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AN INDUCTIVELY COUPLED GONIOMETER FOR WIDE-APERTURE DF ARRAYS

R. F. Gleason and C. V. Johnson, Jr.

RADIO DIVISION

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AN INDUCTIVELY COUPLED GONIOMETER FOR WIDE-APERTURE DF ARRAYS

by

R. F. Gleason
C. V. Johnson, Jr.

12 August 1958

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ABSTRACT

The directional patterns produced by circularly disposed antenna arrays have many technical uses. These various applications may require varying degrees of accuracy, different rotational rates and different coupling efficiencies to produce optimum results for each. For use where directional accuracy is secondary but coupling efficiency a major consideration, inductively coupled goniometers may prove more desirable. This report discusses the various design problems and describes an experimental hand-positioned goniometer using inductive coupling from antennas to the phasing lines.

PROBLEM STATUS

This is an interim report; work on this problem is continuing.

AUTHORIZATION

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AN INDUCTIVELY COUPLED GONIOMETER FOR WIDE-APERTURE DF ARRAYS

INTRODUCTION

Increased consideration is being given to the wide aperture system of radio direction finding (often referred to as the Wullenweber system) employing circularly disposed antenna arrays for purposes other than bona fide direction finding. Such a system, when employing a manually operated goniometer, possesses considerable potential merit as a means of producing a readily and rapidly steerable receiving beam for the reception of intelligence where very weak signals must be received through intentional or natural interferences.

The term "potential" merit is used above because, while the antenna arrays are inherently high gain and are capable of producing narrow beams with a minimum of back lobes, the overall effectiveness of such a system has, in the past, been compromised by the goniometer design and its accompanying antenna matching problems. Heretofore capacitively coupled goniometers have been used almost universally, which, while excellent for direction finding accuracy, possess excessively high insertion losses which tend to nullify the advantages gained by the array as a steerable beam for intelligence reception. Similarly, for this use, where direction finding accuracy can be made a secondary consideration, these goniometers are rather expensive to construct.

In an attempt to overcome these weaknesses in the conventional capacitively coupled goniometers, the Naval Research Laboratory has undertaken an investigation of the practicality of employing inductive coupling in such goniometers, simplifying their design and reducing their insertion loss to a point that would make the use of the antenna arrays really advantageous.

In this report, the various design problems involved are discussed and analyzed and a description, with illustrations, of an experimental inductively-coupled goniometer is given.

GENERAL CONSIDERATIONS

The desire for selective directional reception of radio waves from arbitrary azimuths has led to the consideration of methods to form and rotate a directional antenna beam. In the high frequency band, large circular arrays with suitable phasing networks have been used for this purpose.[1] They are easily adapted to different modes of operation simultaneously for monitoring and providing accurate df information of received signals. For use as direction finder the high-accuracy goniometer may rotate at high speeds (at least 300 revolutions per minute) resulting in a stationary pattern on the screen of a cathode-ray tube. This method is excellent for df purposes but for narrow beam-widths it modulates the incoming signals so that the received portion of the signal may be useless if information content is a main requisite. The use for monitoring purposes in particular requires a method capable of rotating the directional antenna beam so that it can be aimed continuously at the desired target. The use of suitable antenna multicouplers will provide the necessary outputs for simultaneously forming and rotating independent beams. Instrumental accuracy of the phasing device may be considerably reduced; however, it should have a higher coupling efficiency than tapped-delay line goniometers[1,2] now in use for high-accuracy beam positioning. The higher efficiency would insure that the phasing device and receiver were not the limiting factors in obtaining an intelligible signal. This factor may be relatively unimportant for some locations and frequencies, but particularly above 10 megacycles advantage may be taken of the receiving antenna's directivity, making it partially blind to the sources of atmospheric and man-made noise and more selectively responsive to the desired signal.

SOME METHODS FOR BEAM ROTATION WITH CIRCULARLY DISPOSED ARRAYS

Rotation of the antenna beam of an array can be achieved by suitable variations of the phases of the group of antennas to be combined at one time. Coupling methods considered most suitable for circularly disposed arrays in the hf band (3 - 30 Mc) are outlined briefly in the following discussion. These are capacitive, direct, or inductive coupling.

1. Capacitive Coupling

The use of capacitive coupling is a suitable method of forming and rotating beams for high accuracy df use. In this method capacitive coupling (Fig. 1) to a lumped constant delay line (with proper values of L and C) gives the appropriate phase relations to the received signal. By turning the rotor it is possible to continuously rotate the whole antenna pattern.

The insertion loss for this coupler is inherently high since the coupling capacities from the antenna elements must be less than the smallest C used in the delay line. This is necessary to reduce errors as a result of mismatch and variation in delay with changes in position of the rotor and stator plates. As a result of this large insertion loss high-gain antenna amplifiers may be required for monitoring low-level signals.

2. Direct Coupling

The use of direct coupling with brushes or switches is an alternate solution that could conceivably be used for hand rotation of the antenna beam by coupling the phasing network to successive groups of antennas in a circle. In the use of direct coupling, the beam can be steered only to the antenna positions (or some other initially prescribed position) with a single phasing network. For a 40-element array the beam would jump in 9-degree steps. A 120-degree element array would reduce this to 3-degree steps.

In the case of the 40-element array the 3-degree movement of beam position would require 2 additional sets of phasing networks or a method to vary the delays in a single set of delay lines. This method introduces additional switching problems or coupling losses.

3. Inductive Coupling

In this method of coupling, beam rotation is accomplished by the use of input transformers in which the secondary may be moved or varied in coupling to the primary winding. Figure 2 serves to illustrate the position where a particular set of delay lines is inductively coupled equally to two antennas with the result that the beam position will be half way between two antenna positions. In this manner the beam may be directed quite accurately to points in between the antenna positions if the coupling transformers are designed so that the secondary winding couples progressively from one primary transformer winding to the next about the complete circle. The blending in this device will not be as smooth as in the case of capacitive since the coupling using inductances with powdered iron or ferrite cores is not a smooth function of position as in the case of moving rectangular capacitor plates with a constant air gap. The so-called "cogging" or beam-positioning error for the device illustrated in Fig. 2 will be zero at points near 0, 4.5, 9, 13.5 --- degrees, with a maximum error at odd multiples of 2.25 degrees. The maximum error depends on the shape of the cores and degree of coupling but should not exceed 1.5 degrees. Lack of beam symmetry will be evident at the positions of maximum error.

The use of inductive rather than direct coupling to phasing lines can therefore offer several possible advantages. Among these are reduction of coupling loss, elimination of brush noise and the possibility of automatic or high-speed rotation. For these reasons an investigation of various techniques and designs for inductive coupling is in progress at the present time. These studies are aimed at the final design of a small instrument with low insertion loss and "cogging" error less than 2 degrees.

DESCRIPTION OF ANTENNA ARRAY AND PHASING LINES

The antenna array with which the experimental goniometers are to be used is a 430-foot circular array of 40 antennas with a complete reflecting screen behind the elements. The nominal frequency range of the array is 4 to 25 megacycles. Twelve adjacent antennas are used to form a beam at one time.

Figure 3a shows schematically the phase delays required to form a beam along a center line of 12 antennas. The lengths of delay shown are required to bring all antenna outputs in phase along a common baseline A - A' for signals arriving perpendicular to this line and at a 20-degree vertical angle of arrival.

Figure 3b shows schematically the delay lines required to correctly position the beam at 1.5 degree intervals (compared with 4.5 degrees for Fig. 3a) for the same array. For this method of phasing two delay lines may be completely uncoupled from the input transformers so it is necessary to terminate all lines with loads equal to their characteristic impedance. This will result in an additional insertion loss of approximately 6 db since a portion of the power is now absorbed by the terminations and the termination is one-third of the cable characteristic impedance. A previous goniometer or scanner has been constructed in this manner (Figs. 4 and 5). The present scanner is of the type shown in Fig. 3a where all lines are used continuously.

ARRANGEMENT OF GONIOMETER COMPONENTS

The three component parts required for the goniometer are

1. input transformers
2. delay lines
3. output transformers

Figure 2 shows schematically the arrangement of the three components. The input or coupling transformers between antenna inputs and delay lines step the 50-ohm input impedance up to 110 ohms for proper matching with the delay lines to be used. In order to reduce the lengths of several delay lines a modification of the single delay line arrangement (for each antenna) is made as shown by Fig. 6. The delay lines are now arranged so that a combination of 110- and 50-ohm delay lines are used. This is done by delaying the output of antenna 1 by use of a 110-ohm delay line so it is in phase with the output of antenna 2. The junction of the 110-ohm line and number 2 transformer output (110-ohm impedance) is then delayed through a 55-ohm cable by the amount required to phase these two antennas with the other antennas using a common base line. Similarly, the output of antenna 3 is delayed until it is in phase with antenna 4 and a section of delay line of half their impedance (55 ohms) is used to delay this output by an amount until it is in phase with the output of antennas 1 and 2. A similar procedure is used for outputs of antennas 5 and 6. The outputs of these three 55-ohm cables when joined will result in a maximum response for a signal arriving along the center line (and vertical co-phasal angle) of the group of antennas to be phased.

The impedance seen at this output junction will be approximately 18 ohms so , if combined directly with the other half of the array this would result in an impedance of 9 ohms. At this low value of impedance, lead lengths become comparable to that value at the higher frequencies, making the transformer design more difficult. For this reason the two 18-ohm outputs are first transformed to an impedance of 100 ohns before combining.

INPUT TRANSFORMER DESIGN

The design and performance of broad band transformers have been described in considerable detail in the literature[3, 4]. Generally they are physically different from this transformer where the two windings are on individual cores separated by an air gap. In the particular case of auto-transformers the importance of the core material is limited to the lower frequencies where it results in a decrease of the number of turns required and so reduces the capacity effects at the higher frequencies. The use of high-permeability ferrite cores along with spiral tape windings makes it relatively easy to design auto-transformers with losses less than 1 db, and extending beyond 50 megacycles in frequency. Similarly, balanced to unbalanced transformers are fairly easily constructed by use of a single layer wound secondary spaced above a spiral tape primary and again using high-permeability cores good at the lower frequencies.

For the transformers needed in this type of construction the core characteristics are important over the entire frequency range. The type of material desired should have a high permeability that would remain constant beyond 30 megacycles. In addition, cores loss such as hysteresis, residual, and eddy current losses must remain low over the entire frequency range.

The required air gap for moving the secondary cores with respect to the primary has the effect of reducing the permeability; however, it is possible to reduce the high-frequency losses by variation of this air gap.[5, 6] The selection of a ferrite material for this application therefore depends on the air gap dimensions in addition to variation of losses and permeability with frequency for a particular ferrite.

The core material available in quantity at the time of construction was a Manganese zinc ferrite with a high initial permeability (Ferroxcube 3A) but with losses increasing rapidly at frequencies above 5 megacycles. Figure 7 shows the loss from 4 to 30 megacycles for a 1 to 1 turns ratio transformer for this and several other core materials. The size and shape of the cores varied, so that some of the differences may be the result of these factors; however, they should be of secondary order compared to losses in the material. The results of this comparison demonstrate that a lower-loss transformer can be constructed by a better choice of core material. Based on the present material available the Ferroxcube 4B core should give better results at the higher frequencies and about equal results for the lowest frequency.

In all of the transformers spiral tape multi-layer windings approximately 1/8 inch wide (using .001 inch copper foil insulated with .001 mylar tape) were found to give the best results. This is attributed to a lowered value of leakage inductance as compared to a single layer winding.

Since it does not appear possible to couple without some loss in the input transformer, a simple network matching the secondary of the transformers to the delay lines may be required. This will introduce a slight additional loss but result in a more uniform response over the entire frequency range.

INPUT TRANSFORMER CONSTRUCTION

The input transformers, constructed using the Ferroxcube 3A cores, are shown in Fig. 8. These have spiral tape windings on ferrite E-shaped cores. They are designed to work between 50- and 110-ohm unbalanced impedances over a frequency range from 3 to 30 megacycles. The E cores are placed in a circular ring 6.5 inches in diameter (Fig. 9), requiring a tapered form to permit close spacing of the cores for continuous coupling. The 40 antenna input leads enter the bottom of the housing and are connected to the primary winding by short leads. A secondary with a similarly constructed spiral tape winding uses identical ferrite E cores. The E cores are mounted in a circular ring (Figs. 10 and 11) with a radius and angular spacing equal to that of the input transformer cores. The 12 secondary cores shown can then rotate with respect to the fixed cores so that the secondary windings couple to 12 of the primary windings in turn about the circle. Adjustment of the air gap for rotation or coupling variation is made by the screw and locking arrangement supporting the lower bearing (Fig. 12). An air gap of .005 inch is feasible with normal machining tolerances for this small rotor plate.

DELAY LINE DESIGN

In the selection of a delay line for this application the factors of size, shape, in addition to the major consideration of electrical characteristics must be considered. The desire for size reduction resulted in limited space which determines the combined physical length of all the lines that are necessary for beam formation. For certain delay line designs it may require crowding or placement of components so that phase distortion would be increased, and perhaps eliminating their choice.

The electrical properties to be considered are the delay linearity and amplitude characteristic with increasing frequency. In order to limit excessive phase variations the cutoff frequencies of all lines must be similar and higher than 30 megacycles. Since coupling losses were made a major consideration, the insertion loss should be less than 2 db at 30 megacycles for the longest delay line required.

The impedance values previously given (Fig. 6) were necessary to limit the required transformer turns ratio of primary to secondary to less than 1 to 1.5. An increase in turns ratio would make the transformer design more difficult. This relatively low impedance value may be a factor in the selection of a particular type of delay line design.

Several possible choices exist in the design of delay lines that are commercially available or for which design data can be obtained. The two general types are:

1. lumped constant (L and C)
2. single solenoid layer winding

In the design of the lumped constant line the required number of sections can be determined by the product of L and C, since this is a low-pass filter. In order to ensure that all frequencies of the operating band lie within the pass band, LC has to fulfill the condition

$$\sqrt{LC} < \frac{1}{\pi f_o}$$

where f_o is the highest frequency in the operating band. The bandwidth requirement of 30 megacycles would require a minimum of 10 sections for the longest line. The characteristic of this line can be modified by mutual coupling so careful design is required for the higher frequencies.

The solenoidal wound line is an attempt to replace the conductors of a conventional transmission line by a coaxial solenoid. One method employed quite successfully uses a close-spaced layer winding on a cylindrical conductor such as a silver plated ceramic rod. This conductor is usually slit longitudinally to prevent excessive attenuation. Still others are wound on a core of flexible dielectric and enclosed in a concentric shield insulated from the solenoid.

The time delay of both types shows variations with increasing frequencies. To reduce this distortion, variations of mutual coupling with frequency must be compensated by some means, such as coil placement for the lumped constant line or bridging capacities across portions of the lumped inductance, or a portion of the solenoidal wound coil.

Despite these corrective measures most commercially constructed lines available displayed a lack of phase linearity above 10 megacycles and also variations of impedance and length for similar lines of at least 10 percent.

In seeking a method which could better meet these particular requirements and also be reproduced in quantity without variations, the use of printed circuit techniques was considered. Printed circuit techniques to make strip transmission line have been used in the design of microwave components with considerable success.[7] Several variations of printed circuit lines are currently used at the higher frequencies. Perhaps the most widely used type is microstrip or "Strip line". Calculations of attenuation with frequency demonstrated that the loss for a 70-ohm stripline in the high-frequency region from 3 to 30 megacycles would be equal to that in an RG 59/U cable. This line therefore appeared to offer the possibility of a delay line whose impedance and length could be varied with relative ease. The greatest advantage would be linearity of phase with frequency beyond 30 megacycles.

The theory of strip transmission line has been covered in detail in numerous articles. The dominant mode of propagation in the "flat-strip" line, like the coaxial line, is the TEM mode. For wide strip line the field is concentrated in the region of the strip so that there is very little fringing effect. As the strip is narrowed, for higher impedances, the fringing effects become apparent due to interaction between fringing fields at the edges with fields of adjacent conductors. Theoretical curves of impedances versus width to dielectric thickness (Fig. 13) have been calculated[8] and experimentally checked for lines of many types.

DELAY LINE CONSTRUCTION

For this application, or any requirement where the length of delay is in the order of .1 microsecond (100 feet in air), it would be necessary to closely space the "strip line" to achieve an appreciable delay in a small area. A variety of configurations were tried but the best results were achieved for lines in which the conductors were continuous straight strips with equal spacings. These cross back and forth over the material (Figs. 14 and 15). Points on the impedance curve were checked with these designs and relatively little departure existed between the measured and calculated curve for lines below 125 ohms. Nearly all of the experimental designs were made with an epoxy resin impregnated fiberglass material (NEMA G-10). In order to achieve the delays required in the desired space, line widths and spacings of 1/32 inch were used. With this spacing an area of 1.4 square feet (using 1/32 inch double sided copper clad material) gave the equivalent of 100 feet of RG 59/U cable. This seventy-ohm line has a slightly higher loss (Fig. 16) than RG 59/U cable. The phasing lines required (for this scanner) with their lengths and impedance are shown in Fig. 6. The final lines were printed on circular shaped disks (Fig. 15) so that the entire phasing lines could be stacked above each other on the upper side of the rotor plate. The input terminals are placed so that only short vertical leads are required to make all of the necessary connections. Spacings between the circular plates (Fig. 11) are kept uniform by placing circular plastic disks around the outer edge and between adjacent disks.

Stacking the "strip line" as shown resulted in a slight change of impedance and length of line so that it is necessary to measure or calculate these two factors before the final circuits are printed. A change of length can be made easily by heating the "strip line" at the end and then pulling the required length from the plastic material.

Further reduction in losses may be accomplished by increasing the strip width and by using a lower dielectric constant insulating material. The use of a material of lower dielectric constant should result in a shorter electrical length. Losses due to the copper and dielectric material are given by:

conductor loss (α_{cu}) for copper in db/100 ft.

$$\alpha_{cu} = 7.25 \times 10^{-3} \frac{1}{h} (f_{mc} \quad e_r)^{1/2}$$

Dielectric loss (α_d) in db/100 ft.

$$\alpha_d = 2.78 \times f_{mc} \quad F_P \, (e_r)^{1/2}$$

where

h = dielectric thickness

F_p = power factor or loss angle

e_r = dielectric constant

By using the same fiberglass material with an epoxy resin and doubling the conductor width a reduction of 1.5 db can be realized at 30 megacycles. Comparative sizes of 70-ohm lines for the two widths can be obtained from Fig. 14, which shows a 110-foot line with one-sixteenth inch strip and a 65-foot line using one-thirty-second inch strip.

OUTPUT TRANSFORMERS

The completely assembled goniometer rotor is shown in Fig. 17, with the output terminals of the two groups of phasing lines on opposite sides of a center line. The output impedance of each network at this point is 18 ohms so that it requires a transformation to 100 ohms, resulting in a final desired output impedance of 50 ohms after combining all lines for the 12 antennas. The transformers used are broad-band auto-transformers. Their loss remains less than 1 db past 30 megacycles. For this application ferroxcube 4B core material gives a slightly better response at the higher frequency as compared with ferroxcube 3A.

ASSEMBLED GONIOMETER

The completely assembled goniometer is shown in Figs. 18 and 19. The two output terminals from the phasing network mounted on the rotating plate are brought out through a hole in the center of the rotor shaft to brushes mounted on the top cover plate. The instrument can be panel mounted as shown in Fig. 20. A 360-degree scale calibrated in one degree intervals is provided for positioning the beam or in obtaining bearings of signals.

TEST RESULTS

As a result of the inefficient input transformers installed at the present time, high standing wave ratios and losses are to be expected for frequencies above 10 megacycles. The results of measurements of input and output impedance are shown in Fig. 21. For both measurements all of the remaining input or output connections were terminated in the design impedance of 50 ohms.

Similarly the insertion loss increased with frequency in a manner corresponding to the input transformer loss. Figure 22 shows the measured insertion loss for the goniometer. The loss starts at 6 db at the lower frequency and rises to 14 db at 26 megacycles. In this measurement where a signal is applied to only one input it is necessary to take into consideration loss due to the inactive loads on the other input terminals. In order to determine or predict loss with improved transformers, measurements were made with a 1 to 1 turns ratio transformer (using Ferroxcube 4B material) fed through a 65-foot length of 50-ohm coaxial line and an additional attenuation of 3 db. The results of this measurement are also shown in Fig. 22.

Antenna array pattern measurements were made for this inductive coupled goniometer and also for the goniometer with 3 pickup coils for each antenna input coil. Patterns for this goniometer are shown in Fig. 23. These display evidences of "cogging" as shown by lack of pattern symmetry at 10, 12.5 and 15 megacycles. The patterns shown for 5, 10, and 15 megacycles are for signals arriving at one of the 2.25 degree positions, so they demonstrate the maximum amount of odd pattern distortion that can be expected.

Figures 24 and 25 show the patterns of the inductive coupled goniometer with additional delay lines for phasing the antenna outputs correctly at 1.5 degree intervals. This results in a much more symmetrical pattern, but displays some evidence of serrations (10 megacycles) as a result of the characteristics of transformer coupling as compared to capacitive coupling. Figure 25 shows null (Fig. 25a) and inverted null (Fig. 25b) patterns for this goniometer. These are used when the direction of arrival of a signal is to be determined.

CONCLUSIONS

The final design of an instrument suitable for the anticipated use requires a more efficient coupling transformer between the antenna inputs and phasing lines. The use of a ferrite material such as ferroxcube 4B should reduce transformer losses at the higher frequencies so that the loss at 30 megacycles would be improved, so that, it should be no greater than the present loss at 10 megacycles.

A small additional improvement can be made by use of a greater area for the rotor plate so that "strip line" with a wider conductor could be used. Reduction of loss for the longest line used will be nearly 1 db.

Radiation tests demonstrated suitable beam patterns; however, some improvement could be made by tapering or reducing the input of the side elements slightly. The taper would be such as to reduce signals from the outside element by about 3 db, and 0 db for the center elments.

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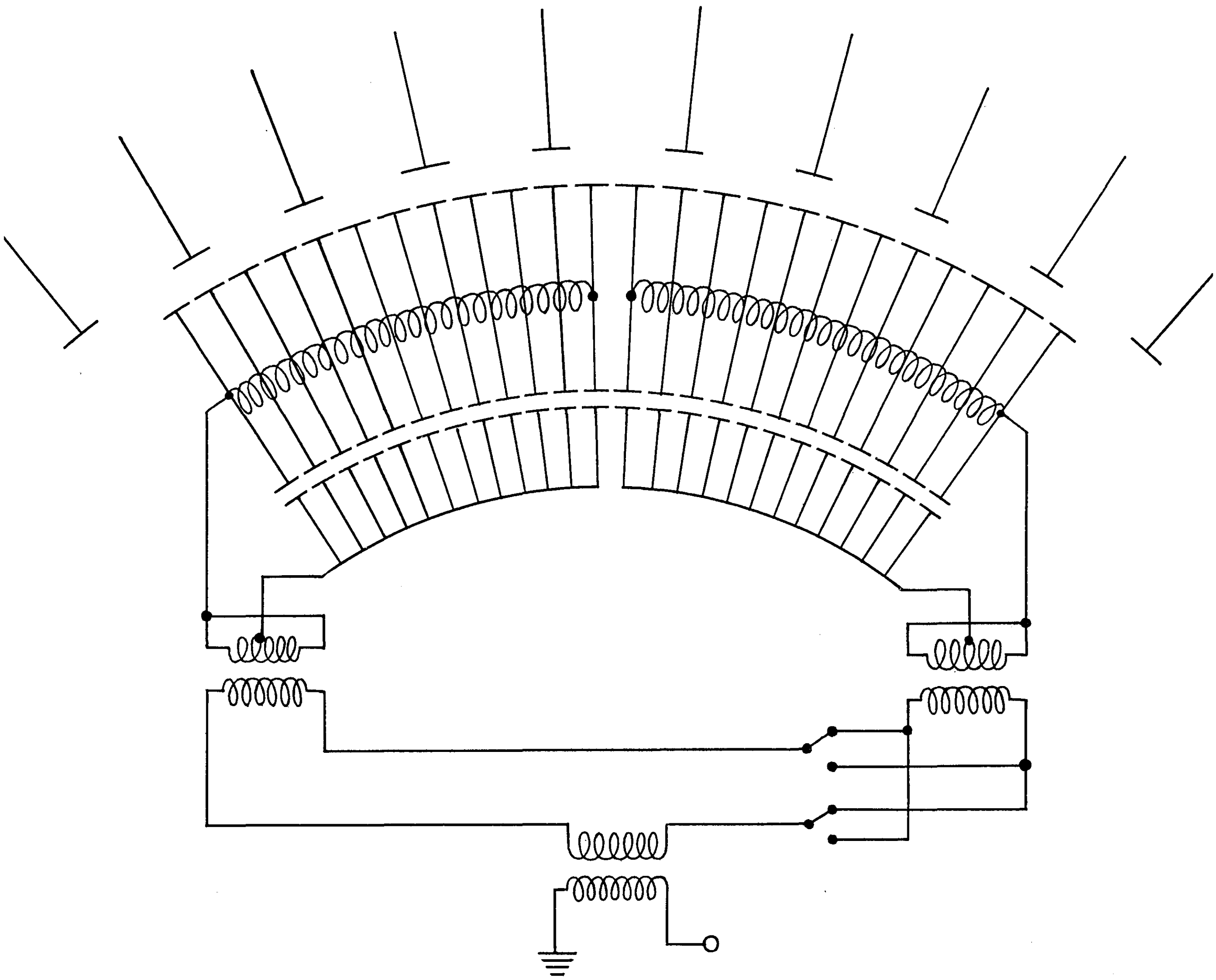


Fig. 1 - Capacitive Coupled Goniometer

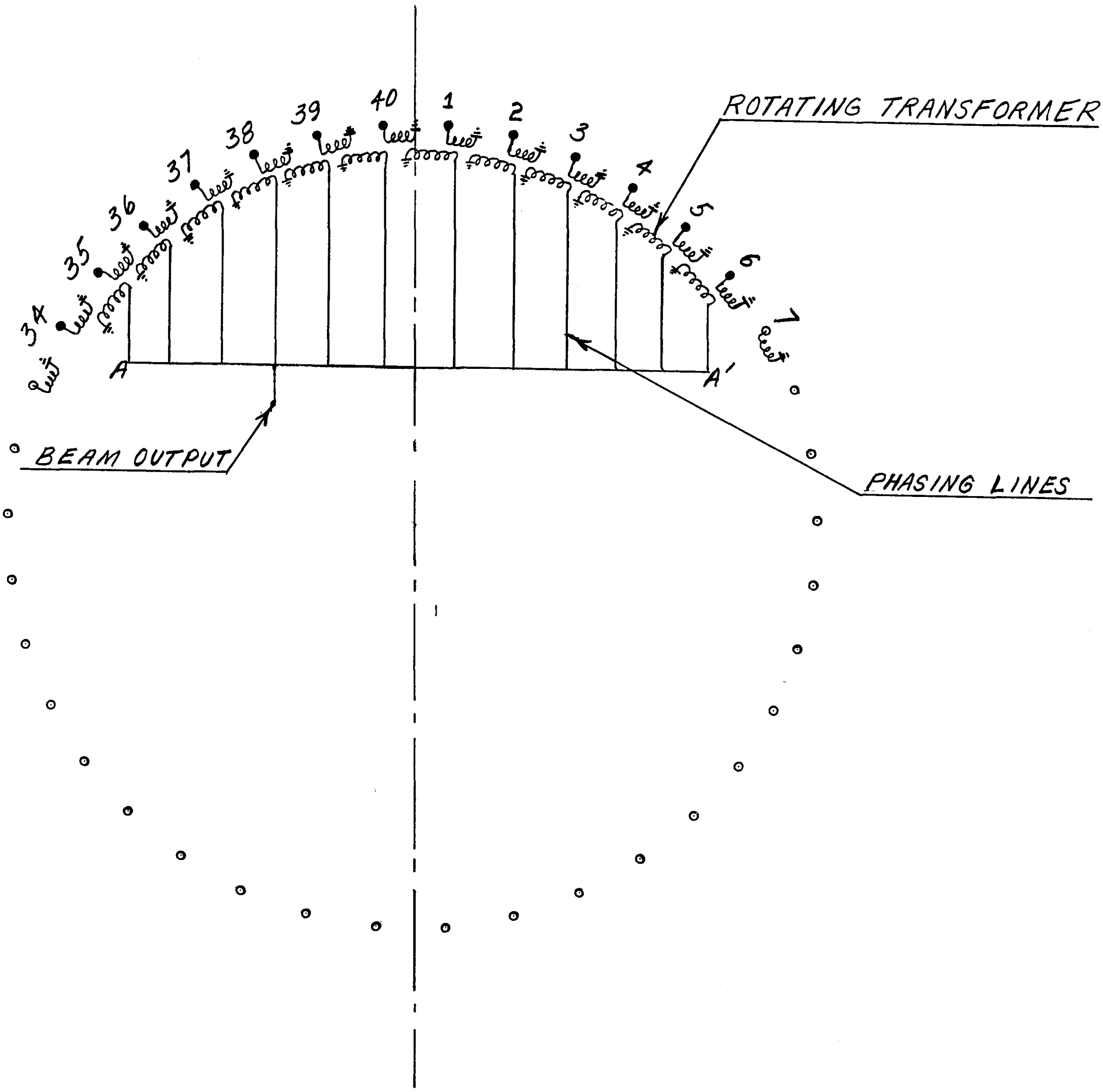
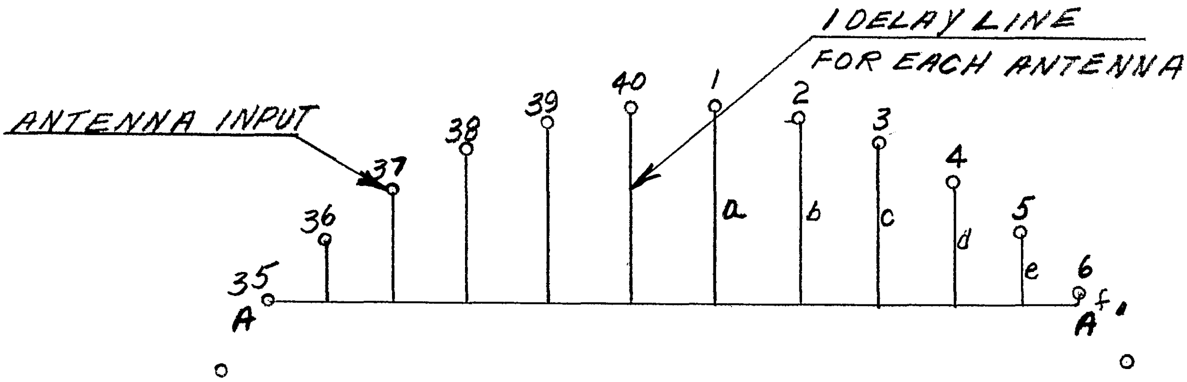
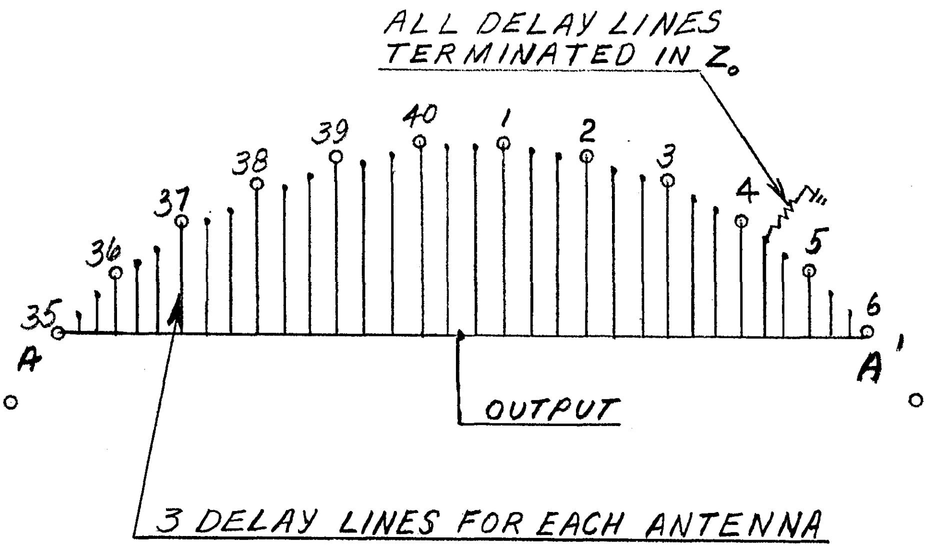


Fig. 2 - Inductive Coupled Goniometer



	DELAY LINE LENGTH (FEET IN AIR)
a	72
b	66
c	56
d	42
e	23
f	0

(a)



(b)

Fig. 3 - Phasing lines

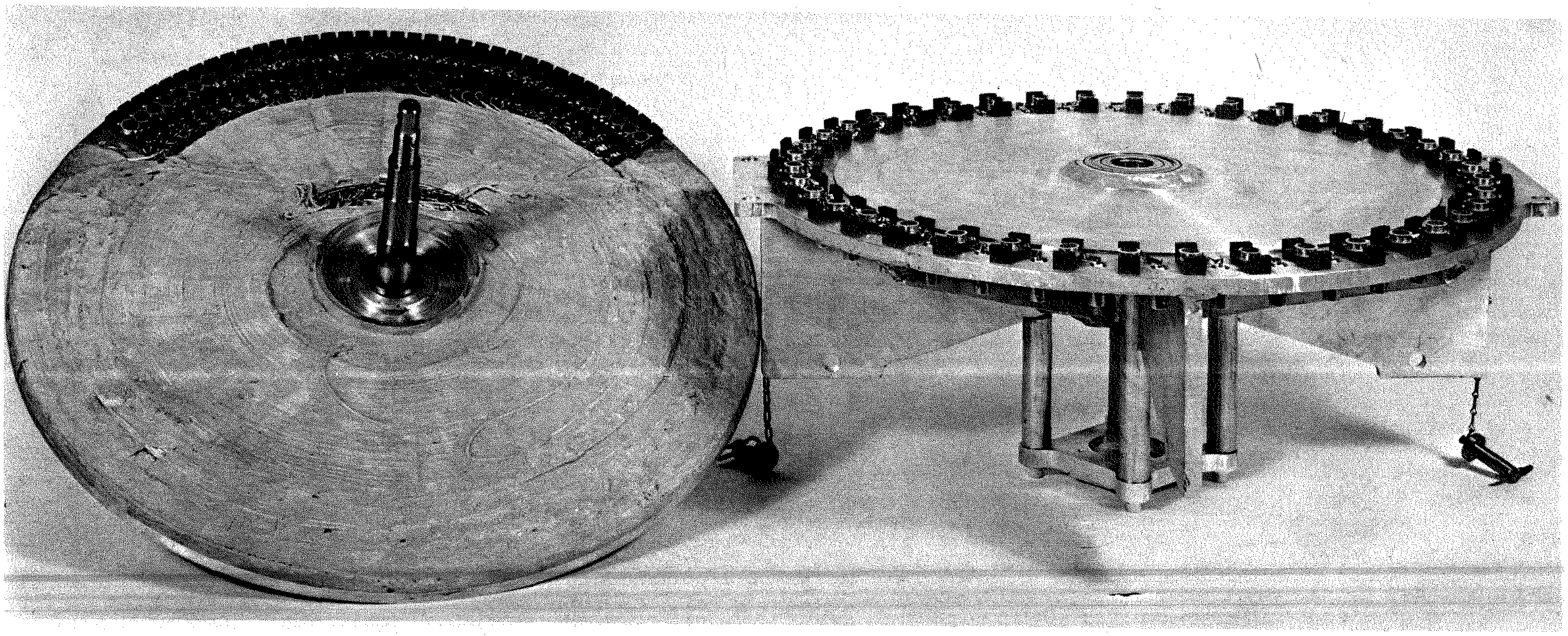


Fig. 4 - Inductive Coupled Goniometer (3 delay lines per antenna)

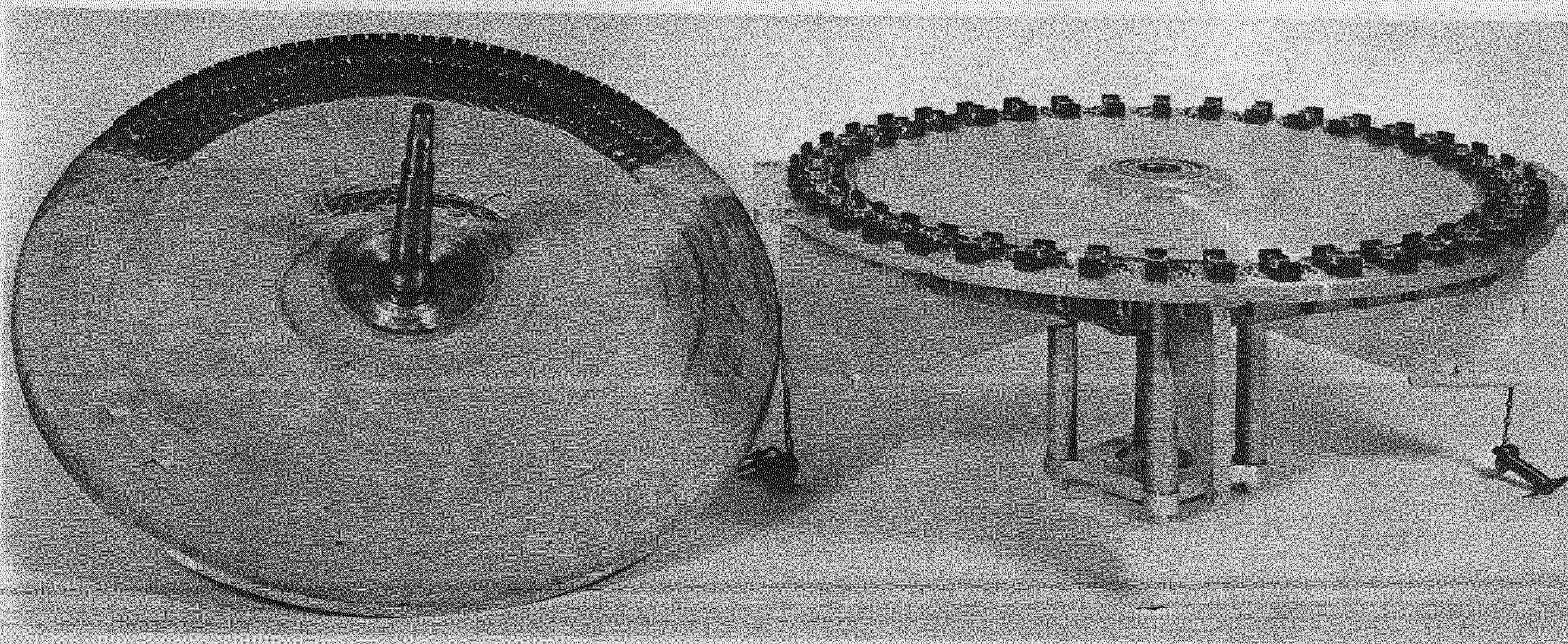


Fig. 4 - Inductive Coupled Goniometer (3 delay lines per antenna)

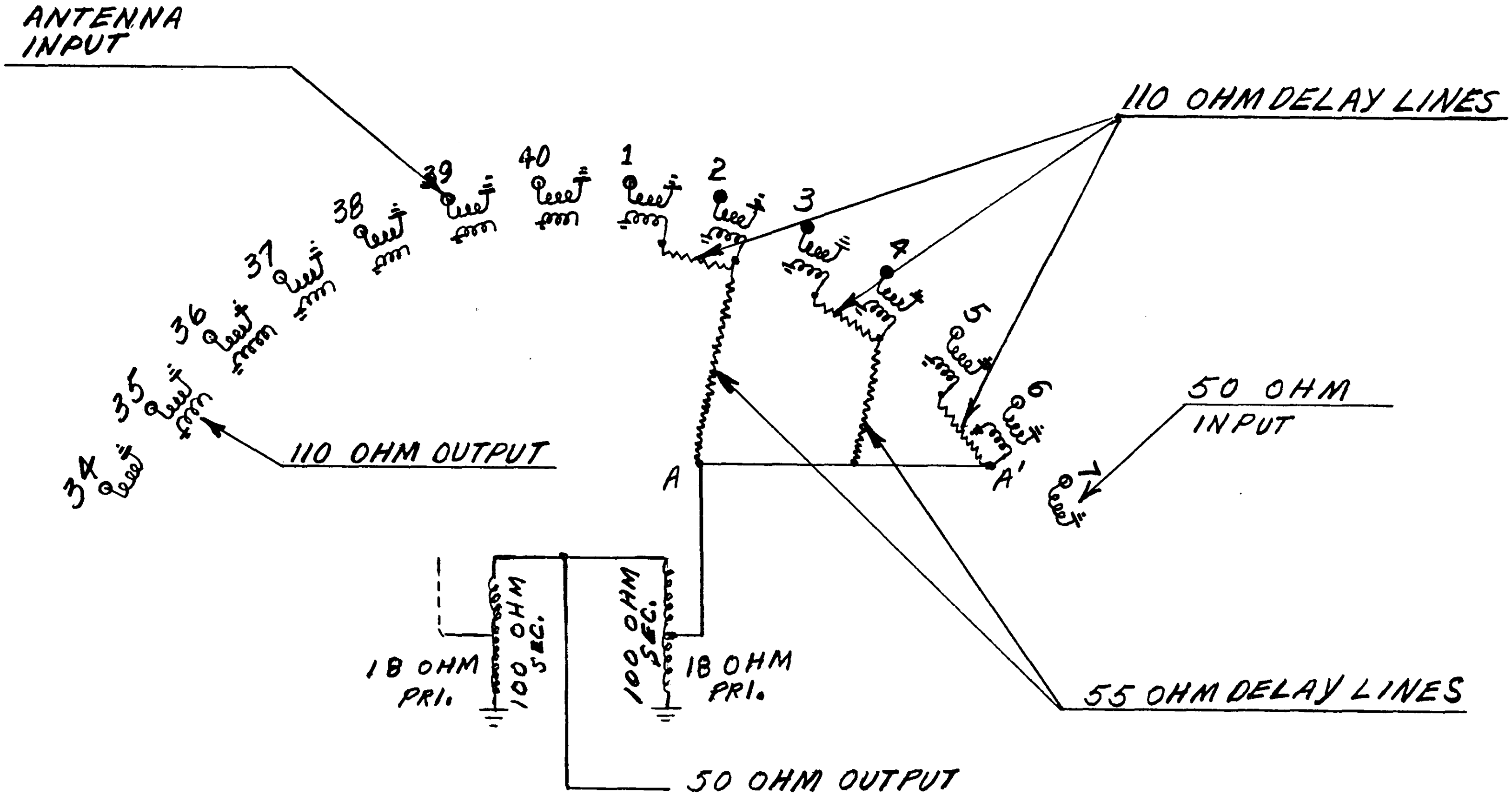


Fig. 6 - Inductive Coupled Goniometer (modified delay lines)

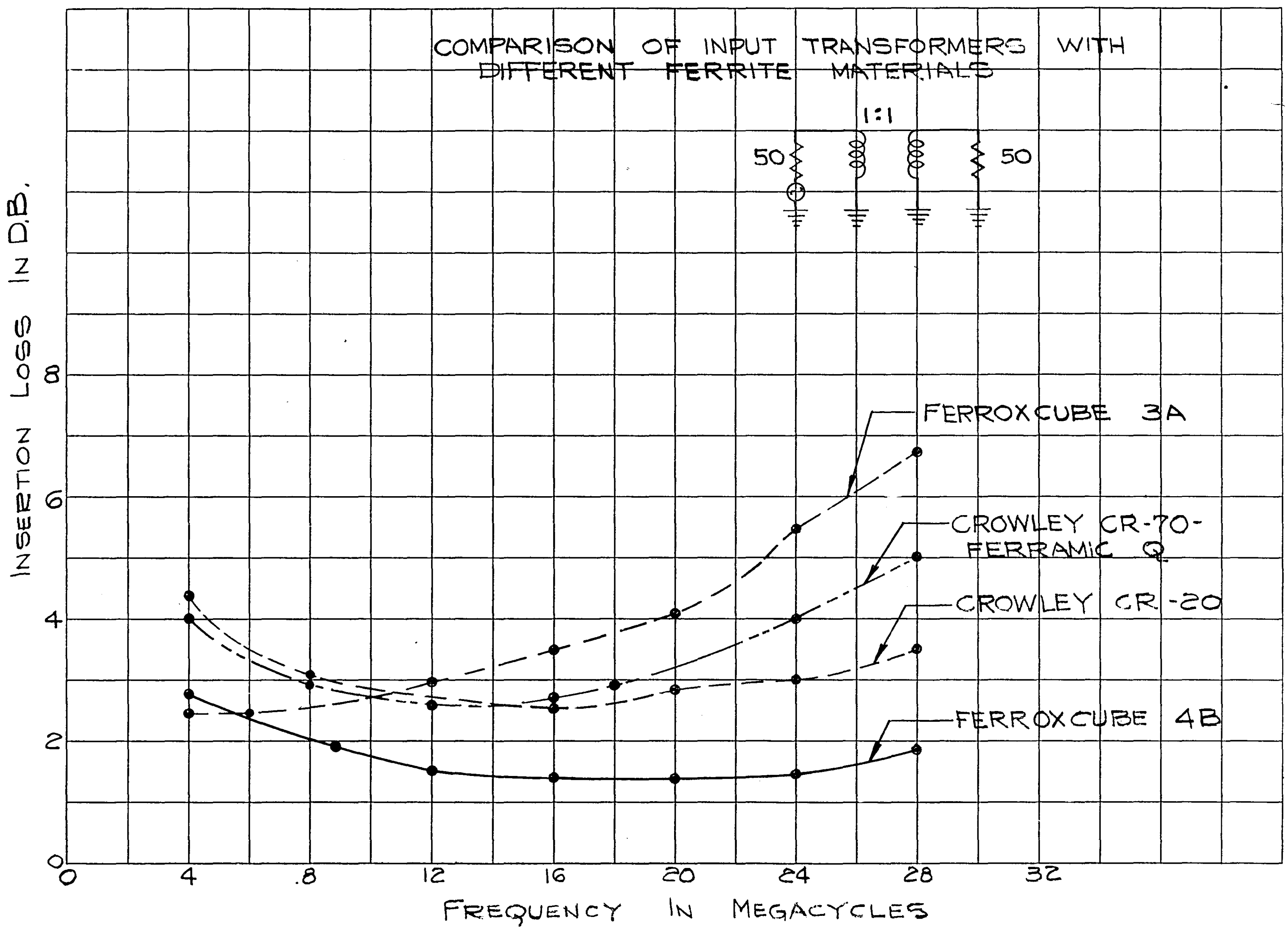


Fig. 7 - Comparison of Input Transformers with Different Ferrite Materials

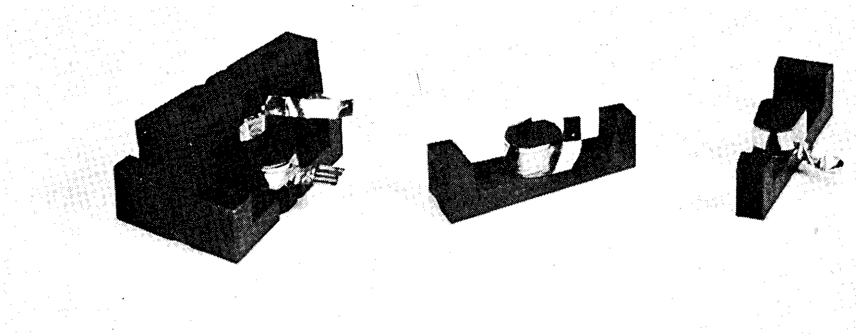


Fig. 8 - Input Coupling Transformers

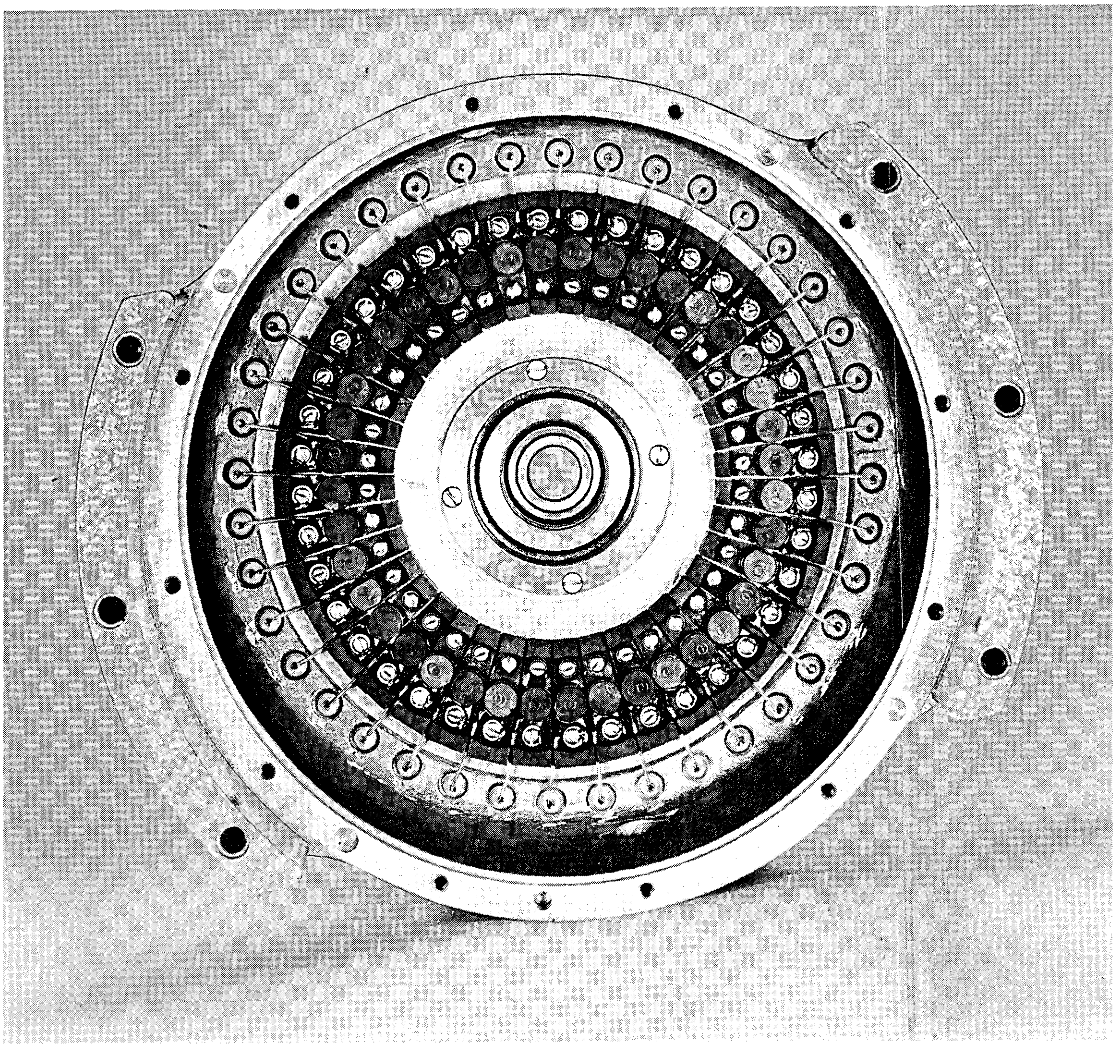


Fig. 9 - Inputs Transformer Assembled (primary)

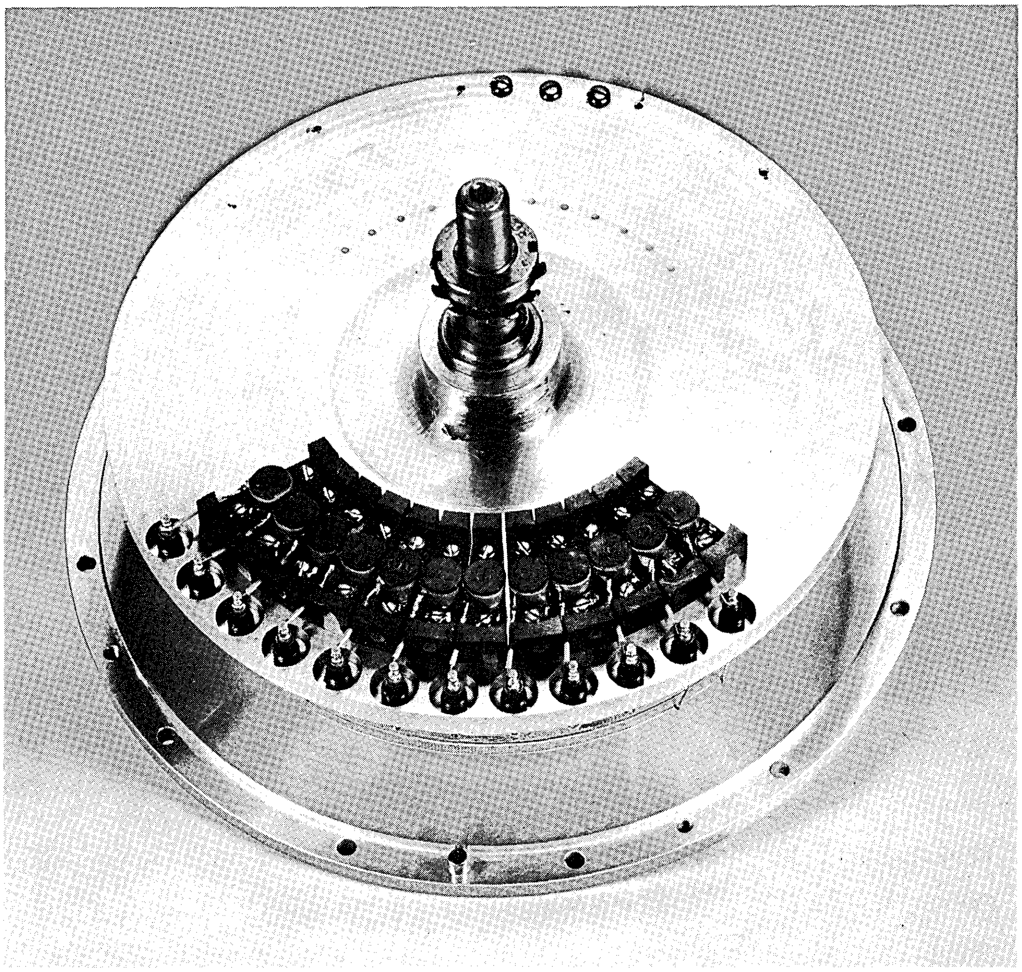


Fig. 10 - Input Transformers (secondary)

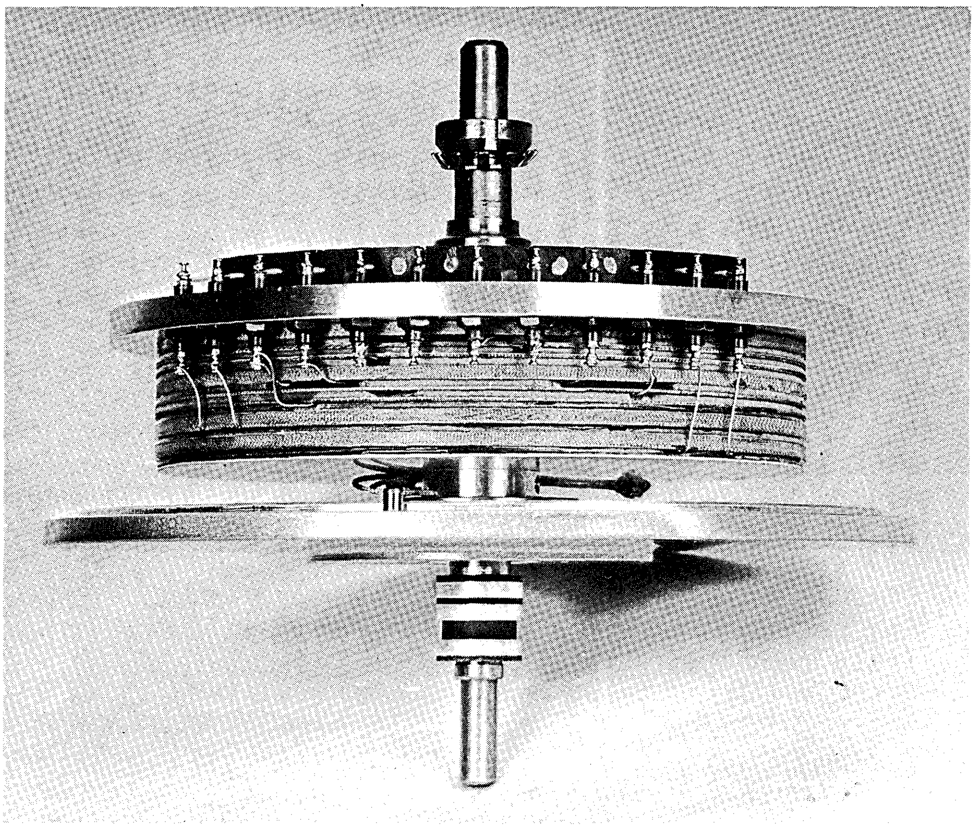


Fig. 11 - Input Transformer Secondary (side view)

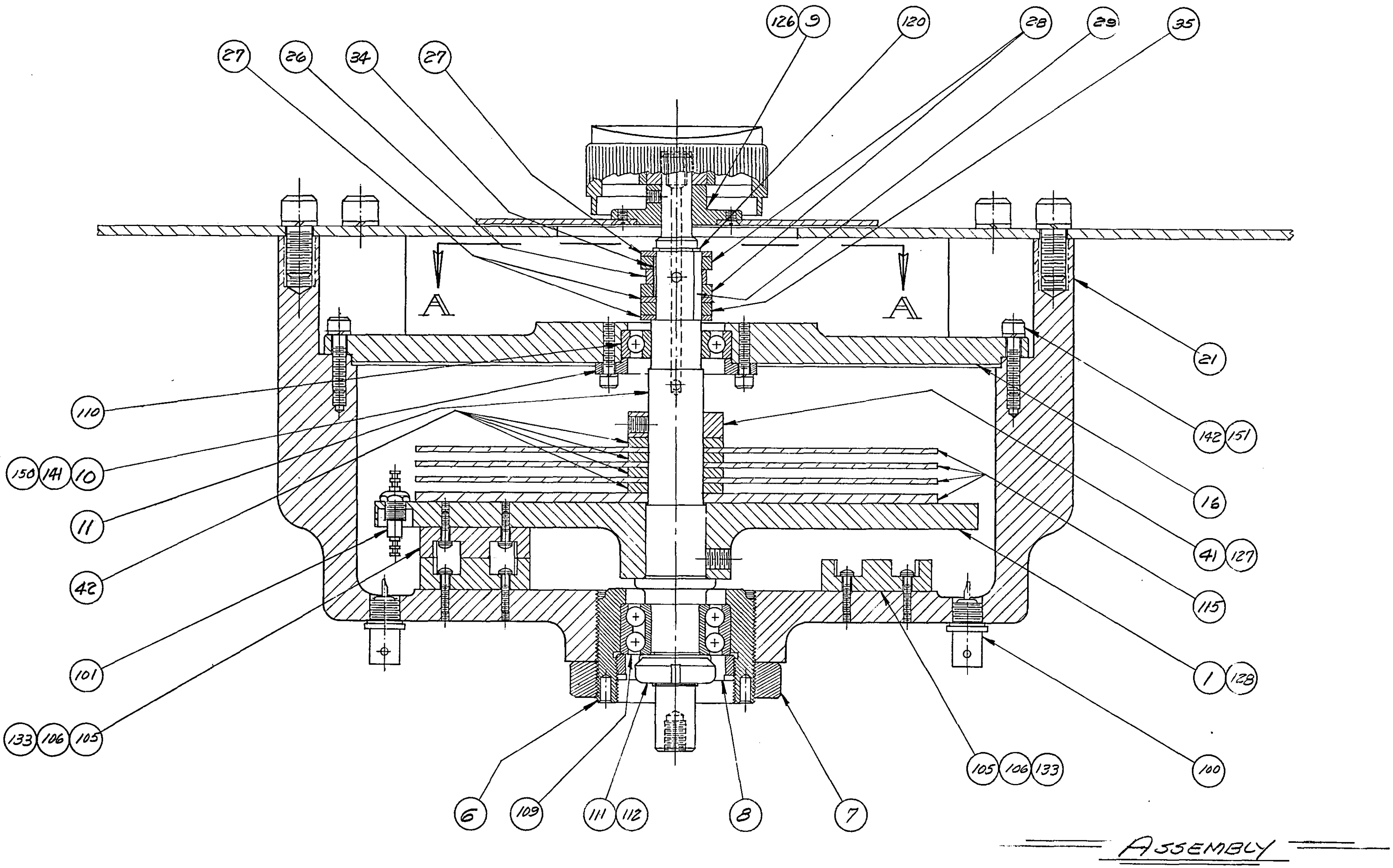


Fig. 12 - Cross Section of Goniometer

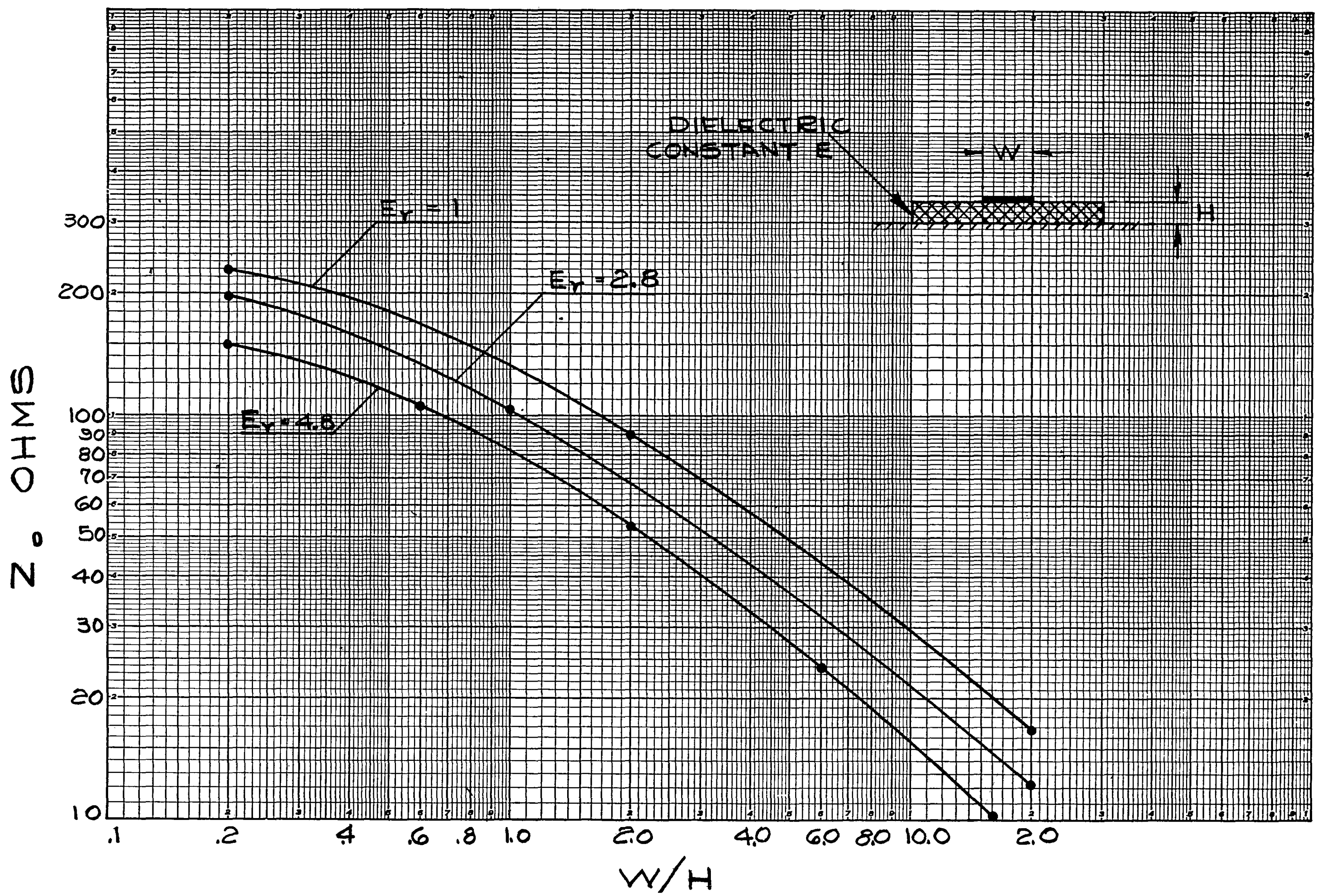


Fig. 13 - Characteristic Impedance of Microstrip

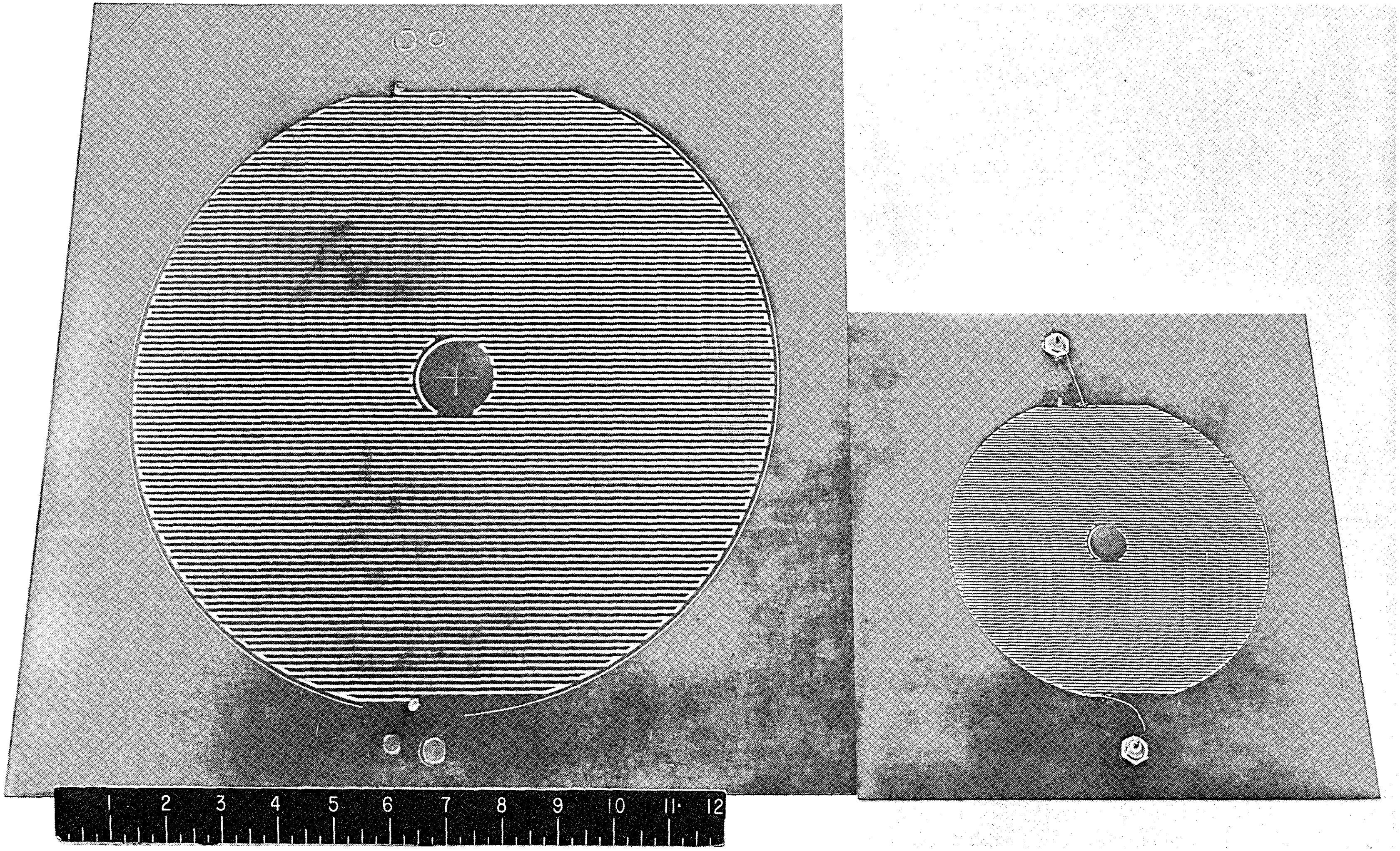


Fig. 14 - "Strip Line" Configurations

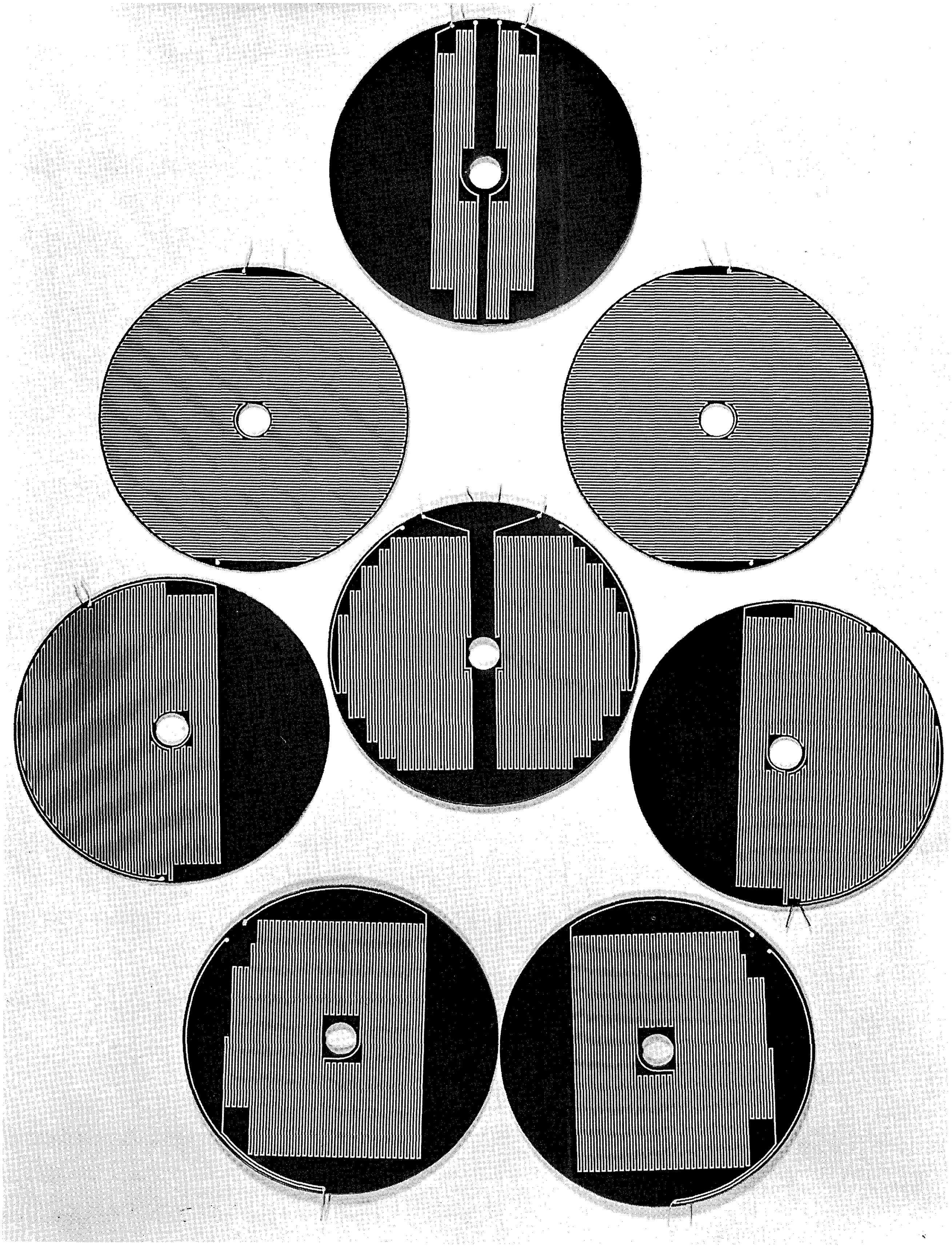


Fig. 15 - Delay lines ("strip line")

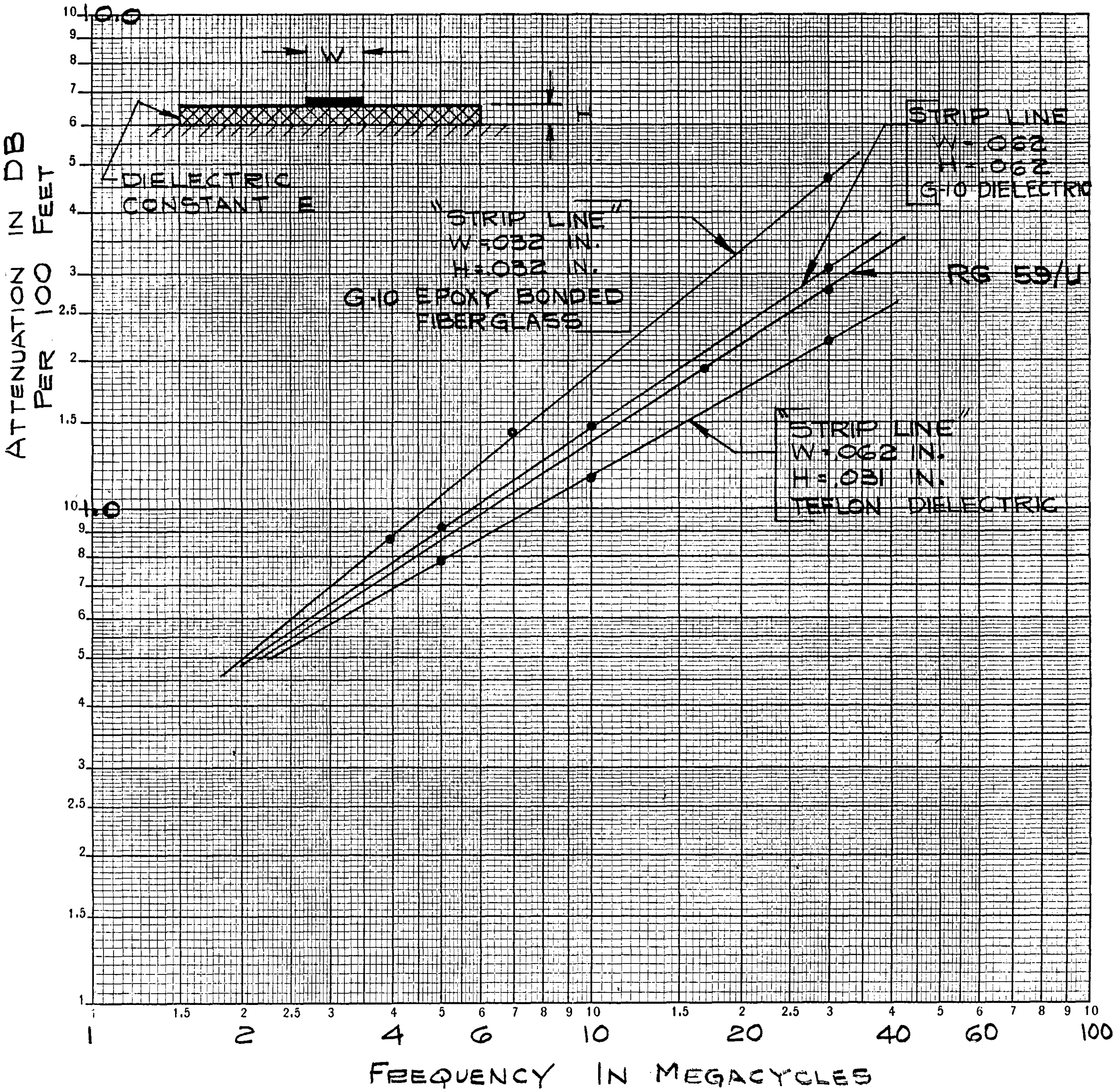


Fig. 16 - Attenuation for "Strip Line"

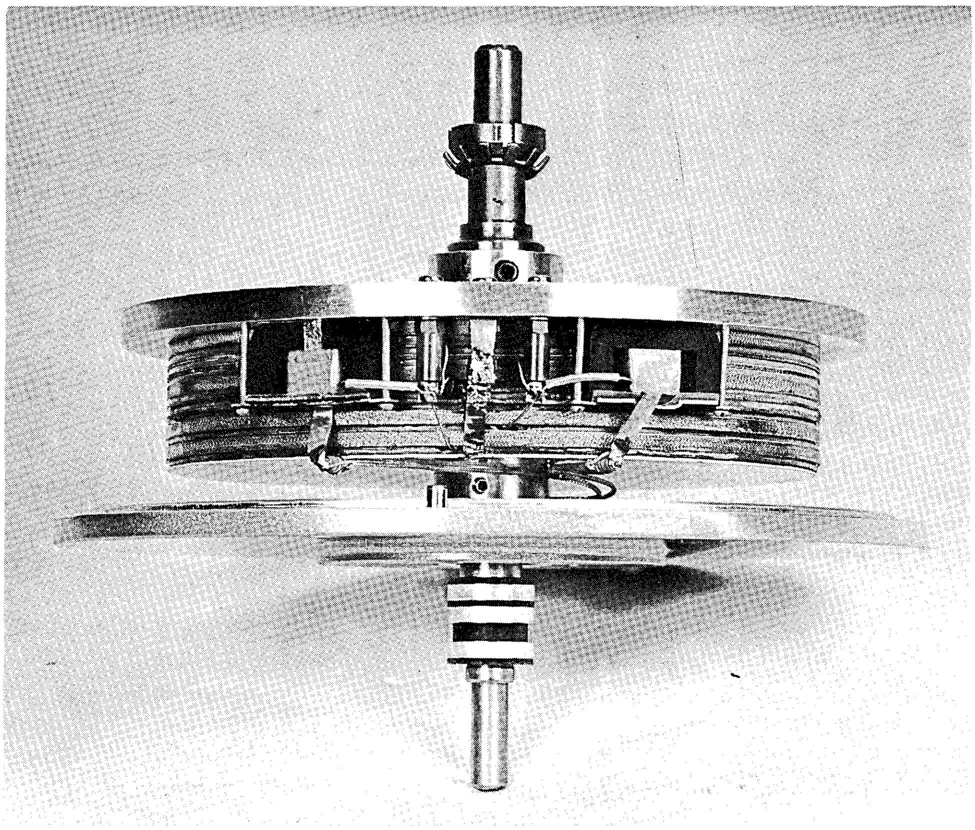


Fig. 17 - Assembled view of Goniometer Rotor

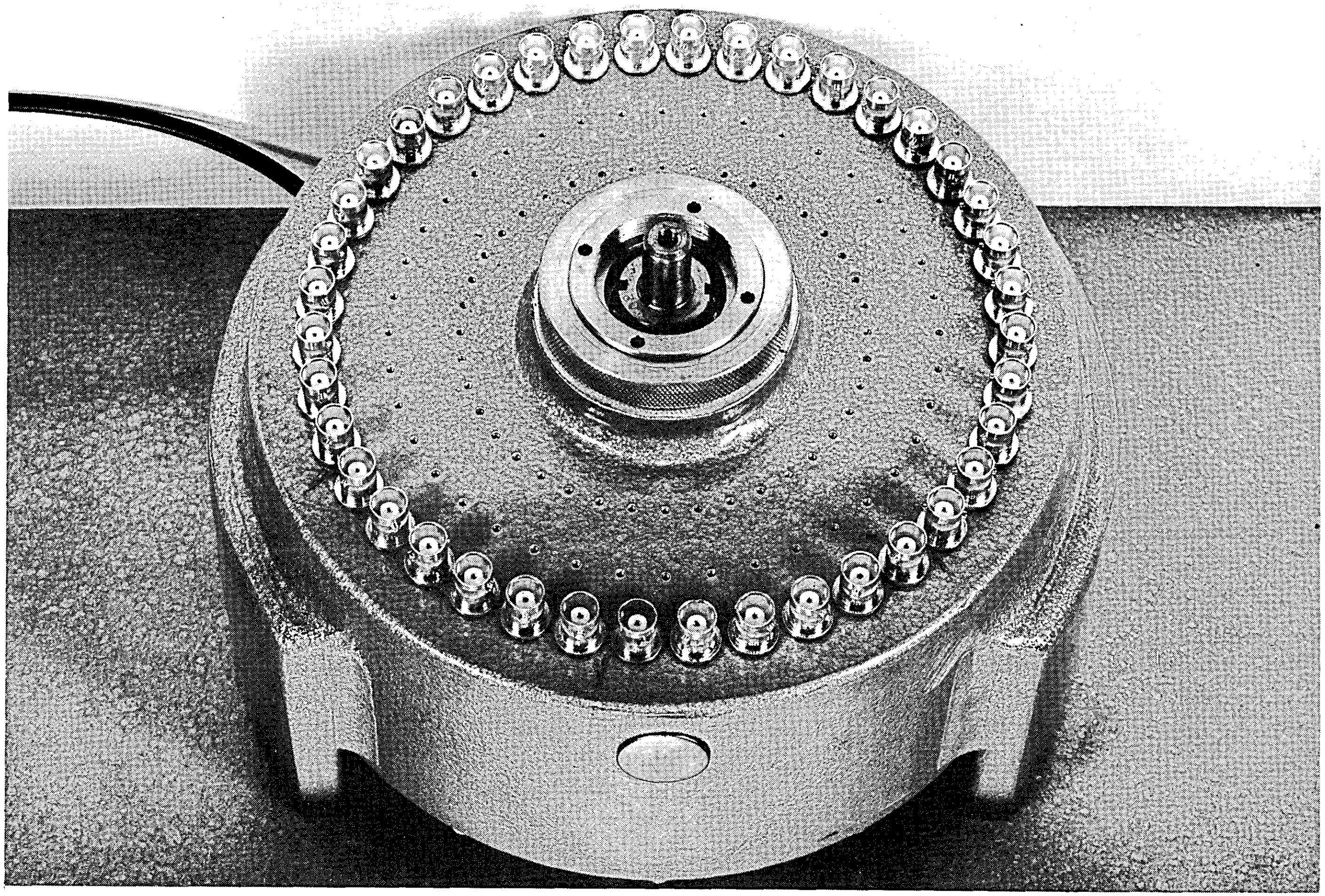


Fig. 18 - Assembled Goniometer (bottom view)

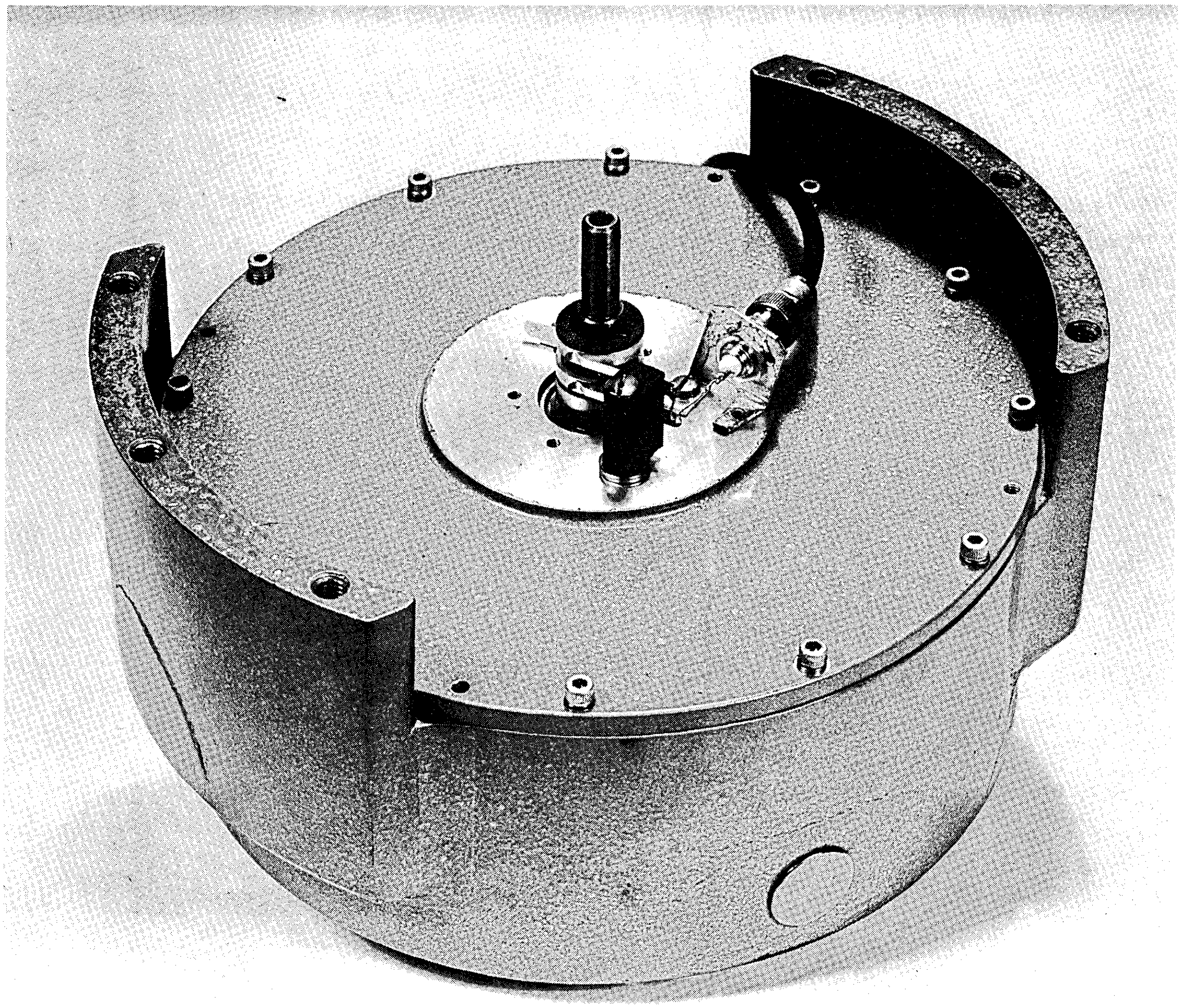


Fig. 19 - Assembled Goniometer (top view)

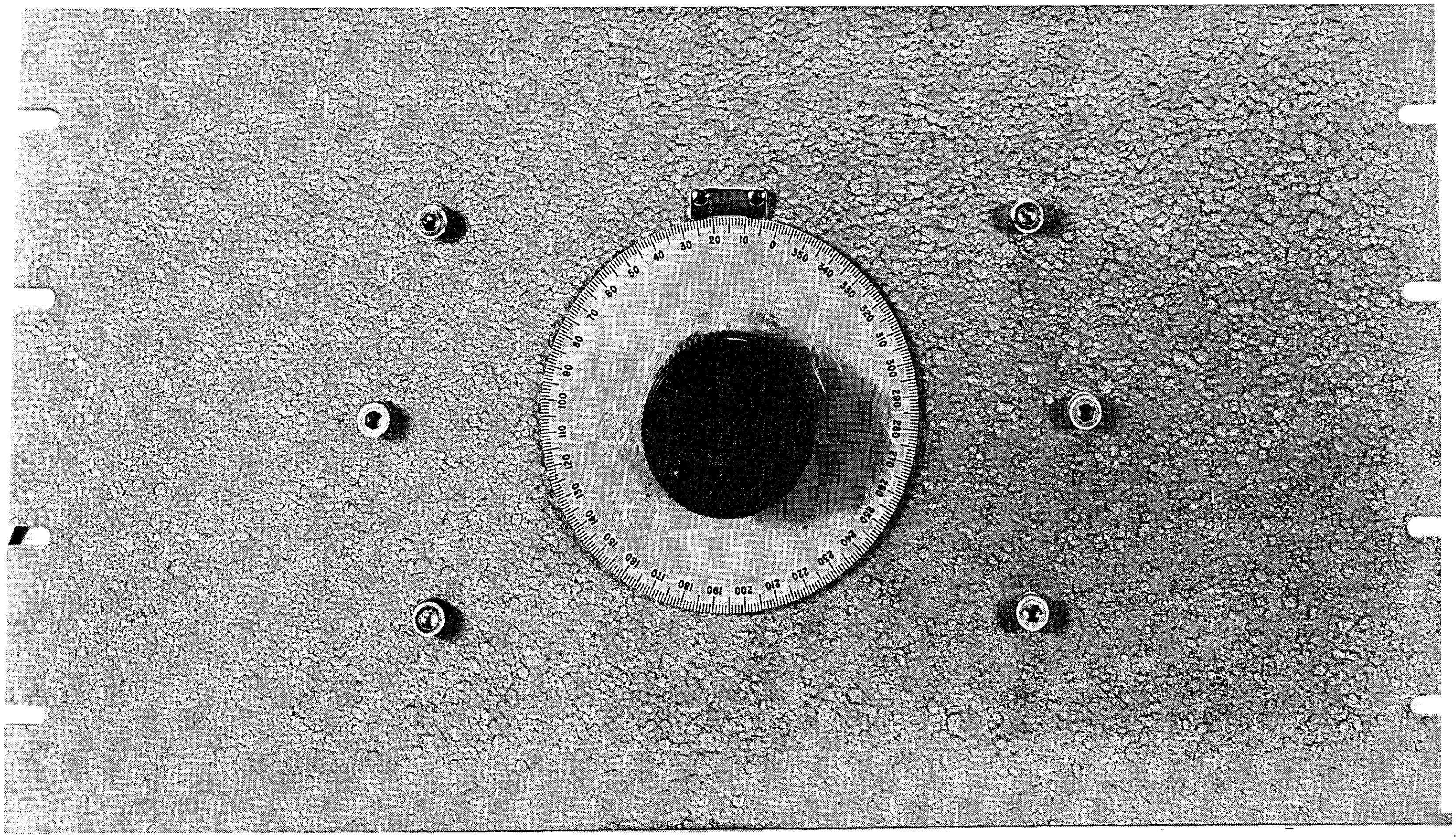


Fig. 20 - Panel Mounting of Goniometer

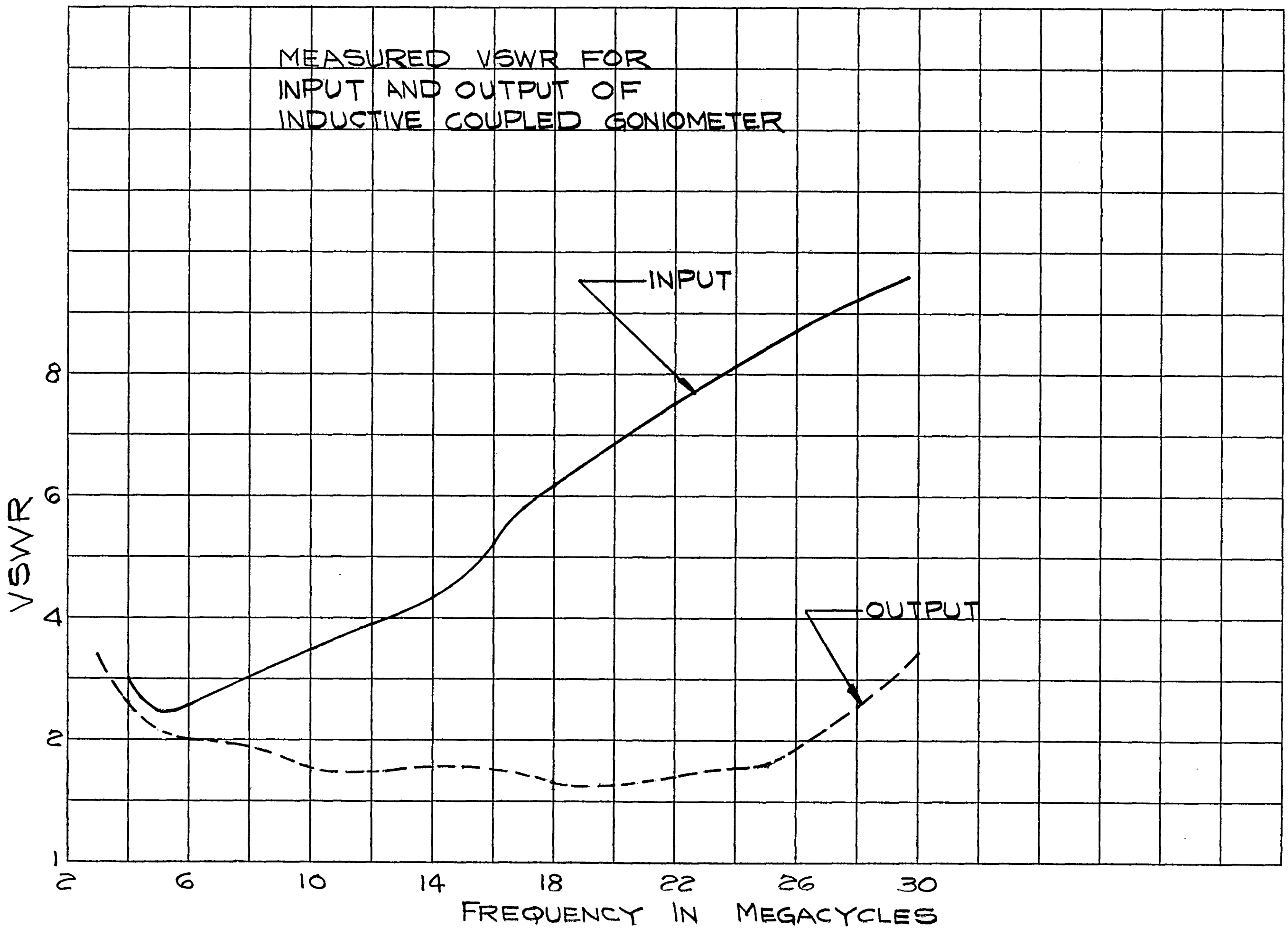


Fig. 21 - Measured VSWR for Input and Output of Inductive Coupled Goniometer

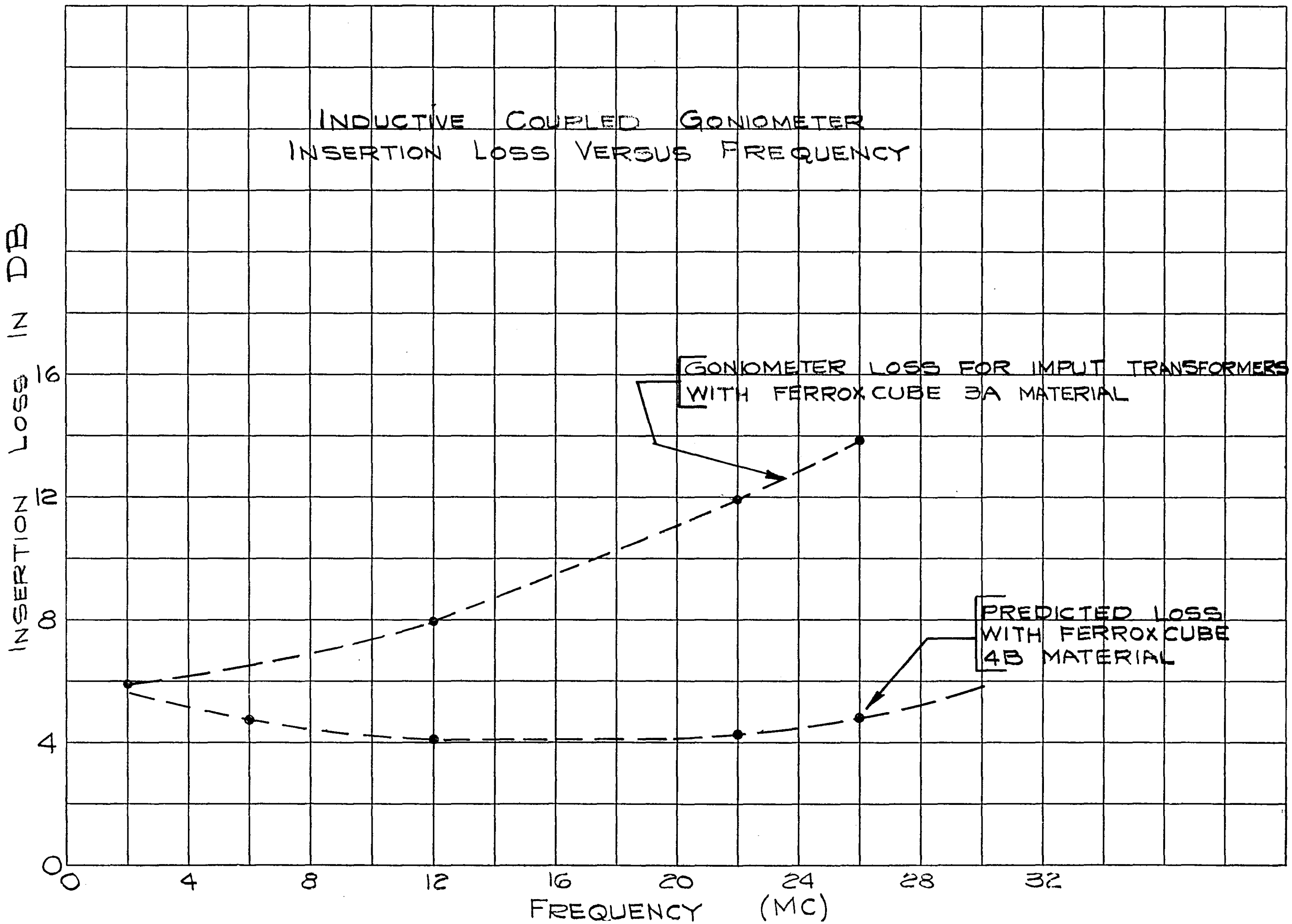
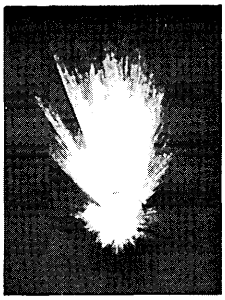
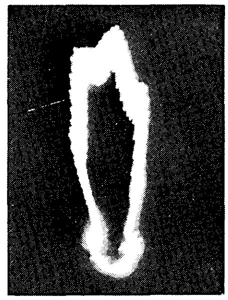


Fig. 22 - Inductive Coupled Goniometer Insertion Loss Versus Frequency

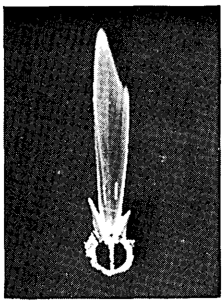


3.9 mc

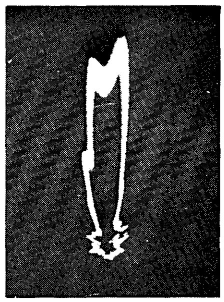
5 mc



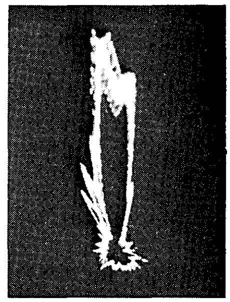
6 mc



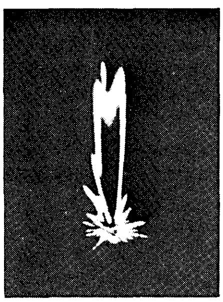
10 mc



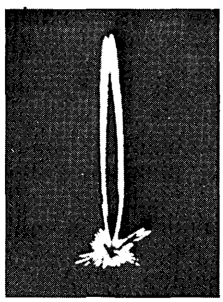
12.5 mc



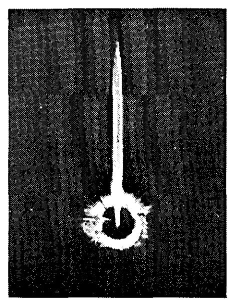
15 mc



17.2 mc

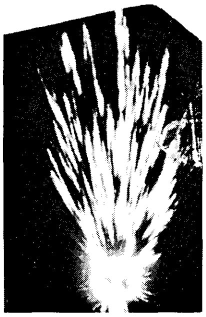


20 mc

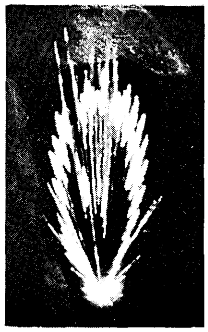


25 mc

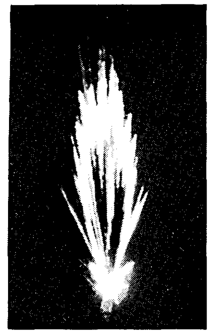
Fig. 23 - Antenna Beam Patterns for Inductive Goniometer
(1 delay line/antenna)



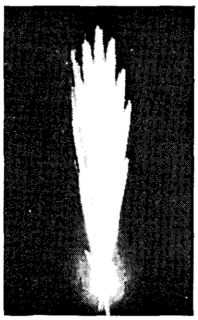
4 mc



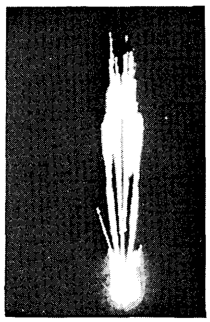
5 mc



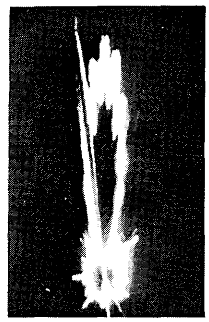
6.5 mc



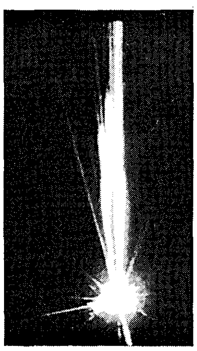
10 mc



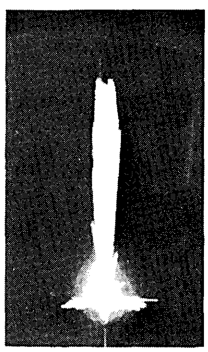
12.5 mc



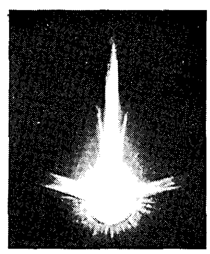
15 mc



17.2 mc

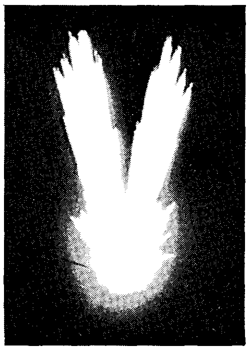


20 mc

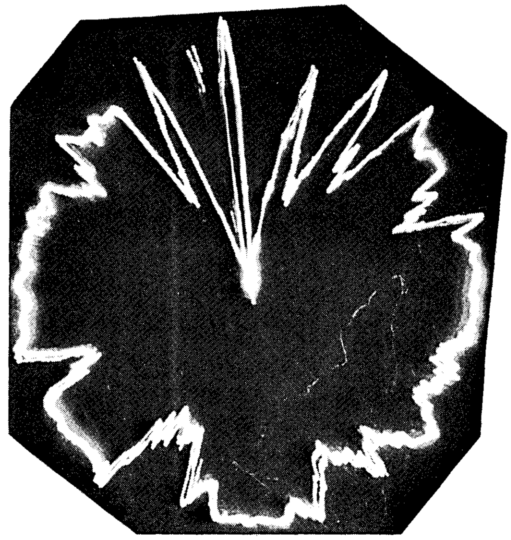


25 mc

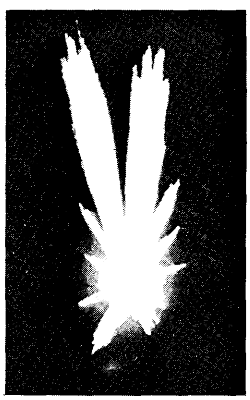
Fig. 24 - Antenna Beam Patterns for Inductive Coupled Goniometer (3 delay lines/antenna)



10 mc

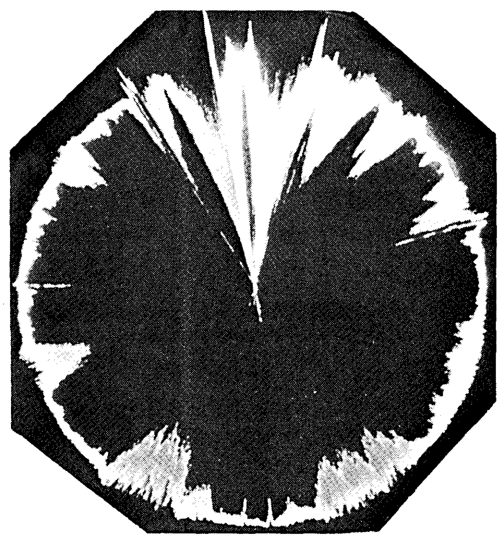


10 mc



15 mc

(a)



15 mc

(b)

Fig. 25 - Antenna Null Patterns for Inductive Coupled Goniometer (3 delay lines/antenna)