductor since there are some capacitative and end effects. It is, however, an extremely good capacitative reactance of very little loss, although its use as a standard of ϕ for the resonance technique is not possible without some knowledge of the size of the capacitative end effect. It has much less dissipation than the brass short circuit because the currents in an open circuit are low. It is actually the center conductor of the coaxial line terminated by a waveguide beyond cut off. It avoids the radiation loading of the usual coaxial open circuit. Some difficulty may still be had with some radiation from the open end directly to the sensitive detection equipment if the center conductor is not retracted far enough. In this case the cap may be placed over the end of the line to eliminate completely the radiation from the open end. The currents in the cap are low since

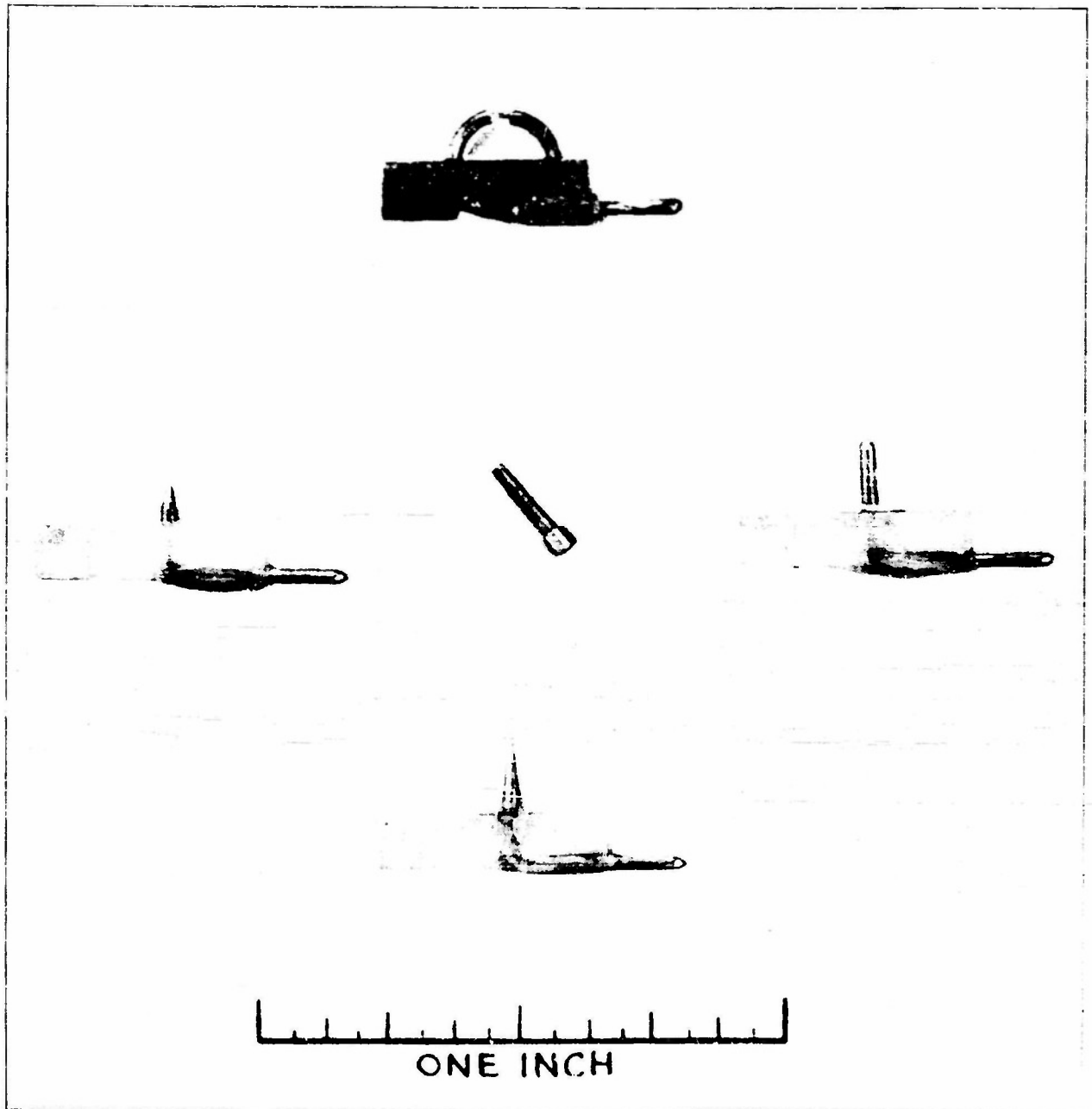


FIG. 10 CURRENT AND CHARGE PROBES

the center conductor is retracted and the quality of the open circuit will be essentially the same as for a well retracted center conductor.

The distributions in the line should be plotted in order to get an accurate measurement of the wavelength. Actually, minimum positions of five-figure accuracy can be obtained conveniently in this way, although such precision is much greater than is usable. The non-uniformity in the sizes of the tubing and in the polyfoam spacing material causes the most difficulty. A scatter of one part in 3000 was found in the wavelength determined in various parts of the coaxial line. This represents an error of 1/10 degree at 600 MC which is quite adequate for most impedance measurements.

The value of the dielectric constant must also be determined and follows immediately from the measured wavelength λ_c in the coaxial line, the free space wavelength λ_s , and $\epsilon_r = (\lambda_s / \lambda_c)^2$. The input frequency can be set conveniently and maintained for a reasonable length of time to one part in 100,000 using the frequency standard, so that an accurate average value of the relative dielectric constant over a half wavelength is available. The polyfoam used in this line has an $\epsilon_r = 1.033$.

Incidental to the determination of wavelength is the setting of the probe vernier indicator so that it reads zero when the probe is at the plane of the image screen and at the end of the coaxial line. In an impedance measurement the apparent impedance at the plane of the image screen is the desired result. Hence, the distance of a minimum in the line distribution from the end of the line should be available immediately from the equipment. In order to set the indicator, either the plane of pickup of the probe must be known and the setting made by a purely mechanical measurement of lengths, or else a load of precisely known ϕ must be available. The plane of pickup of the probe may be more or less located by inspection but it could hardly be found more accurately than to 1/32 of an inch. A load of known ϕ is available, however, in the short circuit used as the standard in the resonance technique. It also has the low loss that will give sharp minima. This short circuiting cap was used and the indicators set so that, for the first minimum in the current distribution, the indicator reads exactly a quarter wavelength from the end of the scale.

6. The Coaxial Low-Pass Filter

The transmitter is a simple free-running oscillator. Its amplitude is limited only by the nonlinearity of the tube characteristics so that, of necessity, harmonics are plentiful. The tube operates in tuned high-Q cavities, but nevertheless a fairly large percentage of harmonic is unavoidably coupled out of them. The main coaxial-line equipment has several stubs and line stretchers to match the transmitter to the line input and the probe output to the receiver in addition to the piston in the main line. It is quite possible that some harmonic of the transmitter frequency may be tuned simultaneously with the 600 MC fundamental to give a large signal in the line output. Even if such actual resonances do not occur, the 600 MC minima in the line may be obscured by the presence of some harmonic maximum occurring at the same point. It is true that the receiver should not respond to the harmonics, but it has spurious responses and it is not to be relied upon in this respect. It is possible, in any case, that harmonics may be picked up on the line and make the measurements meaningless or at least confusing. A filter was therefore deemed necessary to suppress the transmitter's higher harmonics. It would certainly be necessary in a crystal detector-tuned amplifier combination since then the selectivity of the receiver would not be present and there would be only the very uncertain possibility of harmonic suppression by the tuning stubs.

The final design consists of an 11 section constant-K low-pass filter with an additional matching section on each end.⁶ As a filter it may be considered on the usual lumped constant basis as a periodic arrangement of inductors and capacitors. Its elements, however, are realized in the coaxial structure appropriate to the 600 MC frequency. The elements may be considered in at least two ways. The first considers the inductive elements to be merely short lengths of wire with a corresponding inductance, and the capacitive sections to be cylindrical capacitors. It may also be considered to be a series of short coaxial lines of varying impedance, and because of this, it called a short line filter. On a line basis, sections of high impedance -- each short length of wire being a high impedance line -- alternate with sections of low impedance, viz. the short polystyrene-supported coaxial line. Now, a short length of coaxial

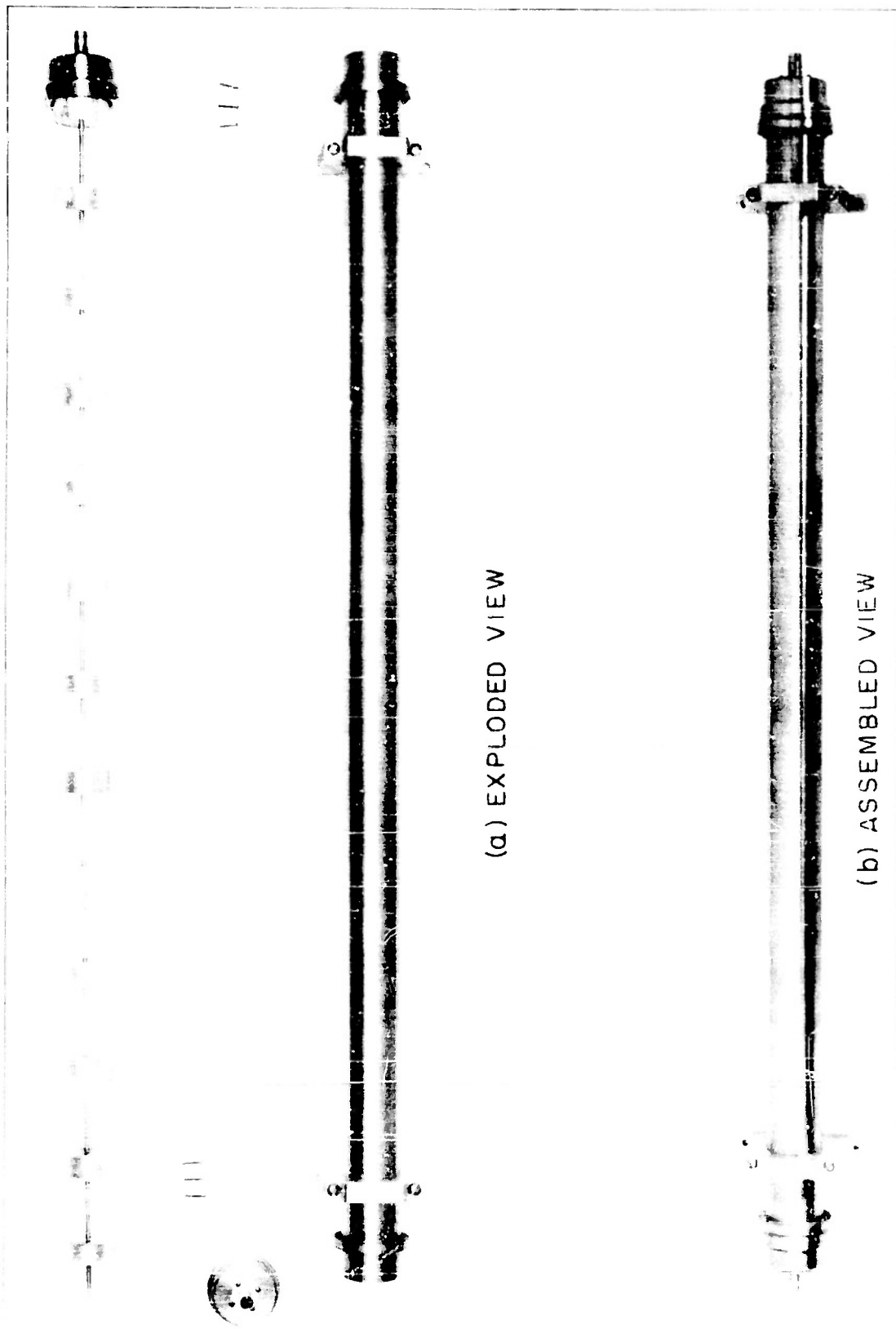


FIG. 11 700 MC LOW PASS FILTER

line terminated in an impedance low compared to its characteristic impedance appears to be inductive as a function of frequency; also a short length of line terminated in a comparatively high impedance appears to be capacitive as a function of frequency. This is the situation that exists in this low-pass filter as can be seen in the figure. On either basis there are alternating sections of series inductance and shunt capacitance necessary to form a low-pass filter structure. The impedance of the capacitive line sections has been further reduced by the use of polystyrene-supporting rings about the very large center conductors. They also provide the necessary mechanical support. This type of filter (the action of whose elements depends upon the variation of their apparent physical size with the frequency) necessarily has spurious pass bands owing to higher order resonances of the supposedly short lines. These are well above cutoff, however, and their positions may be located approximately when the length of the capacitive and inductive elements are determined.

The function of the filter is to suppress the higher harmonics generated in the 600 MC transmitter and to pass the fundamental with as little attenuation as possible. The second harmonic is at 1200 MC so that the cutoff frequency may be set comfortably above 600 MC and the filter will perform properly. 700 MC was chosen as the cutoff frequency with this in mind, and also because experience indicated that the actual filter cutoff was usually a few percent below the design cutoff.

A constant-K filter is supposed to operate into its image impedance which is not independent of frequency as is the 50 ohm characteristic impedance of the coaxial lines used with the equipment. To overcome this difficulty, a matching section is used on each end of the filter to match from the varying image impedance of the internal sections of the filter to the constant load impedance. This was achieved by using a full T-section with the same 50 ohm low frequency image impedance but with a cutoff frequency f_c equal to 1.3 times the cutoff frequency of the internal sections, that is, with a cutoff of 920 MC. This arrangement gives essentially zero insertion loss up to at least $0.9 f_c$ of the internal sections. This may be compared with a filter with no end sections which would have little loss up to $0.5 f_c$ or with a filter designed throughout with a low frequency image impedance equal to 1.4 times the load impedance which would be flat up to $0.7 f_c$.

The design was quite satisfactory. See Fig. 13 for the measured insertion loss and the experimental setup used to obtain it. The insertion loss was measured with no padding on either end, padded with 10 db (at 700 MC) on only one end, and padded with 10 db (at 700 MC) at both ends. The padding was lossy cable of 50-ohm characteristic impedance which is the low frequency image impedance of the filter sections. The insertion loss is flat and zero within 1/2 db when padded on both sides and when unpadded has 2-3 db irregularities, especially in the vicinity of cutoff. An incidental difficulty for the unpadded case is that the calibration of the signal generator is not entirely trustworthy when the generator operates into a load which varies with frequency. The filter is designed to be padded, but its characteristics are quite reasonable without padding. The actual cutoff frequency is approximately 680 MC, which is therefore low by 3 percent. This seems to be usual in this design.

7. The Measurement of Frequency

The frequency of the transmitter must be either fixed and known because it is crystal controlled or else there must be some convenient way of measuring and resetting its frequency. The frequency of a crystal controlled transmitter is difficult to change over any great range because the long chain of harmonic generators and power amplifiers do not have a very wide bandwidth. This chain of harmonic generators is necessary in order to attain an output frequency in the UHF or VHF region. Achieving frequency stability in this way is a cumbersome process. The alternative of periodically measuring the frequency of a power oscillator with a wavemeter or a slotted line is likewise inconvenient. In addition, the conveniently attainable accuracy of this technique can be little better than about one tenth of one percent.

Another possibility, and the one used in this equipment, is the use of a spectrum analyzer. This gives a presentation on a cathode-ray-tube screen of the signals present in a certain frequency region. Energy is fed into this device from the equipment's transmitter and also the signal from a lower frequency and lower powered crystal controlled source. Then, one has on the linear frequency time-base a separate deflection due to each of these, and it is possible to make a continuous visual check of the transmitter's frequency relative to the frequency of the standard. Adjustment of the transmitter amounts

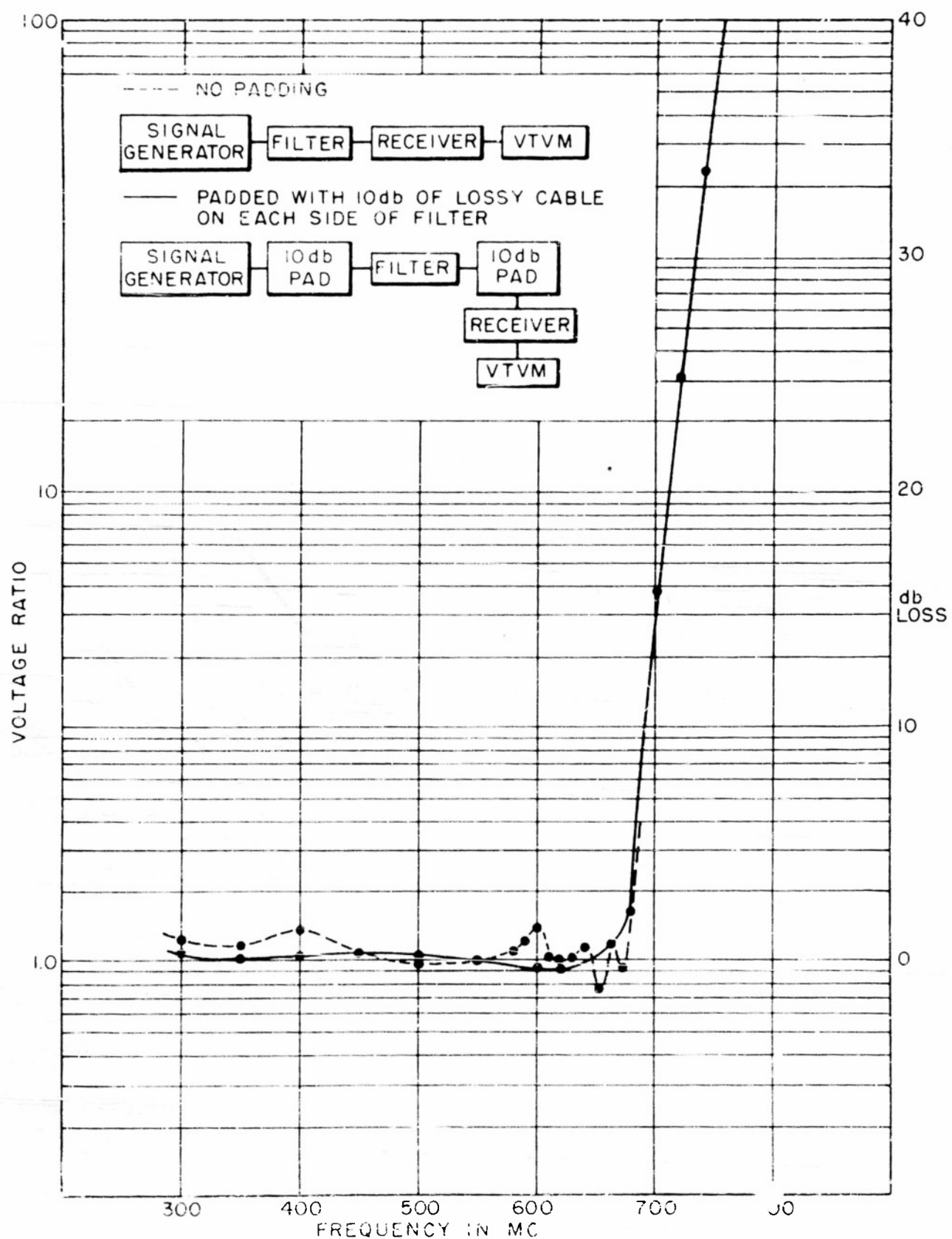
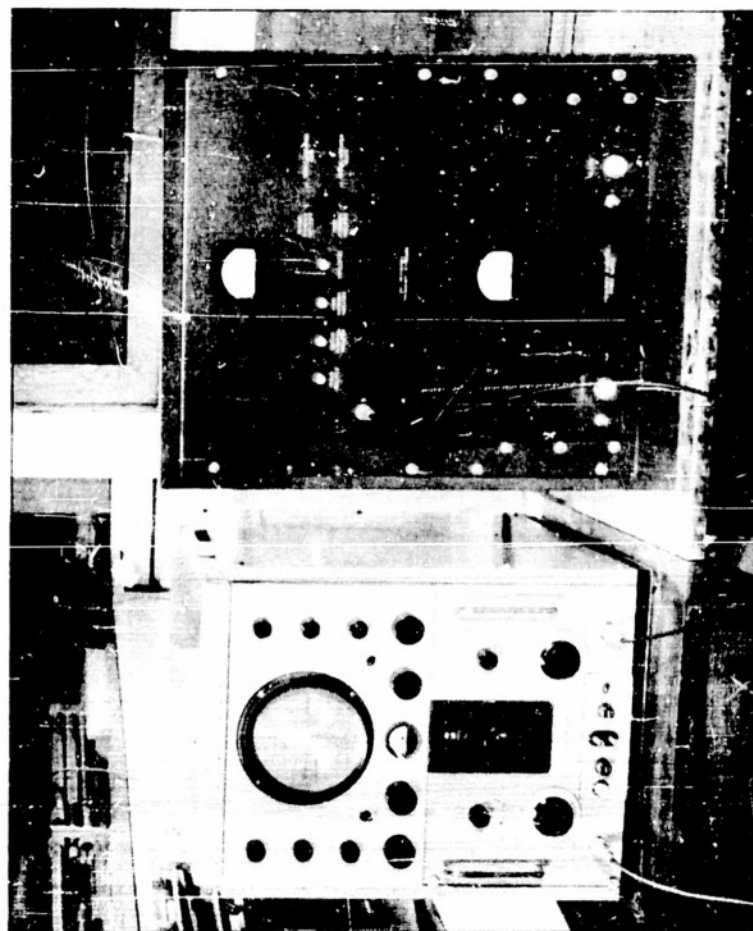
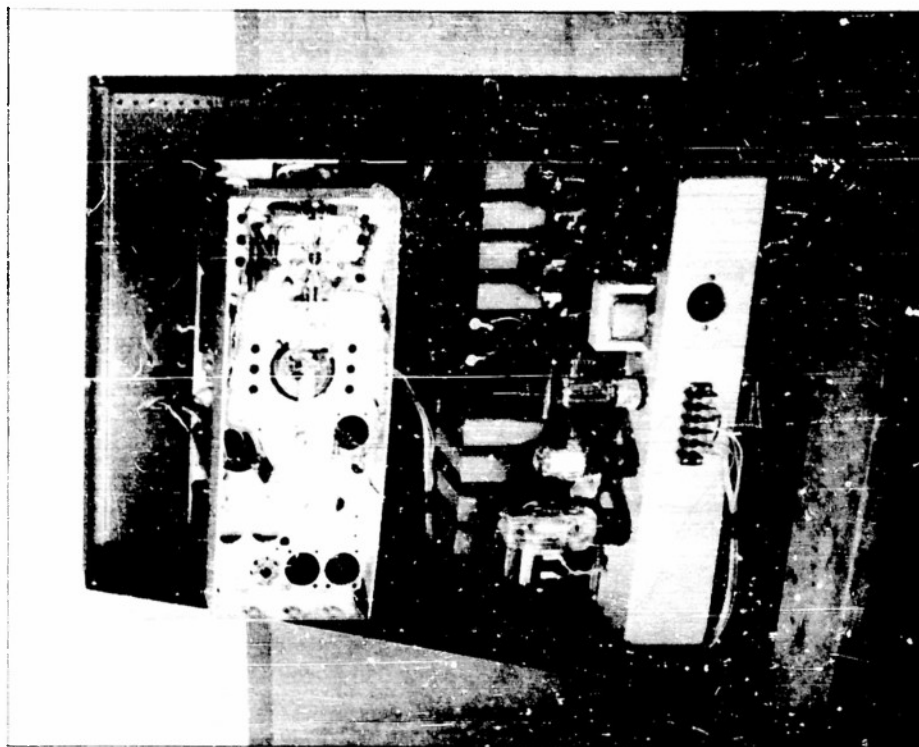


FIG. 13 700 MC LOW PASS FILTER INSERTION LOSS



(a) SPECTRUM ANALYSER
AND FREQUENCY STANDARD



(b) BACK OF FREQUENCY
STANDARD

FIGURE 14

to moving the signal of the transmitter into coincidence with that of the frequency standard. The adjustment is very quickly and easily made since the frequency difference of the two is clearly visible, and the adjustment necessary for coincidence is obvious and readily controlled.

A source of crystal controlled energy at 150 MC is actually used as the frequency standard. This is much lower than the frequency to be measured but still is considerably higher than is actually necessary. It was used because it was available, but a more easily constructed lower frequency standard could be used instead. The signal from the standard is fed into the spectrum analyser and then the harmonics of the standard generated in the crystal of the spectrum analyser's input circuits are coincident with the fundamental and harmonics of the transmitter's signal likewise generated in the crystal. Actually a spectrum analyser covering the 2400-3600 MC frequency band was used, and hence the coincidences observed on the cathode ray tube are not the fundamental of the transmitter and a harmonic of the standard frequency. The ones actually used are the 4th harmonic of 600 MC and the 16th harmonic of 150 MC occurring at 2400 MC. If the equipment used for the standard had not been available, a crystal oscillator and power amplifier operating at 50 MC would have been constructed. This would not have been as difficult as a series of harmonic generators and power amplifiers with several watts output at 600 MC.

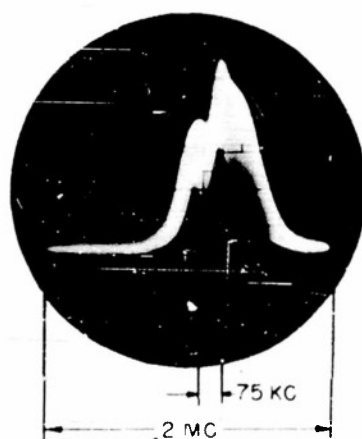
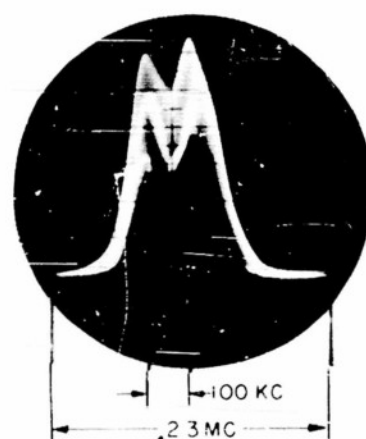
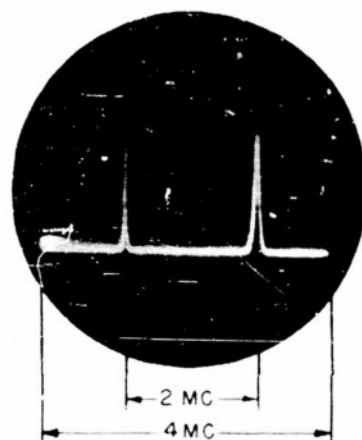
The part of the frequency band presented in the spectrum analyser is adjustable over a wide range from zero to several megacycles. It can be reduced conveniently to have a bandwidth of 100-200 KC. This is the sweep used under usual circumstances. When setting the transmitter frequency originally a slotted line is used to set the wavelength within two or three tenths of a percent. After this the signal will appear in the vicinity of the frequency standard's harmonic on the spectrum analyser, and can be adjusted to the final frequency. A sweep of 1 to 2 MC is necessary when the set is first turned on and during the initial temperature and frequency-drift time of about an hour. With a sweep this wide, the signal appears immediately and drifts toward coincidence with the standard signal. After warmup, a much narrower sweep is used for a precise adjustment and to monitor the transmitter's frequency during the course of measurement. Zero beat of the transmitter and frequency standard's signal could actually be observed, but the transmit-

ter's stability is not sufficient to hold this for more than a moment. A direct measurement of frequency-sweep speeds and frequency deviation can be made by using the fact that the half voltage width of the signals on the face of the tube is 33 KC. In operation it is necessary to monitor the transmitter's drift. This is, at most, about 100 KC after several minutes warmup, and during final operations there is a random drift of about 10 KC. The transmitter's random stability after 15 minutes warmup was then of the order of one part in 10,000. After being on an hour the stability was of the order of one part in 100,000. This is more than adequate for impedance measurements.

The frequencies that may be checked using this technique are by no means limited only to the 600 MC at which the equipment usually operated. The frequencies are not continuous, but there are many harmonic combinations of the 150 MC standard, and harmonic combinations of a transmitter which occur in the 2400 to 3600 MC range of this spectrum analyser. In the range between 550 MC and 650 MC for example there are eight frequencies (550, 570, 575, 600, 625, 630, 638 and 659 MC) which may be checked by suitable coincidences. The 550 MC measurement, for example, is a coincidence of its 6th harmonic with the 22nd harmonic of 150 MC at 3300 MC. In the region from 300 to 1200 MC there are about 90 check points involving up to the 25th harmonic of the 150 MC standard and up to the 12th harmonic of the checked frequency. If a lower frequency standard were to be used, many more combinations would be possible.

The frequency standard itself is a modified BC-625 crystal controlled military aircraft transmitter. The modulator circuits have been removed and a new power supply added. The crystal is in a Pierce oscillator which doubles the crystal frequency in the plate circuit. Following this are two triplers giving a total frequency multiplication of 18 times. A power amplifier is last and is capable of delivering 12 watts maximum to a matched load. A free space wavelength of 50 cm or a frequency of 599.558 MC is desired as the transmitter's frequency with which the standard will be compared. Dividing this by four gives 149.889 as the standard's output frequency and then by 18 to give the crystal's frequency of 8.3272 MC.

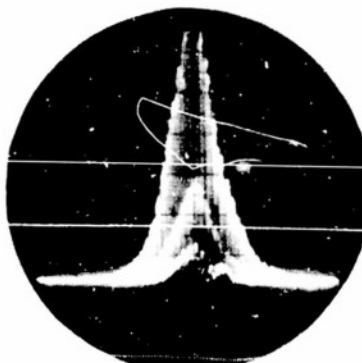
It was necessary to regrind a 8.325 MC crystal to give the exact desired



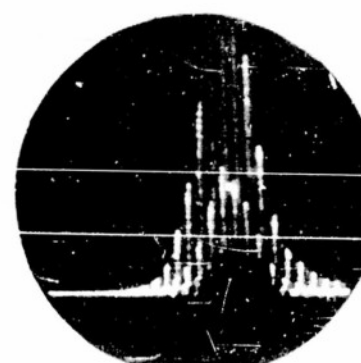
LESS THAN 5KC DEVIATION



LESS THAN 1000cps
DEVIATION
(2 SEC EXPOSURE)



NEAR COINCIDENCE
(1/8 SEC EXPOSURE)



VERY NEAR COINCIDENCE
(1/8 SEC EXPOSURE)

FIG 15 SPECTRUM ANALYSER PRESENTATION WITH VARIOUS DEVIATIONS OF
THE TRANSMITTER FROM THE FREQUENCY STANDARD

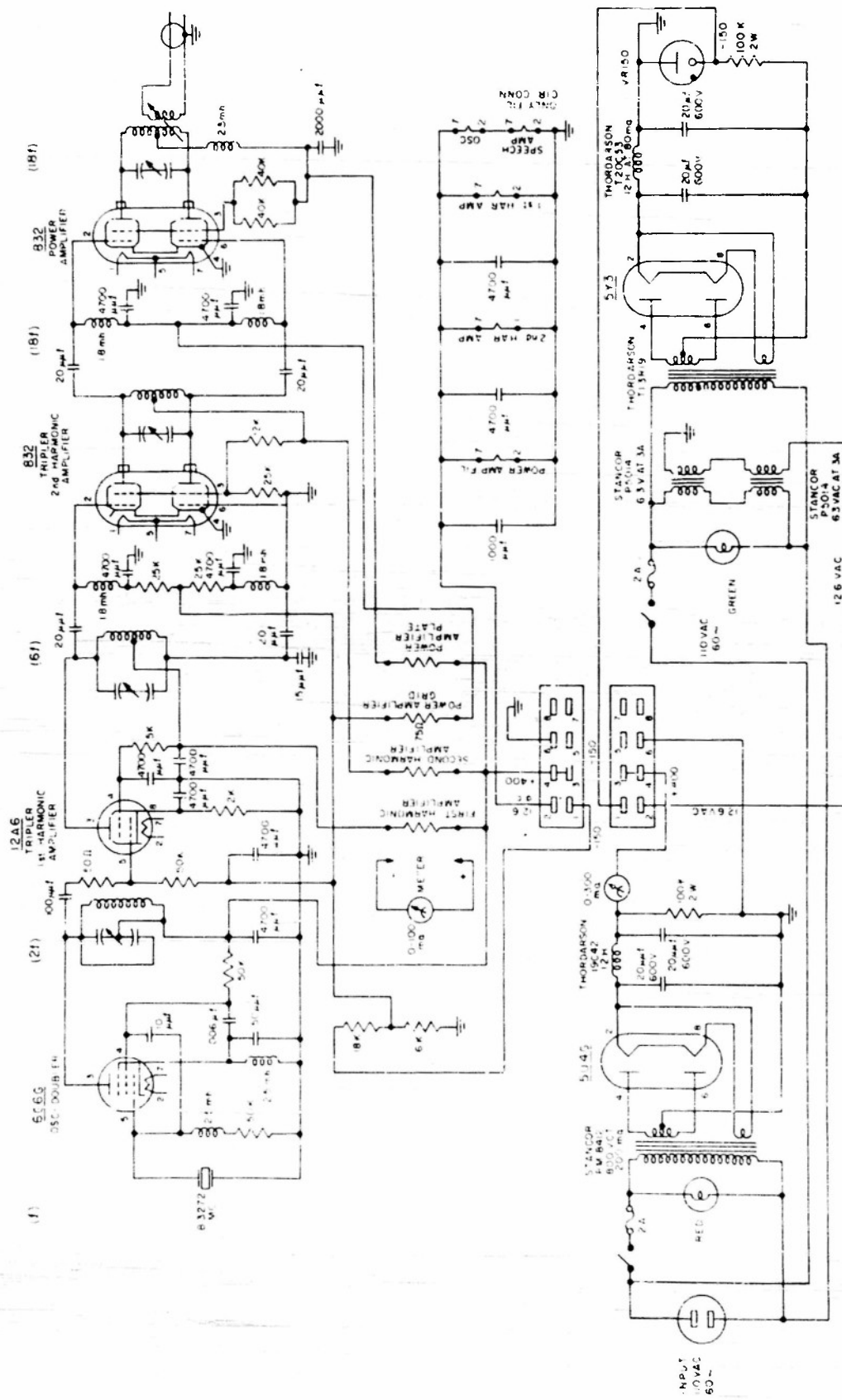


FIG 16 FREQUENCY STANDARD

frequency. During this process a General Radio Heterodyne Frequency Meter was used to compare the standard's output with WWV's transmission at 10 MC. The process consisted of calibrating the frequency meter using WWV rather than its own internal crystal. This may be done by connecting the frequency meter's input terminal and an antenna to a receiver. Some of the energy from the meter's variable frequency oscillator leaks from the meter's terminal into the receiver and a zero beat of the variable frequency oscillator with WWV may be obtained in the receiver.

The frequency adjustment of the crystal required removing the quartz plate from its holder and grinding each side to increase the frequency or loading it with inked lines to decrease its resonant frequency. The loading was found necessary because it was not found possible to grind it sufficiently carefully to obtain the correct frequency. A continual check with WWV was maintained throughout the process. The crystal was originally too low in frequency, but was ground until it was slightly high. It was then loaded and so brought down to the correct frequency. The most that it could be decreased by this process was about five parts in 10,000. It was easily possible to set the frequency of oscillation to within two parts in 100,000 which is more than adequate in this application.

This frequency measuring technique using a spectrum analyser is quite satisfactory. There is considerable convenience in having a continuous visual check in the transmitter's frequency relative to a standard. The system has most of the accuracy of an ordinary crystal source and is certainly adequate for the usual impedance measuring.

8. Signal Detection

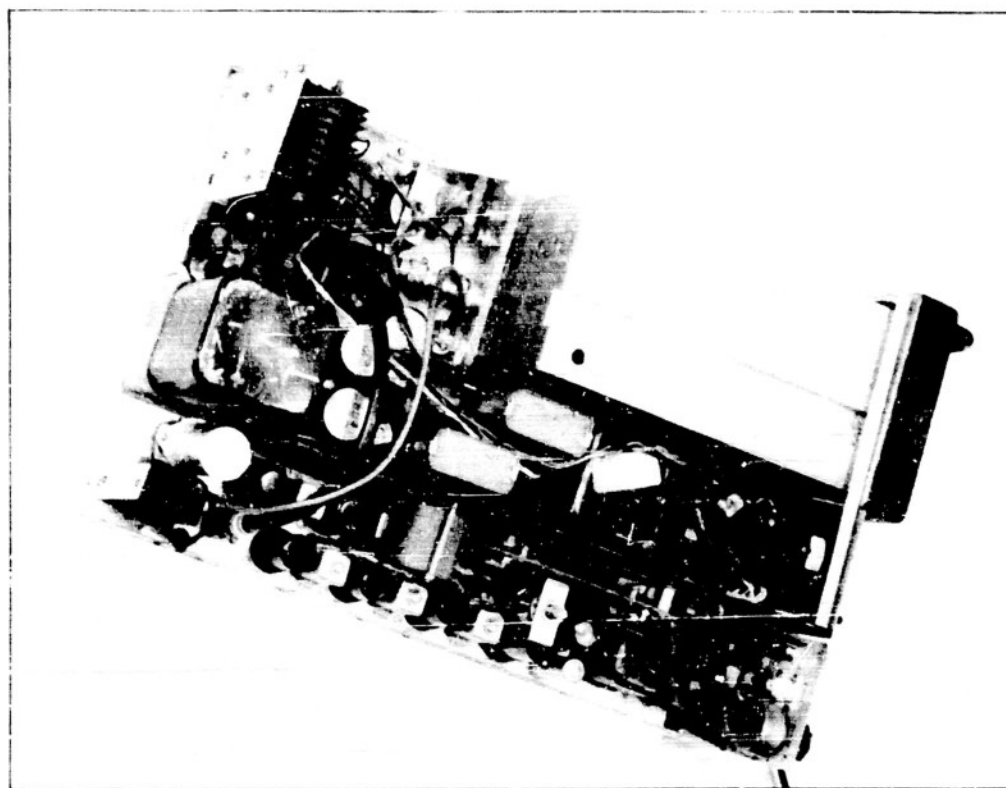
The measurement of the signal levels available from the probes requires amplification of some kind either of the signal itself or of its modulation. The usual arrangements are to have an audio-modulated signal and detection equipment consisting of a silicon crystal and a tuned amplifier or a superheterodyne receiver with tuned audio stages. The audio output of these is then taken as an indication of the radio-frequency signal level.

Several of the above combinations were tried using available equipment in order to find the one having the best signal-to-noise ratio. A signal gener-

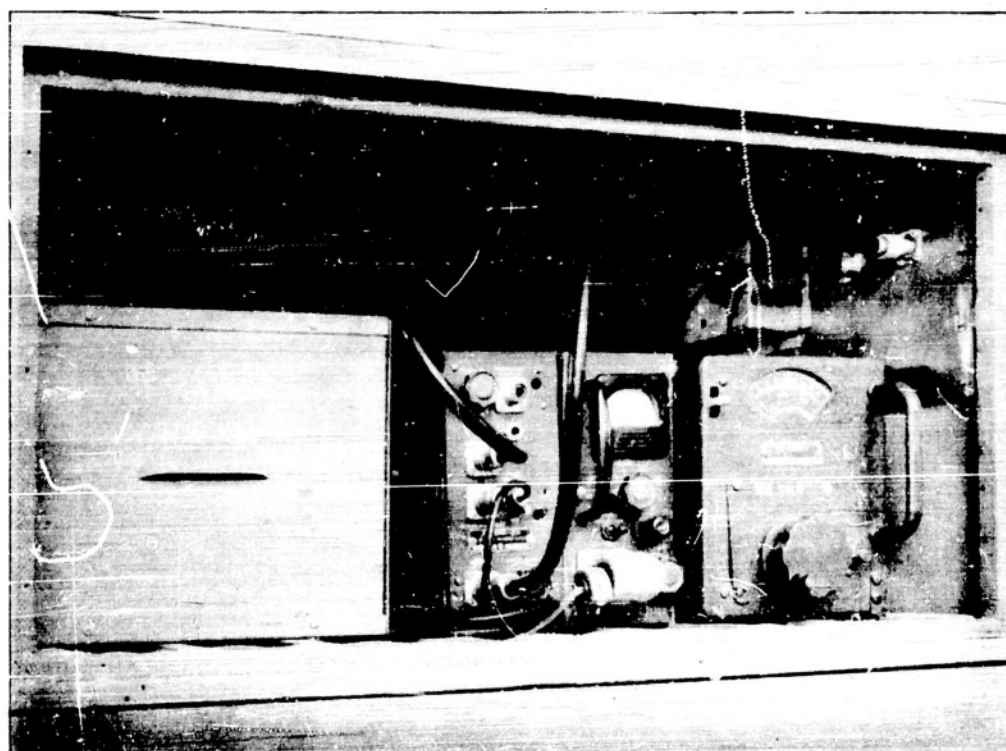
ator padded by 20 db of lossy cable was used as the signal source with tuning stubs to match this source to the various loads. The arrangements tried were a crystal detector with tuned audio amplifier, a superheterodyne receiver (the military APR-1), and the same receiver with a tuned amplifier following it. The signal-to-noise ratio decreased in the order listed. The last set-up was 8 db better than the crystal-tuned amplifier arrangements and 4 db better than the receiver by itself. For this reason it was decided to use a receiver-tuned-amplifier combination.

It is possible to modulate the radio frequency signal at a number of points before the final detection device. Difficulty was immediately experienced on a short-circuited line using a modulated power oscillator as a signal source and either a crystal-tuned amplifier-combination or a receiver as a detector. A peculiar probe-output versus piston-position was obtained as the line was tuned through resonance. There was the usual main peak, but on one side there were also two other peaks 7.5 and 18 db below the main peak. Examination of the waveform coming from the crystal detector showed that the two dips between the three peaks were caused by the fact that the usual 1000 cps crystal output had become distorted to primarily 2000 cps at these points. The same distortion occurred for any modulation frequency. Examination of the transmitter's spectrum revealed that it was several MC wide owing to frequency modulation caused by the direct modulation of the oscillator plate voltage. It was decided that this was the reason for the peculiar line-resonance output and that it would be necessary to modulate at some other point in the transmitter-detector path. One possibility is a silicon crystal in shunt or in series with the signal path at some low power point. Locating it in the receiver seemed reasonable and this was the final solution.

The receiver is a considerably modified SPR-2A military superheterodyne receiver with a TN-3/APR-1 tuning head. Its bandwidth has been reduced to about 5 percent of its original bandwidth by installing higher value plate-loading resistors in the plate circuits of the I. F. stages. This decrease in bandwidths is accompanied by a proportional increase in gain and a maintenance of the output noise level. Any further increase in gain would cause the strip to oscillate with its present mechanical arrangement. The hum level was reduced by more filtering of the power supply. Also a direct-current filament supply



(a) RECEIVER CHASSIS



(b) RECEIVER IN SHIELDED CABINET

FIGURE 17

was added for the second detector and the tuned audio-amplifier stages. An iron shield was needed around the power transformer. The most important aid to hum reduction was the addition of the tuned amplifier

The addition of a signal-path modulation circuit to the receiver may be achieved in many ways. Various points in the I. F. strip were tried but they had the common fault that the usual noise present was modulated as well as the signal. This increases the signal-to-noise ratio owing to the presence of an audio signal in the output with no signal whatsoever in the input of the receiver. The place for such modulation is ahead of as many of the noise sources as possible. It was decided to locate the circuit between the 1st mixer and the 1st I. F. stage. There the only noise which may be modulated is that arising in the 1st mixer and coming from the local oscillator. The circuit used is a series crystal in the signal path and has about 100 percent modulation. The incidental noise modulation is undesirable, but it may be suppressed.

The audio signal resulting from the incidental noise modulation is 180° out of phase from the audio due to an input radio frequency signal. Hence, as this input signal is increased from zero, the output audio drops first to a null at the point where the noise audio output equals the input signal audio output and then increases. The output due to incidental noise modulation is about 14 db above the receiver's random noise level. For this to be negligible an input signal to the receiver would have to be sufficiently large that this could be subtracted from it and not noticeably change its value. At lower levels there would be an increasingly great error in the measurement of the radio frequency input level.

A circuit was devised to suppress this noise modulation and consists in changing the gain of one of the I. F. stages at the proper phase and amplitude to exactly cancel it out. After this stage there is no noise modulation and the audio output in the absence of a signal has no non-random components. The output audio then increases directly from the noise level as the input signal increases. Note that this noise-modulation-suppression circuit tends to increase the amplitude of the desired audio output since the gain modulation is 180° out of phase with the noise modulation and hence in phase with the signal modulation.

The output of the second detector is noise and hum in addition to the desired 1000 cps audio voltage. Two tuned audio stages were added in an attempt to eliminate as much as possible of the undesired components and so improve the overall signal to noise ratio of the receiver. The first narrow-band stage is an ordinary amplifier whose plate load is a high-Q parallel-resonant circuit. The second stage is a cathode follower, the output of which is a ten thousand ohm resistor driven through a high-Q series resonant circuit. The receiver output is taken from across this resistor. It was adjusted to give the narrowest bandwidth which is about 100 cps. The cathode follower is biased quite positively in order that it operate with 50 volts of audio on its grid.

The transmitter power available was sufficient so that leakage directly from it and the coaxial cables and their connectors resulted in a large radio-frequency-signal level in the vicinity of the experimental setup. The receiver is sufficiently sensitive and the room level was sufficiently high that variations in the measured signals were caused by people moving around in the room. This was traced to imperfect shielding of the transmitter and receiver. A small shielded cage was constructed for the receiver and its modulating oscillator and this eliminated the difficulty. A radio-frequency low-pass filter was placed in the input power leads to the cage to eliminate pickup on them.

The 60 cps power available had sufficient fluctuations to give a noticeable effects on the receiver output. This was reduced considerably by using a regulating line transformer for the receiver and its audio oscillator.

9. Phase Measurement

For the measurement of phase a reference source of adjustable phase is required together with a device in which to compare the unknown and the reference. The comparison is made by adding the two voltages and finding the phase of the reference when a minimum (resulting from 180° phase difference) is attained. An increasingly sharp null is observed at this point when these voltages are approximately equal.

The reference signal used here is of adjustable phase, but since its phase is not independent of its amplitude, the magnitude of the reference signal cannot be changed during a run. The source is a coaxial line terminated ideally

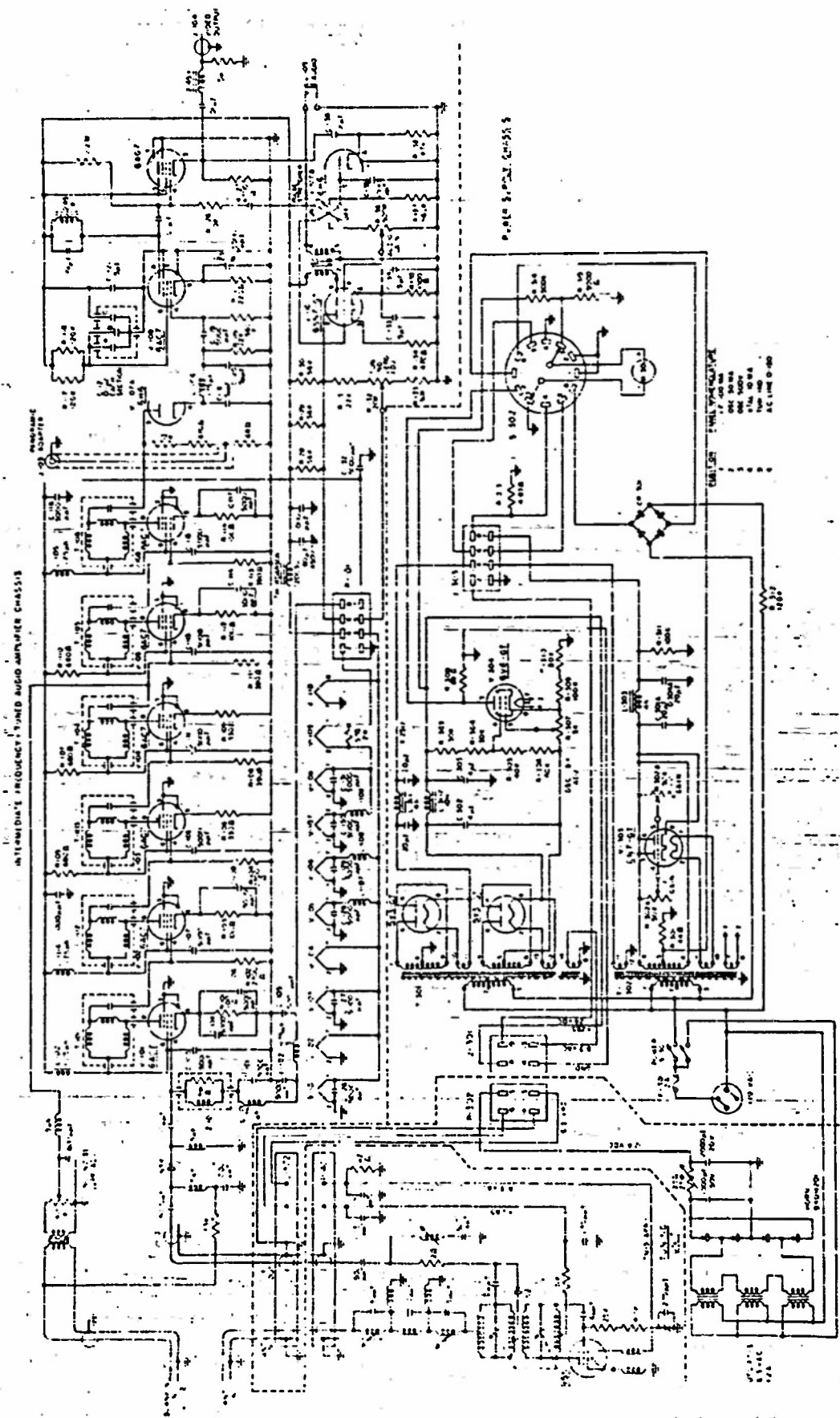


FIG 10 RECEIVER SCHEMATIC DIAGRAM

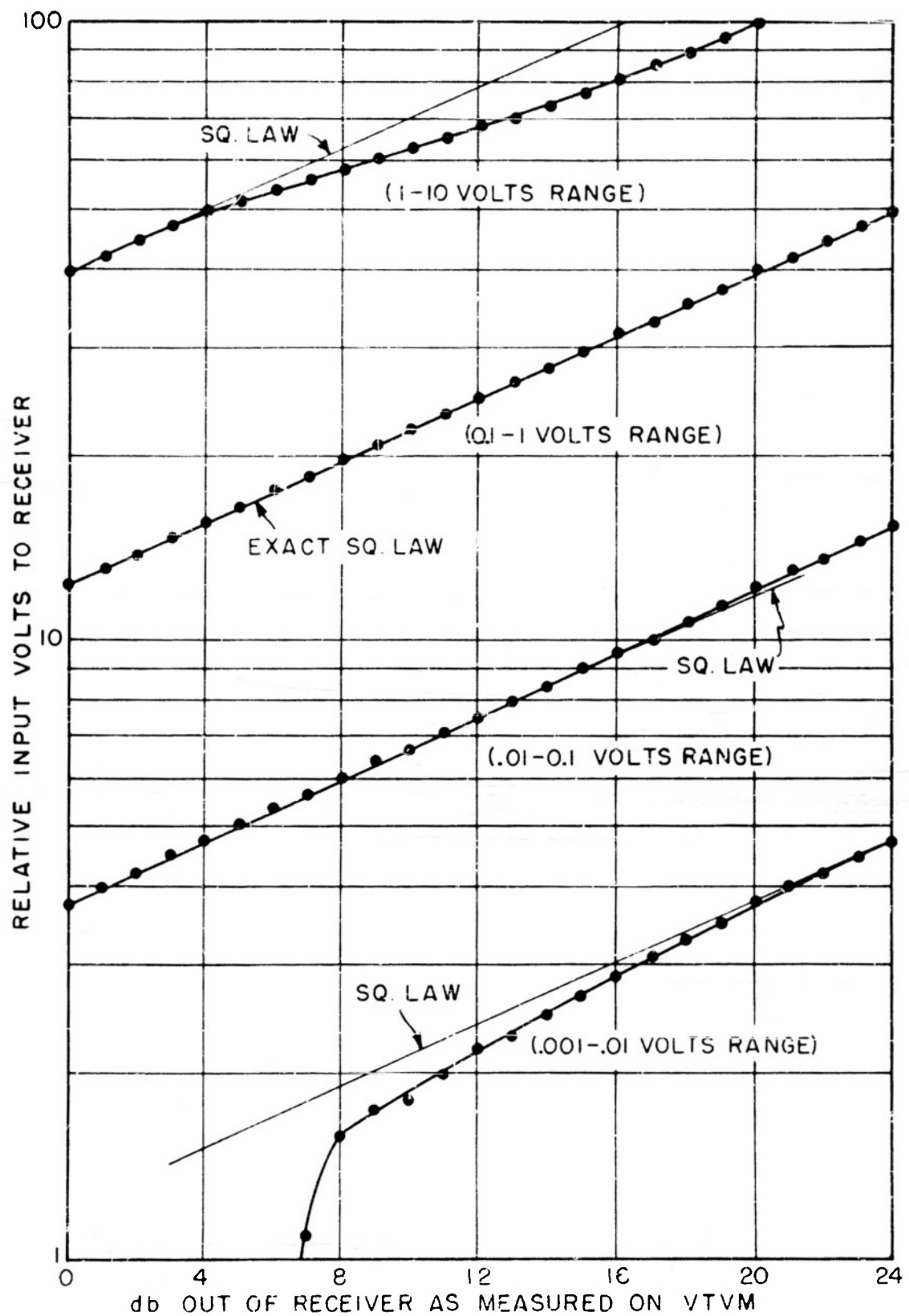


FIG 19 RECEIVER CALIBRATION

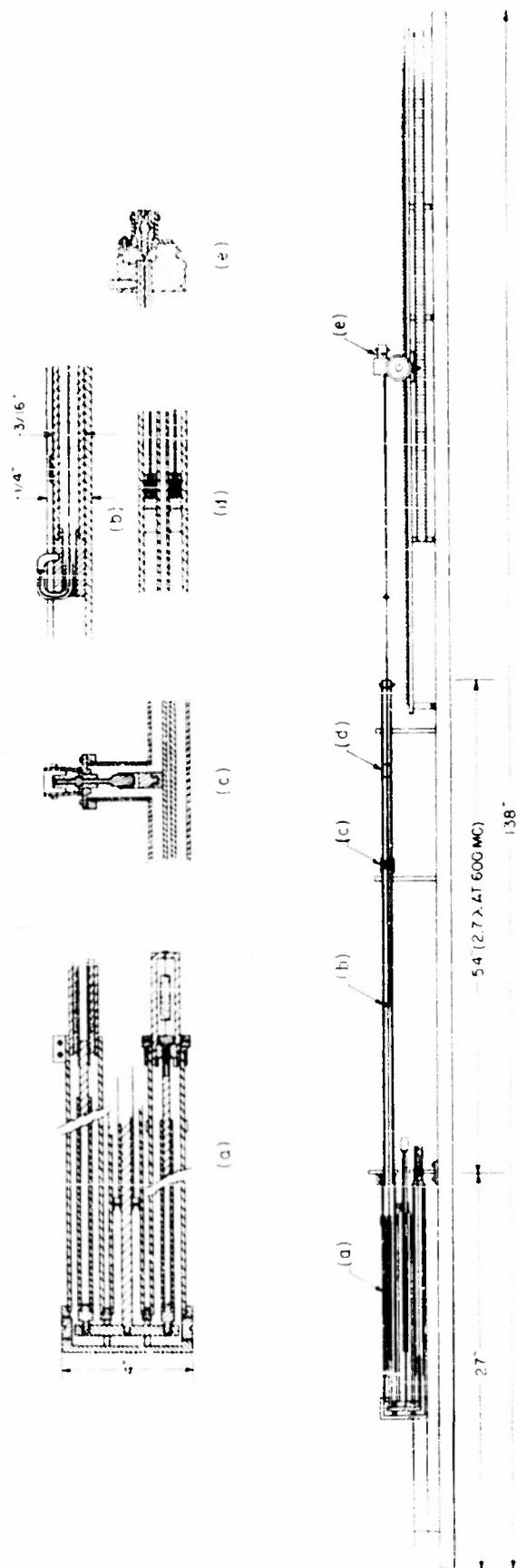


FIG. 20 PHASE LINE

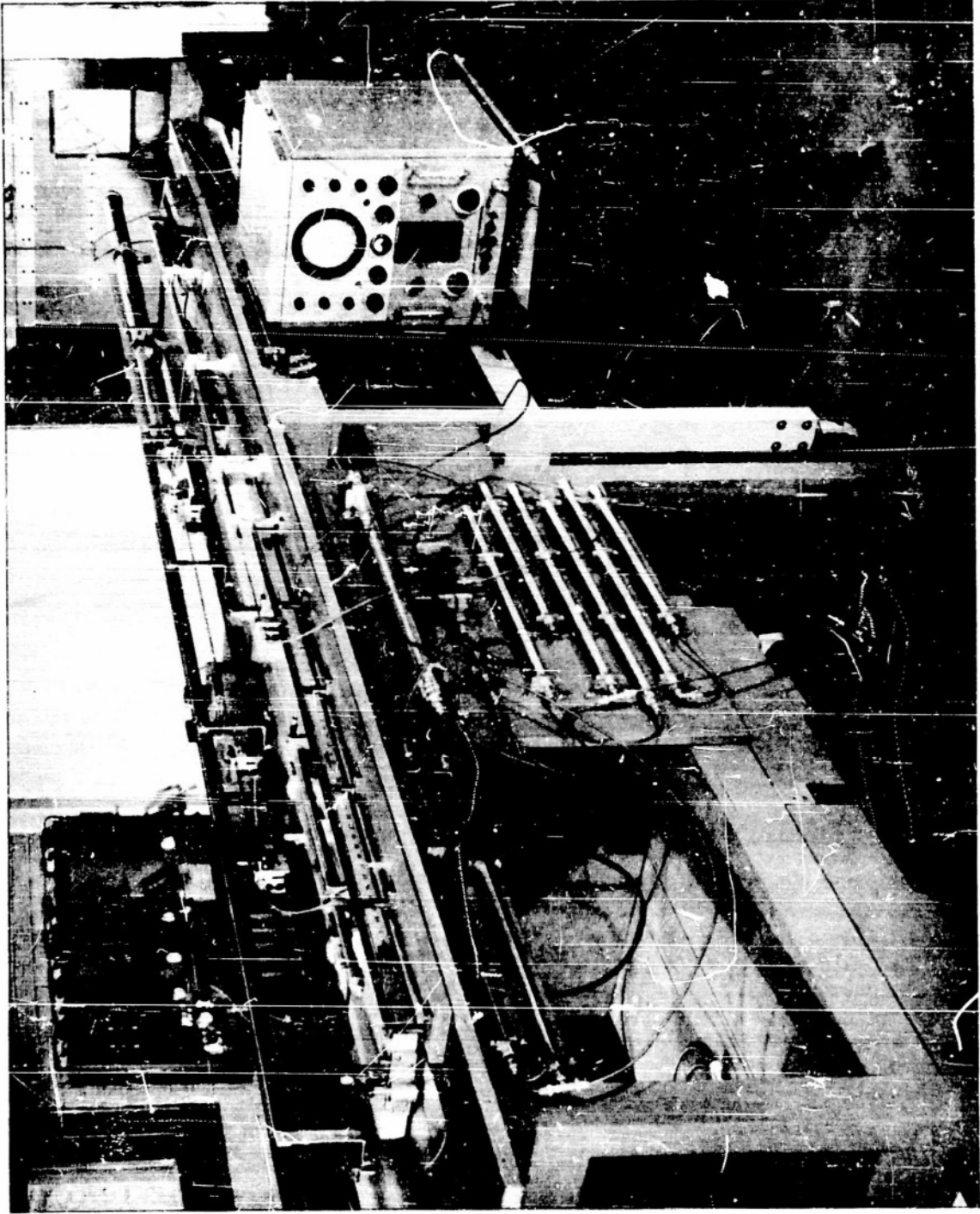


FIG. 21 BACK OF MEASURING LINE AND JUNCTION BOARD

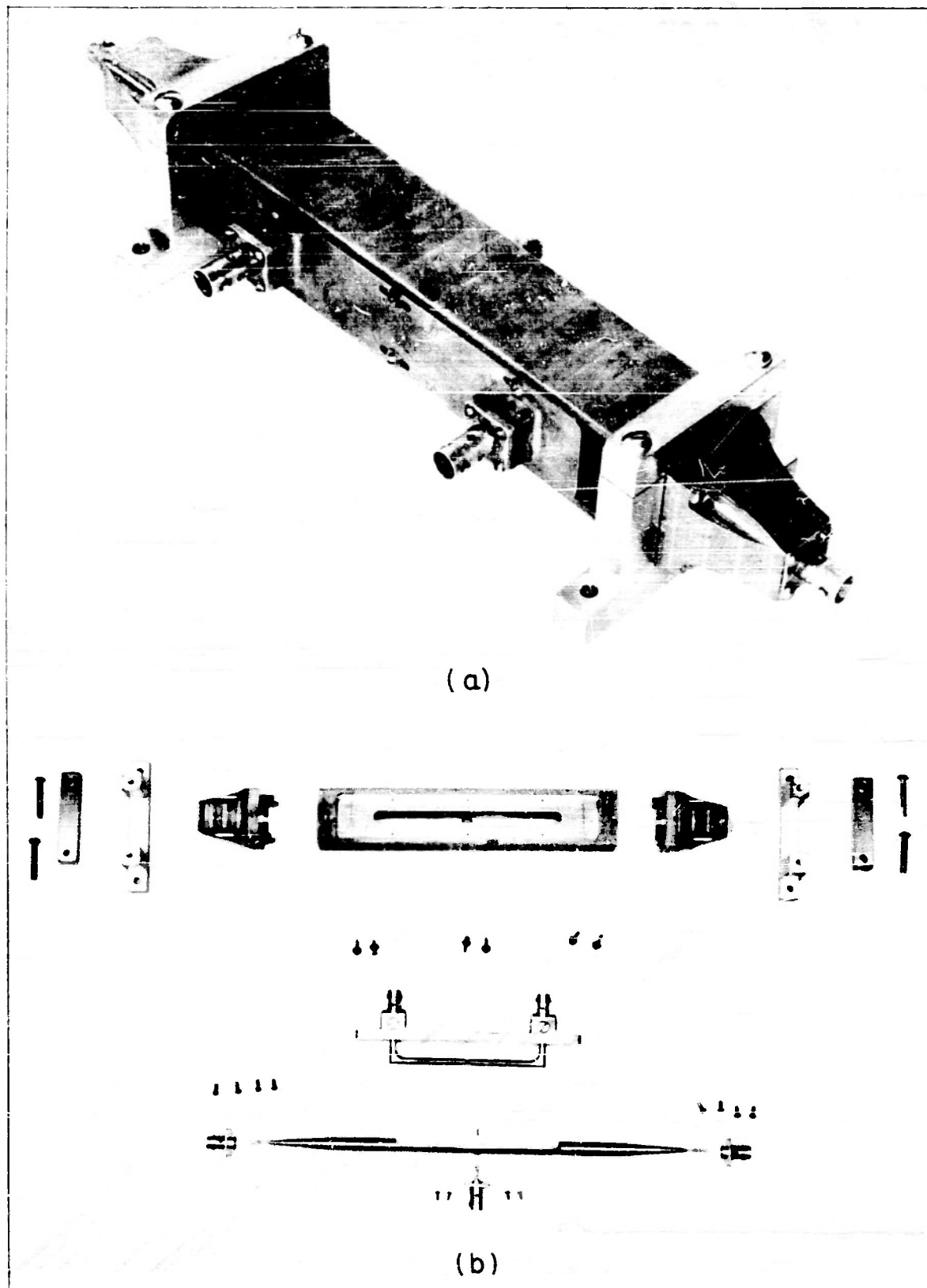


FIG. 22 HYBRID JUNCTION

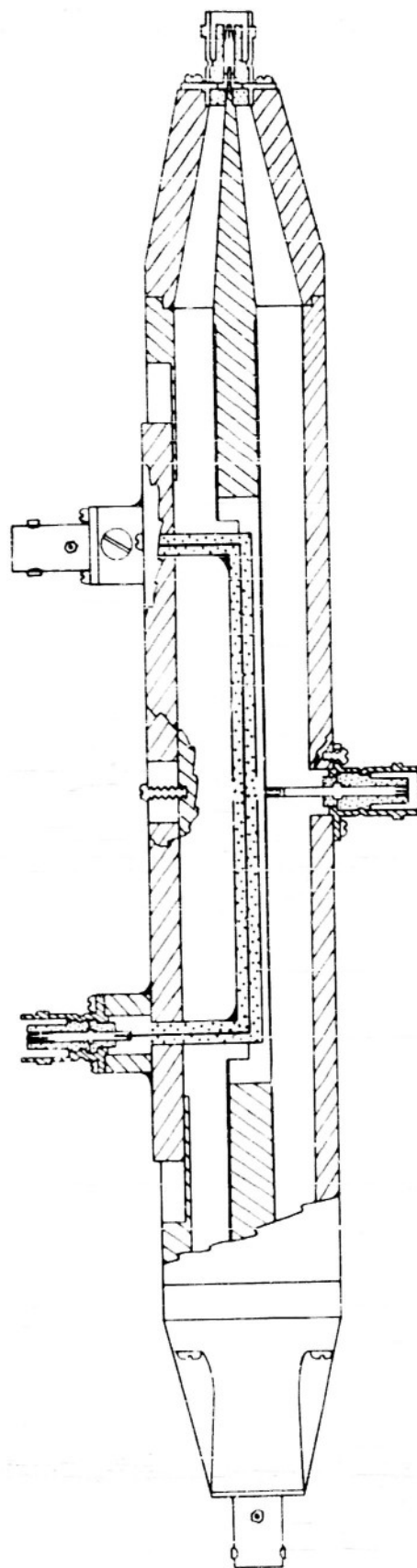


FIG.23 COAXIAL HYBRID JUNCTION

in its characteristic impedance so that it has no standing waves and hence a phase delay that is linearly proportional to distance down the line. The phase is known precisely except for variations in the velocity of propagation and a small residual standing-wave ratio. The load is obtained by tuning out the reactance of a resistive termination.

The construction of the phase line is similar to that of the main coaxial line. It is built of two concentric brass tubes, the center one being continuously supported by polyfoam and slotted over most of its length. A shielded loop probe protrudes through this slot into the line. At the end is a tuning piston for adjusting the line to resonance and thus making amplitude changes. The power feed to the line is by a direct contact to the center conductor. The termination on the line is a two-watt carbon resistor. The tuning configuration used requires a rather large resistance; one several times the characteristic impedance of the line was found necessary. It was tuned up using the probe to measure the amplitude on the line and adjusting the stub and line stretcher until the amplitude was constant over the line length. A standing-wave ration of unity was not obtainable but actually was equal to 1.05. For this a phase error from linearity of about three degrees maximum exists.

A comparison device is needed to add the phase reference and the unknown, but without coupling between them or other variable loads that will effect their phase or amplitude. Such a device is available in the coaxial hybrid junction.⁷ It has two uncoupled input terminals and two output terminals. From one of the input terminals a signal appears at both output terminals with the same phase. This is the T connection from input to output. From the other input the signals appearing at the two output terminals are 180° out of phase. This is the loop connection from input to output. Hence, one of the outputs has the sum of the input signals and the other has the difference of the signals. The device is connected with the reference signal on the T input and the unknown on the loop input. The receiver is connected to one output and a lossy load to the other. The phase is determined by adjusting the probe position for a minimum reading from the receiver. The amplitude is set at the beginning of a run to have the two signals as nearly equal as possible throughout the run.

10. Parasite Current Distribution

An auxiliary probe must be used to measure the current distribution on any parasite used with the driving antenna. Connections must be made to this probe, the supports for it will be large, and the probe will be used quite close to the currents being measured. In this manner the unavoidable possibility arises of large currents being excited upon the probe leads.

In an attempt to minimize these currents a series of detuning sleeves have been built onto the first wavelength of the coaxial conductor leading from the probe back to the measuring equipment.⁸ They consist of a group of four quarter-wavelength coaxial lines the inside conductors of which are the outer conductors of the coaxial lead of the probe. They are filled with polystyrene and short-circuited at one end. The open ends of the sleeves then present a high impedance in series with currents on the outside surface. The electrical length in the polystyrene is $1/4$ wavelength so that the exterior length is less than $1/6$ wavelength. Thus, the distances between high impedance points are not multiples of $1/4$ wavelength and it is impossible for current resonances to occur on the sleeve's surface. In addition to this preventive measure, the probe leads should be maintained perpendicular to the antenna to reduce coupling, and should be led away as close to the image screen as possible. The probe itself is a tiny shielded loop quite similar to the probe used in the main coaxial line.

11. Coaxial Switch

As seen in Fig. 1 two completely different circuits are needed for the measurement of phase and of amplitude. The receiver must be switched from the amplitude position where it measures the main line probe output to the phase position where it measures the output of the hybrid junction.

A shielded wafer switch was first tried but it had too much distributed capacitance between contacts. There was only about 20 db insertion loss when the switch was supposedly an open circuit. An almost completely continuous coaxial design was devised. Its insertion loss is very high and is limited only by leakage between the connections themselves. See Fig. 25b for a picture and cross section of the switch. It consists of an outer cup-

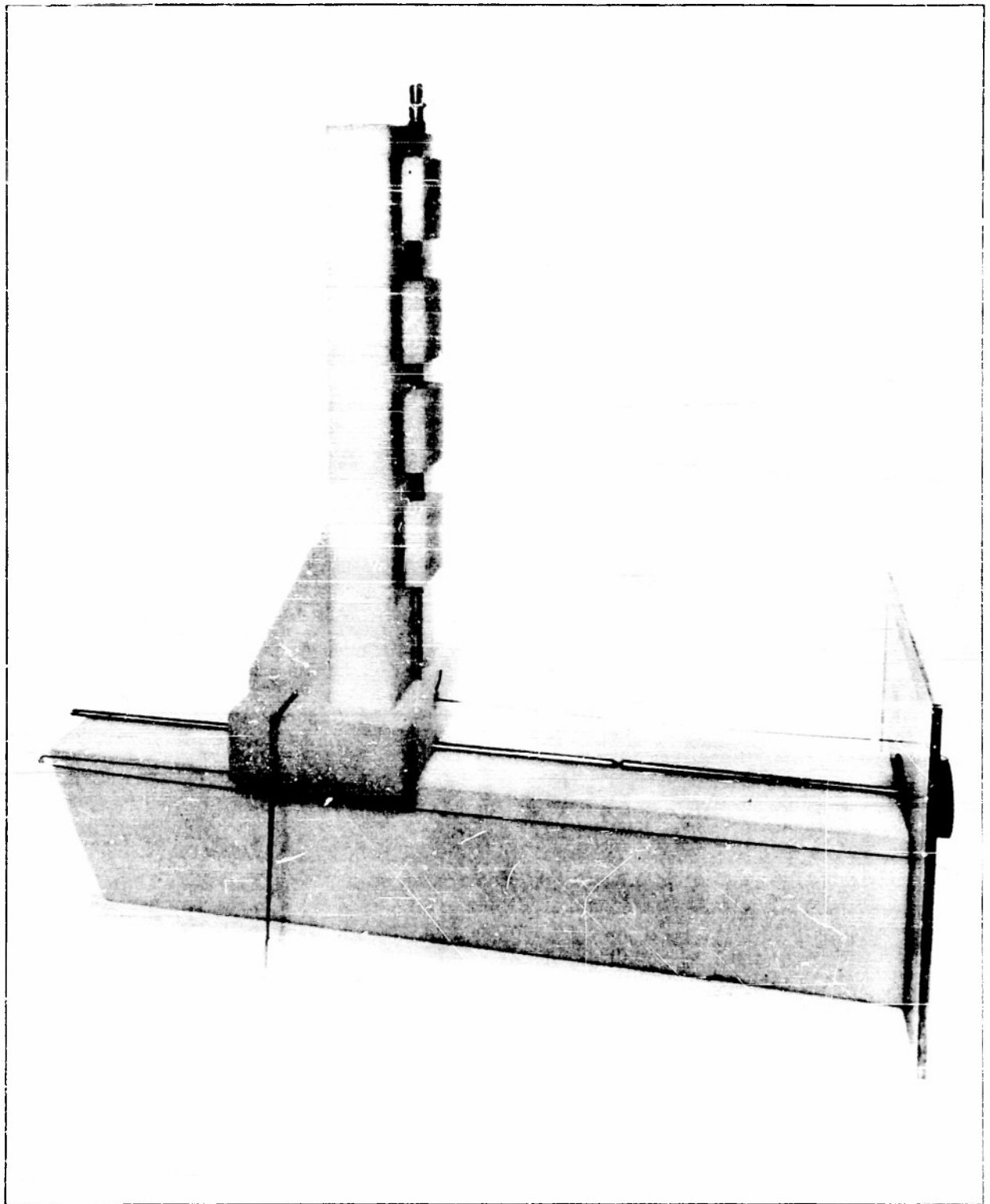


FIG. 24 TRAVELING PROBE AND ANTENNA SUPPORT.

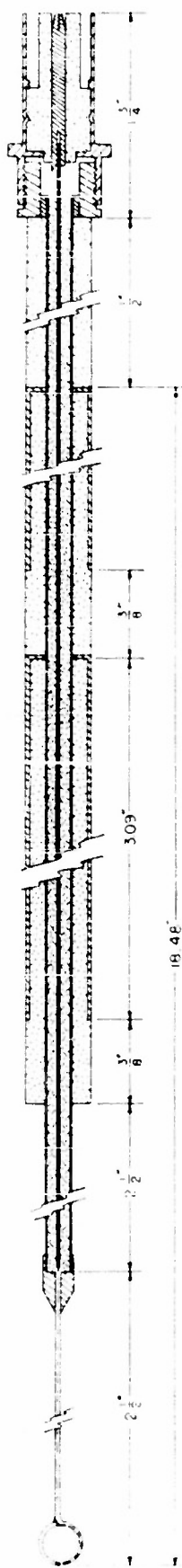


FIG 25a AUXILIARY PROBE

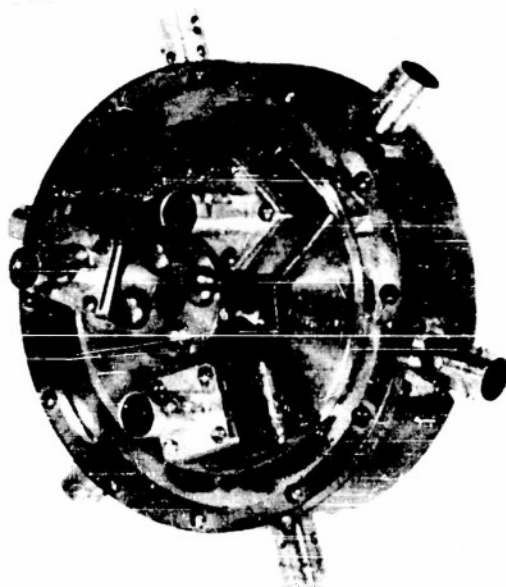
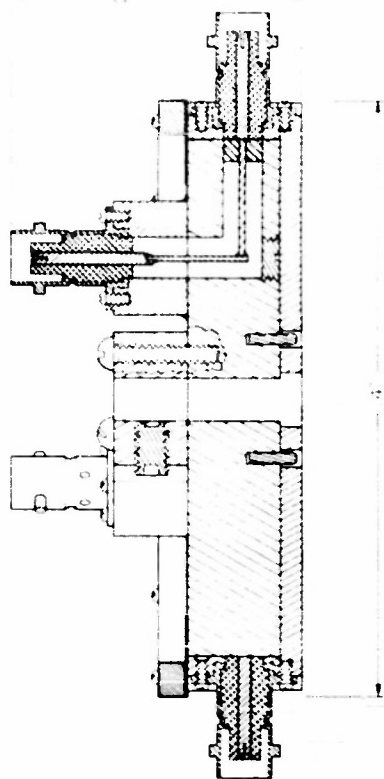


FIG 25b COAXIAL SWITCH



shaped part with the six stationary contacts and an inner rotating disk with the three moving contacts. The two parts are held tightly together with a thin circular phosphor-bronze spring. Note that the moving contacts are shorted to the outer part of the switch as they move between the fixed contacts.

12. Cables and Connectors

The quick disconnect BNC series of coaxial connectors are used as much as possible throughout the equipment. The type N series has more desirable leakage characteristics, but at the frequency used, the improvement is not sufficient to justify their use in view of their greater bulk.

The small-size double-shielded 50 ohm RG-55/U flexible coaxial cable is used to connect the various components in the assembly. A length of about 35 feet (12 db. at 600 MC) of lossy RG-21/U cable is used to isolate the transmitter from the various line-tuning devices and to locate it some distance from the measuring equipment.

References

1. J. E. Storer, "The Impedance of an Antenna over a Large Circular Ground Screen," Cruft Laboratory Technical Report No. 119, Harvard University, November 30, 1950.
2. T. Morita, "The Measurement of Current and Charge Distributions on Cylindrical Antennas," Proc. I.R.E. 38, 898 (1950).
3. W. H. Huggins, "Broad Band Non-Contacting Short Circuit for Coaxial Lines," Proc. I.R.E. 35, 906-913 (September 1947).
4. T. Morita, "The Measurement of Current and Charge Distributions on Cylindrical Antennas," Cruft Laboratory Technical Report No. 66, Harvard University, February 1, 1949.
5. H. E. Sorrows, W. E. Ryan, R. C. Ellenwood, "Evaluation of Coaxial Slotted-Line Impedance Measurements," Proc. I.R.E. 39, 162-168 (February 1951).
6. Staff of Radio Research Lab., Harvard University, Very High Frequency Techniques, Vol. 2, Chap. 27, McGraw-Hill, 1947.
7. T. Morita and L. S. Sheingold, "A Coaxial Magic T," Cruft Laboratory Technical Report No. 162, Harvard University, October 10, 1952.
8. R. M. Hatch, Jr., "Current Distribution on Conducting Sheets Excited by Arrays of Slot Antennas," Cruft Laboratory Technical Report No. 103, Harvard University, July 5, 1950.

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- 6 D. D. King, "Impedance Measurements on Transmission Lines," 1946. Proc. I.R.E., May 1947.
- 8 B. C. Dunn, Jr. and R. W. P. King, "Currents Excited on a Conducting Plane. . .," 1947. Proc. I.R.E., Feb. 1948.
- 11 D. D. King et al, "Bolometer Amplifier for Minimum Signals," 1947. Electronics, Feb. 1948.
- 12 C. T. Tai, "Theory of Coupled Antennas," 1947. Part I Proc. I.R.E., April 1948; Part II, ibid, Nov. 1948.
- 16 Tung Chang, "Impedance Measurements of Antennas Involving Loop and Linear Elements," 1947.
- 18 C. T. Tai, "Propagation of Electromagnetic Waves from a Dissipative Medium to a Perfect Dielectric," 1947.
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- 22 C. H. Papas and R. W. P. King, "Radiation Resistance of End-Fire Collinear Arrays," 1947. Proc. I.R.E., July 1948.
- 23 R. W. P. King, "Field of Dipole with Tuned Parasite at Constant Power," 1947. Proc. I.R.E., July 1948.
- 25 J. V. Granger, "Low-Frequency Aircraft Antennas," 1947.
- 27 C. H. Papas and R. W. P. King, "Surface Currents on a Conducting Plane. . .," 1948. J. Appl. Phys., Sept. 1948.
- 28 C. T. Tai, "Reflection and Refraction of a Plane Electromagnetic Wave. . .," 1948.
- 32 C. H. Papas and R. King, "Currents on the Surface of an Infinite Cylinder," 1948. Quart. Appl. Math., Jan. 1949.

- 35 P. Conley, "Impedance Measurements with Open-Wire Lines," 1948. J. Appl. Phys., Nov. 1949.
- 39 S. B. Cohn, "Theoretical and Experimental Study of a Waveguide Filter Structure," 1948
- 40 C. T. Tai, "Reflection of Plane Electromagnetic Waves from Perfectly Conducting Grounded Half-Cylinder," 1948.
- 41 R. W. P. King, "Theory of Antennas Driven from a Two-Wire Line," 1948. J. Appl. Phys., Sept. 1949.
- 42 J. V. Granger, "Note on Broad-Band Impedance Characteristics of Folded Dipoles," 1948.
- 43 D. G. Wilson and R. King, "Measurement of Antenna Impedance Using Receiving Antenna," 1948.
- 44 E. Hallén, "Properties of Long Antennas," 1948. J. Appl. Phys., Dec. 1948.
- 46 E. Hallén, "Admittance Diagrams for Antennas. . .," 1948.
- 47 C. T. Tai, "On the Theory of Biconical Antennas," 1948. J. Appl. Phys., Dec. 1948.
- 48 K. Tomiyasu, "Problems of Measurement on Two-Wire Lines with Application to Antenna Impedance," 1948. Condensed version, J. Appl. Phys., Oct. 1949.
- 49 E. Hallén, "Traveling Waves and Unsymmetrically Fed Antenna," 1948.
- 50 D. D. King, "Measurement and Interpretation of Antenna Scattering," 1948.
- 52 C. H. Papas and R. King, "Input Impedance of Wide-Angle Conical Antennas," 1948. Proc. I.R.E., Nov. 1949.
- 53 D. K. Reynolds, "Surface-Current and Charge Measurements on Flat Metal Sheets," 1948.
- 55 C. T. Tai, "Study of the EMF Method," 1948. J. Appl. Phys., July 1949.
- 56 T. W. Winternitz, "The Cylindrical Antenna Center-Driven by a Two-wire Open Transmission Line," 1948. Quart. Appl. Math., 1949.
- 58 C. H. Papas, "On the Infinitely Long Cylindrical Antenna," 1948. J. Appl. Phys., May 1949.

- 61 C. H. Papas, "Radiation from a Transverse Slot in an Infinite Cylinder," 1948. J. Math. and Phys., Jan. 1950.
- 63 J. V. Granger and N. G. Altman, "Full-Scale Aircraft Antenna Measurements," 1949.
- 66 T. Morita, "Measurement of Current and Charge Distributions on Cylindrical Antennas," 1949. Proc. I.R.E., Aug. 1950.
- 67 T. Morita and C. E. Faflick, "Measurement of Current Distributions along Coupled Antennas. . .," 1949.
- 69 J. E. Storer and R. King, "Radiation Resistance of a Two-Wire Line," 1949.
- 70 J. V. Granger, "Shunt-Excited Flat-Plate Antennas. . .," 1949. Proc. I.R.E., March 1950.
- 71 }
72 } B. C. Dunn, Jr., "Microwave Field Measurements," I (with
73 } R. King), II and III, 1949.
- 74 R. King and K. Tomiyasu, "Terminal Impedance and Generalized Two-Wire Line Theory," 1949. Proc. I.R.E., Oct. 1949.
- 75 C. T. Tai, "Application of a Variational Principle to the Study of Biconical Antennas," 1949.
- 76 C. H. Papas, "Radiation from a Circular Diffraction Antenna," 1949.
- 77 C. T. Tai, "On Radiation and Radiating Systems in the Presence of a Dissipative Medium," 1949.
- 78 J. V. Granger and T. Morita, "Current Distribution on Aircraft," 1949.
- 81 K. Tomiyasu, "Loading and Coupling Effects of Standing-Wave Detectors," 1949. Proc. I.R.E., Dec. 1949.
- 83 C. H. Papas, "Diffraction by a Cylindrical Obstacle," 1949. J. Appl. Phys., April 1950.
- 84 R. King, "Theory of N Coupled Parallel Antennas," 1949. J. Appl. Phys., Feb. 1950.
- 86 K. Tomiyasu, "Unbalanced Terminations on a Shielded-Pair Line," 1949.
- 91 R. King, "Theory of Collinear Antennas," 1949.

- 92 C. H. Papas and R. King, "Radiation from Wide-Angle Conical Antennas. . .," 1949. Proc. I.R.E., Nov. 1949.
- 93 R. King, "Asymmetrically Driven Antennas and the Sleeve Dipole," 1949.
- 94 T. Morita, E. O. Hartig, and R. King, "Measurement of Antenna Impedance. . .," (Supplement to T. R. 43), 1949.
- 95 C. P. Hsu, "Theory of Helical Waveguides and Helical Radiators," 1950.
- 96 R. King, "Theory of V-Antennas," 1950.
- 98 D. J. Angelakos, "Current and Charge Distributions on Antennas and Open-Wire Lines," 1950.
- 100 H. Levine and C. H. Papas, "Theory of the Circular Diffraction Antennas," 1950.
- 101 J. E. Storer, "Variational Solution to the Problem of the Symmetrical Cylindrical Antenna," 1950.
- 104 G. Wheeler, "Coupled Slot Antennas," October 25, 1950.
- 105 R. D. Kodis, "An Experimental Investigation of Microwave Diffraction," 1950.
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- 119 J. E. Storer, "The Impedance of an Antenna over a Large Circular Screen," 1950. J. Appl. Phys., August 1951.
- 121 R. King, "Theory of Collinear Antennas, II," 1950. J. Appl. Phys., December 1950.
- 122 J. Taylor and T. Morita, "Antenna Pattern-Measuring Range. 1951.
- 126 J. E. Storer, "The Radiation Pattern of an Antenna over a Circular Ground Screen," 1951.

- 128 J. Taylor, "The Sleeve Antenna," 1951.
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- 132 R. King, "An Improved Theory of the Receiving Antenna," June 1951.
- 134 T. E. Roberts, Jr., "Properties of a Single-Wire Line," 1951.
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- 139 T. E. Roberts, Jr., "An Experimental Investigation of the Single-Wire Transmission Line," 1952.
- 141 R. King, "Theory of Electrically Short Transmitting and Receiving Antennas," 1952.
- 146 C. Moritz, "The Coupled Receiving Antenna, I.," 1952.
- 147 C. Moritz, "The Coupled Receiving Antenna, II.," 1952.
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- 172 D. B. Brick, "The Radiation of a Hertzian Dipole over a Coated Conductor," May 10, 1950.
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