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ANTENNA MEASURING INSTRUMENT

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1.0 Introduction

This report describes an instrument that has been developed for measuring changes in the shape of the surface of a parabolic reflector.

A C.W. radar technique is used to measure distance from the focal point of the reflector to various points on the reflector surface. The system described is able to measure changes in distance from the focal point to 21 points on the surface simultaneously with a short-term accuracy of 0.002". The long-term stability of the instrument is affected by changes in the atmospheric refractive index. Compensation for these changes is possible and is routine on distance measuring equipment using a modulated light beam. No attempts at compensation have been made in the present equipment.

By using frequency switching techniques it would be possible to modify the instrument to measure absolute distance rather than changes in distance.

This type of measuring technique has been considered previously [1] and the radio telescope at Pulkova in the USSR reportedly uses a similar system although the accuracy claimed is less than the system described in this report.

The instrument has been installed on the 140-foot telescope and measurements made from the focal point to five points on the surface.

2.0 Principle of Operation

2.1 General Description

Figure 1 shows the basic components of the measuring system.

A stable oscillator at X-band (11.8 GHz) transmits a signal via a circulator and a broad beam horn antenna. A transponder is situated at the other end of the path over which the distance has to be measured. This transponder receives the carrier frequency, amplitude modulates it at a frequency of 455 kHz and retransmits it. The receiver is sensitive only to the transponded signal and rejects signals at the carrier frequency.

The output of the receiver is a signal at 455 kHz. The phase of this signal, referred to the modulating signal provided to the transponder, is a measure of the phase difference between transmitted and received signals at the carrier frequency. The wavelength of the carrier frequency is 1 inch, so a phase change of 1° corresponds to a change in the path length of about 0.0027".

2.2 Analysis of System

To avoid the introduction of many constants the magnitude of the various signals will be ignored and phase relationships only will be considered.

The output from the transmitter is given by cos $\omega_0 t$ where ω_0 is carrier frequency (in radians/sec).

The signal received at the transponder is cos $(\omega_0 t + \phi_1)$, where ϕ_1 is the phase shift at the carrier frequency and is given by

$$\phi_1 = \frac{2\pi d}{\lambda_0}$$

where d is the distance from the transmitter and λ_0 is the wavelength of the carrier frequency.

The transponder retransmits this signal as cos $\omega_m t$ cos $(\omega_0 t + \phi_1)$, where ω_m is the modulating frequency.

This may be rewritten as

$$\cos\left[\left(\omega_0 - \omega_m\right)t + \phi_1\right] + \cos\left[\left(\omega_0 + \omega_m\right)t + \phi_1\right]$$

or

$$\cos\left[\omega_L t + \phi_1\right] + \cos\left[\omega_u t + \phi_1\right]$$

where $\omega_u = \omega_0 + \omega_m$ (the upper sideband)

$\omega_L = \omega_0 - \omega_m$ (the lower sideband).

The retransmitted signal will be phase shifted again on its return to the receiver. The signal received will be

$$\cos\left[\omega_L t + \phi_1 + \frac{2\pi d}{\lambda_L}\right] + \cos\left[\omega_u t + \phi_1 + \frac{2\pi d}{\lambda_u}\right]$$

where λ_u = wavelength of upper sideband

λ_L = wavelength of lower sideband.

The inputs to the two mixers (assuming equal path lengths in the receiver) will be given by

$$\cos\left[\omega_L t + \phi_1 + \phi_2 + \frac{2\pi d}{\lambda_L}\right] + \cos\left[\omega_u t + \phi_1 + \phi_2 + \frac{2\pi d}{\lambda_u}\right]$$

where ϕ_2 is the additional path length added in the microwave circuit of the receiver.

This may be rewritten as

$$\cos\left[\omega_L t + \Phi_1\right] + \cos\left[\omega_u t + \Phi_2\right] \tag{1}$$

where $\Phi_1 = \phi_1 + \phi_2 + \dfrac{2\pi d}{\lambda_L}$

and $\Phi_2 = \phi_1 + \phi_2 + \dfrac{2\pi d}{\lambda_u}$

The output from mixer A will be the input multiplied by $\cos \omega_0 t$. This gives

$$\cos\left[\left(\omega_L + \omega_0\right)t + \Phi_1\right] + \cos\left[\left(\omega_L - \omega_0\right)t + \Phi_1\right]$$

$$+ \cos\left[\left(\omega_u + \omega_0\right)t + \Phi_2\right] + \cos\left[\left(\omega_u - \omega_0\right)t + \Phi_2\right]$$

The IF amplifier following the mixer will pass only frequencies at ω_m, so the output from this amplifier will be

$$\cos\left[\left(\omega_L - \omega_0\right)t + \Phi_1\right] + \cos\left[\left(\omega_u - \omega_0\right)t + \Phi_2\right] \tag{2}$$

The output from mixer B will be the input (equation 1) multiplied by $\cos\left(\omega_0 t - \frac{\pi}{2}\right)$, which gives

$$\cos\left[\left(\omega_L + \omega_0\right)t + \Phi_1 - \frac{\pi}{2}\right] + \cos\left[\left(\omega_L - \omega_0\right)t + \Phi_2 + \frac{\pi}{2}\right]$$

$$+ \cos\left[\left(\omega_u + \omega_0\right)t + \Phi_2 - \frac{\pi}{2}\right] + \cos\left[\left(\omega_u - \omega_0\right)t + \Phi_2 + \frac{\pi}{2}\right]$$

Once again the IF amplifier rejects frequencies other than ω_m, so the output will be

$$\cos\left[\left(\omega_L - \omega_0\right)t + \Phi_1 + \frac{\pi}{2}\right] + \cos\left[\left(\omega_u - \omega_0\right)t + \Phi_2 + \frac{\pi}{2}\right]$$

which may be rewritten in terms of ω_m as

$$\cos\left[-\omega_m t + \Phi_1 + \frac{\pi}{2}\right] + \cos\left[\omega_m t + \Phi_2 + \frac{\pi}{2}\right]$$

or

$$\cos\left[\omega_m t - \Phi_1 - \frac{\pi}{2}\right] + \cos\left[\omega_m t + \Phi_2 + \frac{\pi}{2}\right]$$

This is phase shifted $-\frac{\pi}{2}$ at the modulating frequency which gives

$$\cos\left[\omega_m t - \Phi_1 - \pi\right] + \cos\left[\omega_m t + \Phi_2\right] \tag{3}$$

Equation (2) may be expressed in terms of ω_m as

$$\cos\left[-\omega_m t + \Phi_1\right] + \cos\left[\omega_m t + \Phi_2\right]$$

or

$$\cos\left[\omega_m t - \Phi_1\right] + \cos\left[\omega_m t + \Phi_2\right] \tag{4}$$

The outputs of the two IF amplifiers are then added. When equation (3) and equation (4) are added, the first two terms cancel and a term remains that is proportional to $\cos\left[\omega_m t + \Phi_2\right]$.

This signal, after passing through a limiting amplifier, is phase detected using ω_m as a reference. This phase detector is linear and has a range of 360°. The output will be proportional to Φ_2 which is equal to

$$\phi_1 + \phi_2 + \frac{2\pi d}{\lambda_u}$$

but $\phi_1 = \frac{2\pi d}{\lambda_0}$ and $\frac{1}{\lambda_u} = \frac{1}{\lambda_0} + \frac{1}{\lambda_m}$. ϕ_2 is a constant.

The output of the phase detector is then

$$\frac{2\pi d}{\lambda_0} + \frac{2\pi d}{\lambda_0} + \frac{2\pi d}{\lambda_m} \tag{5}$$

$\lambda_m >> \lambda_0$, so the output is proportional to

$$\frac{d}{\lambda_0} \tag{6}$$

λ_0, the wavelength in air, is variable and this variability imposes long-term stability limitations on the instrument. This is analyzed in detail in section 6.0.

3.0 Signal Strength Analysis

Microwave oscillators have improved very rapidly over the last year. At the time of ordering the oscillator for the system it seemed that a reasonable choice would be to use a source that utilized a low frequency oscillator multiplied to X-band. A better and cheaper source today would be a Gunn-diode oscillator. This is discussed further in section 7.0.

The oscillator used was a Frequency Sources type FS-24 with a power output of 50 mW at 11.802 GHz ($\lambda = 1.000''$).

The design of the transmitter/receiver horn antenna was quite difficult. The angle subtended by the 140' reflector at the focal point is 135° and, unlike the normal feed horn, we would like full illumination at the edges. Open-ended waveguide at this frequency gives a half-power beamwidth of about 70° in the H-plane and 120° in the E-plane. The design of very wide beamwidth horns seems to be empirical. It was found that by sliding the plate behind the horn (see Figure 6) we were able to obtain a half-power beamwidth of 110° in the H-plane and 140° in the E-plane.

Assume a value of 3 (5 dB) for the gain of the horn. The power transmitted, assuming some loss in the circulator and microwave circuit and subtracting 10 mW for LO power, will be approximately +15 dBm.

The power received by a transponder will be

$$\frac{P_T G_T A_R}{4\pi R^2}$$

where P_T = transmitted power

G_T = gain of transmitting antenna

A_R = area of transponder antenna

R = distance between transmitter and receiver.

The power received back at the transmitter after retransmission will be

$$\left[\frac{P_T G_T A_R}{4\pi R^2}\right] \left[\frac{G_R A_T}{4\pi R^2}\right]$$

where G_R = gain of transponder horn

and A_T = area of transmitting horn.

The ratio of received transmitted power is then given by

$$\frac{P_R}{P_T} = \frac{G_T A_R G_R A_T}{16\pi^2 R^4} \tag{7}$$

The gain and effective area of an antenna are related by

$$G = \frac{4\pi A}{\lambda^2}$$

so $A_T = \dfrac{G_T \lambda^2}{4\pi}$ and $G_R = \dfrac{4\pi A_R}{\lambda^2}$

Therefore equation (7) simplifies to

$$\frac{P_R}{P_T} = \frac{G_T^{\,2} A_R^{\,2}}{16\pi^2 R^4} \tag{8}$$

Say the area of the receiving antenna is equal to the square of some dimension, d. Then equation (8) simplifies to

$$\frac{P_R}{P_T} = \frac{9}{160} \left[\frac{d}{R}\right]^4$$

with $G_T = 3$.

Assume d = 4", then we have, expressing $\frac{P_R}{P_T}$ in dB,

$$10 \log \frac{P_R}{P_T} = -12.5 - 40 \log (210)$$

$$= -105 \text{ dB}$$

so $$P_R = -90 \text{ dBm}.$$

A suitable horn for the transponder is a 23 dB horn with dimensions as shown in Figure 12. This horn has a beamwidth of 18°, so alignment of the transponders on the telescope should not be difficult.

The receiver noise is given by

$$P_n = KT_0F \times 2B$$

where B is the bandwidth after the phase detector and F is the receiver noise figure. KT_0 may be conveniently expressed as

-144 dBm per kHz

for $T_0 = 300$ °K, so for a bandwidth of 100 Hz and a receiver noise figure of 15 dB the noise power comes out at -136 dBm, giving

$$\frac{P_R}{P_N} = 46 \text{ dB}.$$

This noise shows at the output of the limiting amplifier as phase jitter. The rms value of this may be calculated by considering the noise voltage to be added vectorially to the signal voltage. In this case the noise voltage is $\frac{1}{200}$ of the signal voltage, so the rms value of the phase jitter will be $\frac{1}{200}$ radians or 0.29°.

This analysis neglects the effect of any noise in the spectrum of the oscillator at 455 kHz away from the carrier. In practice this noise was found to degrade the noise figure of the receiver to approximately 19 dB which should give a signal to noise of 42 dB and a path length jitter of 0.0012". The experimental results obtained were slightly worse than this owing to fall off in oscillator power.

4.0 Description of System

4.1 General Description and Operating Procedures

The block diagram of the system is shown in Figure 2. The transmitter/receiver is mounted in a temperature controlled box at the focal point of the reflector. The display and control sections are in the telescope control room. About 300' of RG 58 cable connect each transponder to the control section.

A photograph of the control and display unit is shown in Figure 3. Each transponder is identified by an indicator lamp which has associated with it a thumbwheel switch (for coarse phase control) and a ten-turn potentiometer (for fine phase control). A digital display reads displacements in increments of 0.001".

Two operating modes are possible. In the manual mode any one transponder may be selected and distance changes monitored on the digital meter. Normally, this mode is used for initial set up of each transponder output at the start of a measuring period.

The automatic mode involves switching a transponder on for a period of 18 ms, measuring the distance and storing the result in a sample-hold circuit. The next transponder is then switched on and the procedure repeated. This scanning technique gives 21 voltage outputs that represent the 21 distance changes. These voltages are available continuously and are updated every 0.44 sec.

A typical set up procedure involves setting the output from each channel to be zero with the telescope set at zenith. This is done by using the manual mode and setting the distance output to zero for each transponder by using the fine and coarse phase controls. The coarse phase control is a thumbwheel switch giving 45° steps in phase to the modulating signal provided to the transponder. This corresponds to total path length change of 0.125", or a distance change of 0.0625". The final nulling of the meter is done with an offset controlled by the fine phase control pot. The instrument may then be switched to the automatic mode and the 21 outputs continuously recorded as the telescope is moved.

4.2 Detailed Description

The modulating frequency and scanning frequency are derived by digitally dividing the output of a 3.64 MHz clock. A 3-bit counter driven by the clock generates the 455 kHz reference frequency for the linear phase detector in the transmitter/receiver package. The clock also provides an input to the programmable digital phase shifter. The output of this phase shifter is a 455 kHz signal, the phase of which may be adjusted with respect to the 455 kHz reference in steps of 45°. The phase shift is controlled by the thumbwheel switches on the front panel, each switch having control when its particular channel is activated. This phase shifted signal is supplied to the transducers via 21 diode switches. The programming signal is derived from the 455 kHz signal via a 13-bit counter. This 55 Hz signal drives a ÷ 24 counter which is used to provide the switching signals. Three outputs from this counter are not used.

To make the operations of the system clearer, we can trace what happens when the ÷ 24 counter is in position A. Diode switch A_3 closes and supplies transponder A with a 455 kHz signal. The phase of this signal is determined by the position of thumbwheel switch A as this is gated into the phase shifter via A1, which is now closed. The linear phase detector gives out a voltage proportional to change is distance. This voltage is added to the voltage from the fine phase control pot via FET switch A_2, which is now closed. After a few milliseconds to allow for build up in the low-pass filter, the output of the summing amplifier is sampled by the sample-hold circuit A. In the automatic mode the programmer will switch to position B after a period of 18 ms.

4.21 Transmitter/Receiver

The transmitter/receiver is mounted in a temperature controlled box, the dimensions of which are 22" x 13" x 9".

Figure 4 shows the circuit diagram of the IF amplifiers, summing circuits, and phase detector.

The mixers are Aertech type MX 12000 which have a noise figure of 10 dB. The IF amplifiers have a low-noise tuned FET input stage followed by a 10 dB amplifier (Q2 and Q3). Two integrated circuit amplifiers and a ceramic filter give additional gain and selectivity.

The noise figure of one channel of the receiver was measured using an X-band noise source at the RF input and a value of 19 dB was obtained. This was surprising as the mixer-IF amplifier used should have < 14 dB. The problem was found to be noise in the oscillator spectrum 455 kHz away from the carrier. This was confirmed by using a klystron as the oscillator and a noise figure of about 13 dB was obtained.

The oscillator was returned to the manufacturer for modifications, but these resulted in only a slight improvement, and it was decided to accept the rather poor receiver noise figure.

The gain and phase tracking of the IF amplifiers are important. Rejection of the unwanted sideband depends on the cancellation of the first terms in equations (3) and (4) in section 2.2. Careful tests showed that a temperature swing of 5 °C resulted in a phase difference of 2° between the amplifier outputs. Further tests showed that most of this was due to the Clevite filters. This effect would be serious if the box were not temperature controlled. If we do further work with the system in which long-term stability is important, we may substitute filters with fewer poles and so reduce the phase shift. An alternative would be to add the two signals earlier in the chain.

The phase shifter consists of Q4 and Q5 and associated components. The signals are added in T4 and further amplified by limiting amplifiers A3 and A4.

The signal and reference square waves are differentiated in the phase detector (CD 2150) and used to drive the set-reset inputs of an AC toggled flip-flop. Complementary outputs are averaged and differentially drive an operational amplifier. The result is a linear phase-voltage transfer function of 10 mV per degree of phase difference. The output is bipolar about ground and extends to nearly ± 180°.

The temperature control system is shown in Figure 5. A thermistor is mounted on the mounting plate close to the microwave oscillator as this was thought to be the most sensitive component to temperature changes. The thermistor is connected in a bridge circuit which is set by means of the 5 K potentiometer to give zero output at 40 °C. If the thermistor is lower than this temperature, the bridge gives an output which is amplified and used to drive a current through the four 1 ohm resistors. This heats the plate and forms a closed loop system that keeps the plate at about ± 0.1 °C of 40 °C.

The horn antenna is shown in Figure 6. This horn has a 3 dB beamwidth of 110° in the H-plane and a beamwidth of 140° in the E-plane.

The power supply arrangement is shown in Figure 7. The transmitter/receiver is provided with two voltages from the control room — +28 V and -28 V. Internal power supplies generate the necessary voltages for the circuitry as shown in Figure 7.

The only other connections needed between the transmitter/receiver and the control room are a RG 58 cable for the 455 kHz reference and a wire for the phase detector output voltage.

4.22 Control and Display Unit

The digital part of the control unit is shown in Figures 8 and 9. Figure 8 shows the 455 kHz generator. The generator gives two 455 kHz outputs, one a reference waveform to the phase detector, and the other at a phase that may be set in 45° increments from 0° - 360°. This variable phase signal drives the transponder modulator.

A 3.64 MHz clock is fed into two ÷ 8 counters. One counter (providing the phase shifted signal) is reset at one of the other counters eight possible states. In this way the two 455 kHz outputs may be given relative phase of 0, 45, 90, etc., to 360°.

The particular thumbwheel switch controlling the phase shifter is selected by applying a voltage to its wiper. This voltage comes from the 55 Hz counter.

Figure 9 shows the 55 Hz counter and lamp driver and is fairly self-explanatory. In the manual mode, advance pulses are provided by a push-button driving a monostable circuit. The counter has 24 outputs, three of which are not used.

Figure 10 shows the circuits needed for one channel (with the exception of the diode driver). All these circuits are standard and need no explanation.

Figure 11 shows the modulator drive circuit. Twenty-one of these circuits are mounted on a printed circuit in the back of the control rack.

4.23 Transponder

Figure 12 is a drawing of a transponder. The horn is made out of aluminum sheet and the open end covered with thin teflon sheet. The modulator is a Somerset Radiation Model X-405. A sliding short circuit behind the diode is used to tune the modulator. The whole assembly is housed in a weatherproof cover. Four bolts attach the assembly to the telescope surface.

For measuring absolute distance, the phase center of the transponder is of interest. Reference [6] gives details of phase center measurements at X-band for horn antennas in which phase centers were determined to ± 0.002".

5.0 System Tests

5.1 Laboratory Tests

The system was tested in the laboratory prior to outdoor tests. A diode modulator was connected to the antenna port of the instrument via an attenuator and phase shifter to simulate both path loss and path length changes. The results obtained are shown in Figure 13. The wavelength of the carrier is 1", so a change in distance between transmitter and receiver of 1/2" gives a path length change of 1" and a phase change of 360°. As may be seen from Figure 13, the voltage output from the phase detector varied linearly with microwave phase shift. This test was a very convenient way of checking the calibration of the digital display meter.

5.2 Outdoor Tests

The complete measuring system was tested outside under conditions as close to actual operating conditions as possible. Five transponders were placed on I beams buried in the ground at distances from the transmitter corresponding to the 140' installation.

Some problems with illuminating the outer transponders resulted in a redesign of the transmitting antenna, but after this problem was solved the system worked well.

The transmitter was mounted on a slide rail and its position monitored with a dial gauge. Tests on the calibration of the instrument confirmed the laboratory results. The system noise was slightly higher than expected, being about 0.002" rms.

One important aspect of operation is the effect of reflections. Multipath reflections of the modulated return signal and also strong reflections of the carrier are both potential problem areas. Tests were done outdoors by moving large sheets of aluminum around the test area. No effects were noticed except near the direct line of sight.

Long-term stability tests showed drifts of about 0.025". This instability is attributed to atmospheric effects and thermal expansion or contraction in the transponder and feed horn. The atmospheric effects are discussed in section 6.0.

5.3 Tests on the 140'

The measuring system was installed on the 140' in May 1970. It was decided to use five transponders rather than the full 21. One transponder was mounted at the vertex; the remaining four were mounted at a radial distance of 625" from the telescope axis on a North-South line and on an East-West line.

The results are shown in Figure 14. The North transponder gave a weak return signal, probably due to misalignment. The results were found to be very repeatable. The apparent, sudden change in distance in the South transponder during the scan in hour angle is due to the 360° linear phase detector crossing over.

6.0 Stability Analysis

The long-term stability of the instrument is mainly determined by the following:

1) Microwave oscillator stability.
2) Variations in atmospheric refractive index.
3) Phase shifts in the coaxial cable supplying the modulating signal to the transponders.

The microwave oscillator stability required to give a maximum variation of .001" over a path length of 70 feet (840") will be approximately 1.2×10^{-6}, a fairly easy figure to achieve with a crystal oscillator.

The variations in refractive index at radio frequencies are discussed in reference [2].

The required relationship is

$$N = \frac{77.6}{T}\left[p + 4810\,\frac{e}{T}\right]$$

where $N = (n - 1)\ 10^6$, the scaled up refractive index.

n = refractive index.

T = absolute temperature.

p = total pressure in mb.

e = partial pressure of water vapor in mb.

This expression is considered to be good to 0.5 percent in N for frequencies up to 30 GHz.

The effects of the variability of the various terms may be calculated by partially differentiating the expression for N with respect to the different variables.

Assuming nominal values of

$$T = 300\ °K$$
$$p = 1000\ \text{mb}$$
$$e = 10\ \text{mb},$$

the various dependencies come out as follows:

$$\frac{\partial n}{\partial p} = +0.26\ \text{ppm/mb}$$

$$\frac{\partial n}{\partial T} = -1.0\ \text{ppm/°K}$$

$$\frac{\partial n}{\partial e} = +4.1\ \text{ppm/mb.}$$

From equation (6) we have

$$\phi = \frac{d}{\lambda_0}$$

which may be rewritten as $\phi = \dfrac{nf_0 d}{c}$

where c = velocity of light in vacuo,

so $\partial\phi = \dfrac{fd}{c}\,\partial n.$

A two-way path length of 120 feet gives

$$\partial\phi = \frac{11.8 \times 10^9 \times 120 \times 10^{-6}}{9.8 \times 10^8} \quad \text{rads/ppm change in refractive index.}$$

$$= 0.080°/\text{ppm}$$

which gives a distance measurement error of 1.100×10^{-4} inches/ppm.

A change of 10 mb in the partial water vapor pressure should therefore result in a change in distance measured of 0.0045". A change of 20 °C would result in 0.0022".

For long-term measurements, it should be fairly easy to measure atmospheric conditions and apply an automatic correction factor within the equipment.

The changes in phase of the modulating signal will now be considered. Reference [4] gives coefficients of phase stability for various coaxial cables. For RG 58, the coefficient is 480 ppm/°C. The length of cable involved is approximately 100 m and the velocity of propagation will be 2×10^8 m/sec. The wavelength in the cable will be

$$\frac{2 \times 10^8}{455 \times 10^3} = 440 \text{ m}$$

and the phase shift will be

$$\frac{1}{4.4} \times 360 = 82°.$$

The phase shift/°C will be 0.04°/°C.

A 20 °C temperature swing would give a distance error of approximately 0.001". This is a borderline effect: certainly any increase in the length of cable or increase in modulation frequency would lead to the introduction of an error that could not be ignored.

7.0 Future Development

Ways in which the instrument may be improved fall into three main categories:

a) Improved signal/noise.

b) More convenient form of transponder.

c) Improved stability.

a) Improved signal/noise.

Although the present system gives adequate performance in this respect, any increase in range or in the number of transponders would result in a decrease in signal/noise.

Rewriting equation (8) as

$$\frac{P_R}{P_T} = \frac{A_T^2 A_R^2}{R^4 \lambda^4}$$

gives an idea of what factors influence the S/N ratio.

If we wish the beamwidths of the antennas to remain the same, A_T, A_R, and λ must remain unchanged. This leaves only an increase in transmitter power.

Leakage at the carrier frequency would impose a limit on the transmitted power in the present system. The mixers operate at a LO power of approximately 10 mW. The circulator leakage will be approximately 30 dB and we may expect some reflections at the carrier frequency. Even taking the figure of 30 dB and assuming a perfectly matched feed, we would expect a carrier leakage of 1 mW with a transmitted power of 1 W. A good solution would be to increase the modulation frequency to about 30 MHz and have a narrow band filter ahead of the receiver to reject the carrier. The phase shift in this filter would have to be very stable. Inquiries from filter manufacturers indicate that a filter with a 20 MHz bandwidth at 11 GHz would have a phase-shift/temperature characteristic of about 0.10/°C. Distribution of this modulating frequency to the transponders would pose a problem in the present system. The phase-shift/temperature at 455 kHz was just acceptable: at 30 MHz it would be far too much. An alternative form of modulation that solves this problem is discussed later.

b) More convenient form of transponder.

The present method of supplying the modulating signal to the transponders would be rather cumbersome when using a large number of transponders. An attractive alternative to supplying the modulating signal is to use a modulated laser beam. A suitable form of antenna would be a dipole deposited on a substrate with a light-sensitive PIN diode at the center. The transponder would be about 4" in diameter and completely self-contained.

A possible alternative would be to use a slot antenna.

A transponder would be switched on merely by pointing the laser beam at it. The pointing accuracy required of the laser would be low.

Apart from the convenience of eliminating cables to the transponders, this scheme has the advantage of being able to use a high modulating frequency without the disadvantage of phase shift through a cable. The phase shift of the modulating signal due to atmospheric changes will be approximately $\frac{f_m}{f_0}$ x phase shift at the carrier frequency.

A HeNe laser with 10 mW output would be very suitable for the proposed system. With no collimation the spot size would be about 5 cm at a distance of 50 m.

With no focusing and 100 percent modulation this would give about 4 μA change in current through a typical PIN photo diode. What depth of modulation this would give to the microwave signal would depend on the diode used and the nature of the dipole, but rough calculations show the idea to be feasible.

The modulator for the laser beam presents no problems. A modulator capable of modulating 100 percent at 100 MHz may be purchased for less than $1,000.

c) Stability.

As has been shown, the main cause of instability is changes in the atmospheric refractive index. These changes may be calibrated out in a number of ways. One way would be to measure temperature and water vapor pressure and apply a correction to the readout electronics. Another possible method would be to have a transponder at a reference distance and to make the crystal-controlled carrier-oscillator voltage tunable. The oscillator could then be set so the readout read the reference distance and atmospheric changes would be calibrated out.

In the future we may do further work on measuring systems using different principles. During the last year or two measuring systems using a modulated light beam have been developed and these seem well suited to making measurements of the kind we need.

The principle used is very simple. A modulated light beam is produced, using either a laser or a light emitting diode, and the phase shift at the modulating frequency over the path provides a measure of the distance. The reflector used may be a simple, optical-corner reflector.

A modulating frequency of 100 MHz is possible, and with a lot of effort, a phase detection of 0.01°. This gives a measurement accuracy of about 0.1 mm.

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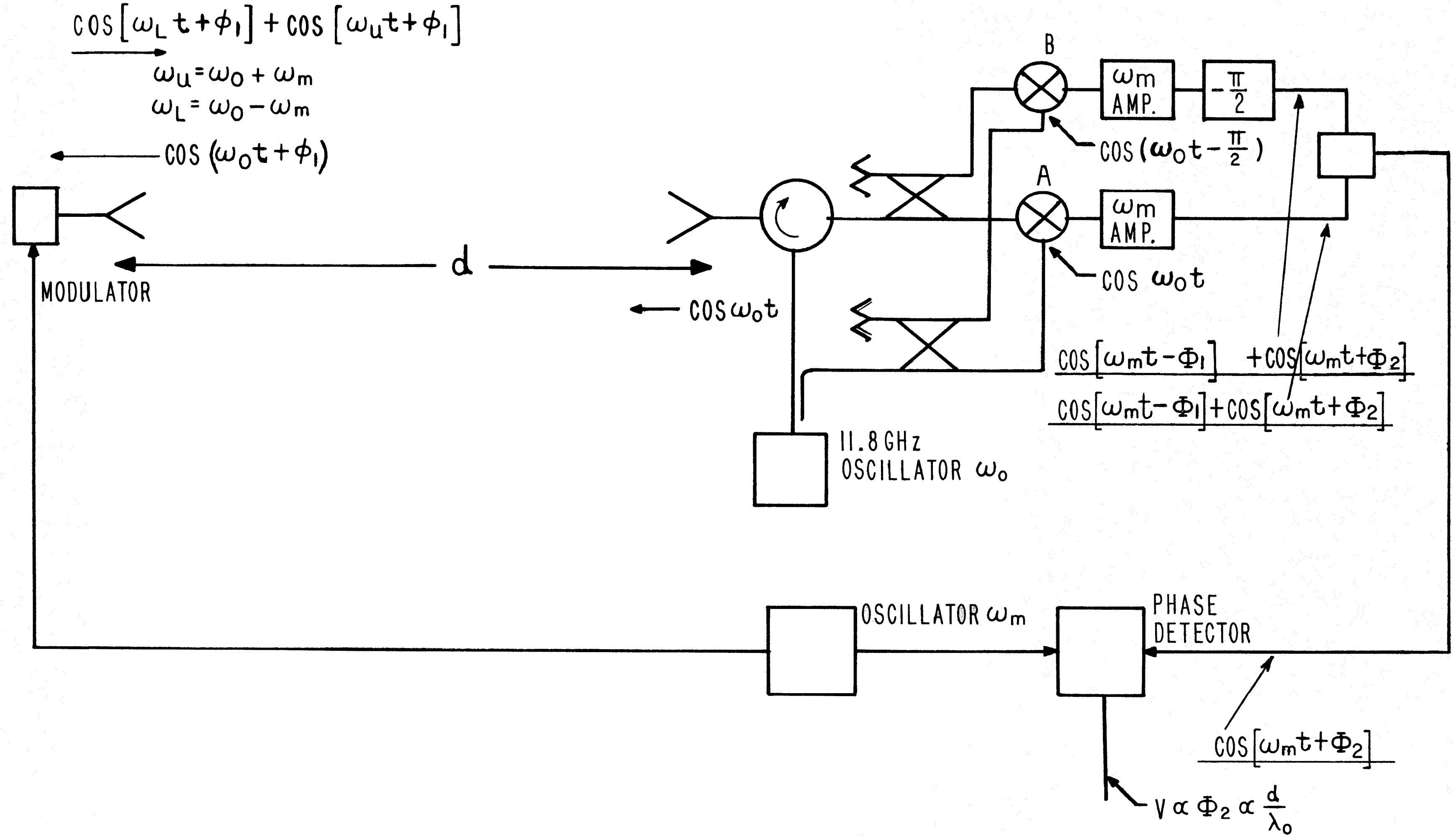


λ_o = WAVELENGTH OF CARRIER FREQUENCY

λ_L = WAVELENGTH OF L.S.B.

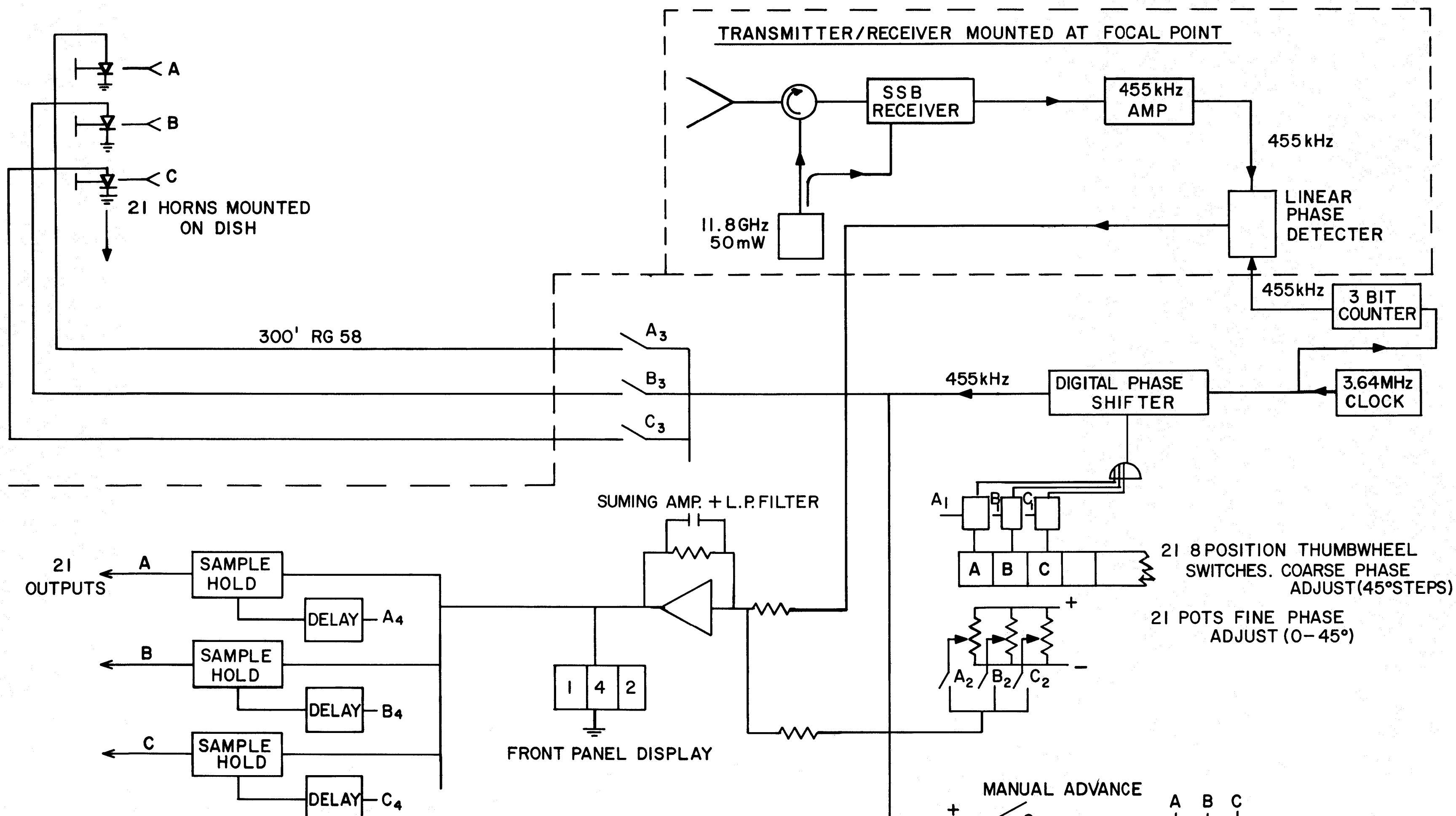
λ_u = WAVELENGTH OF U.S.B.

$$\Phi_1 = K_1 + \frac{2\pi d}{\lambda_L} + \phi_1$$

$$\Phi_2 = K_2 + \frac{2\pi d}{\lambda_u} + \phi_1$$

BASIC BLOCK DIAGRAM OF
MEASURING SYSTEM
FIG. I

Gray template: Condensed coffee (2.5)
KE 61-0699
500



ANTENNA MEASURING INSTRUMENT
BLOCK DIAGRAM
FIG. 2

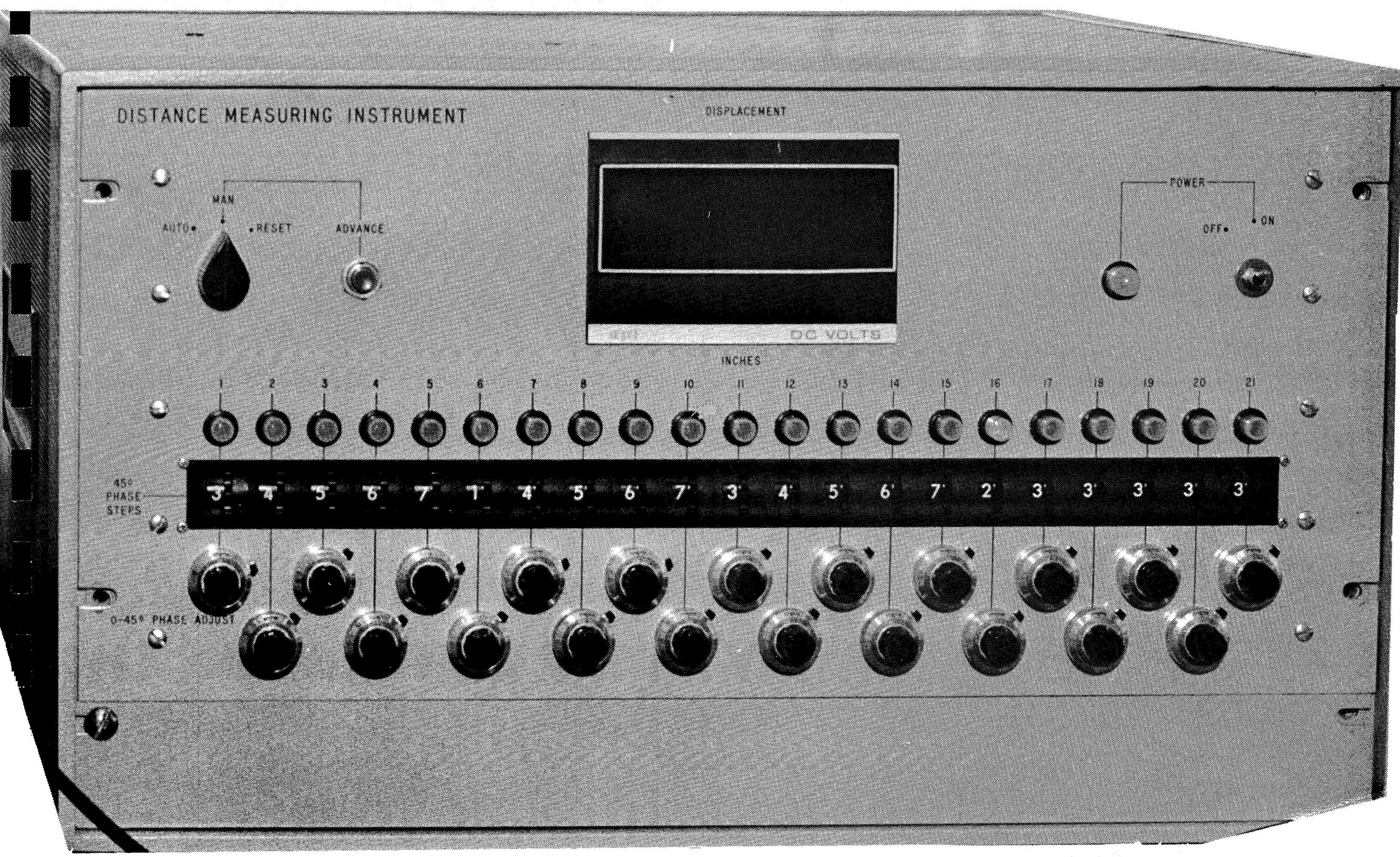
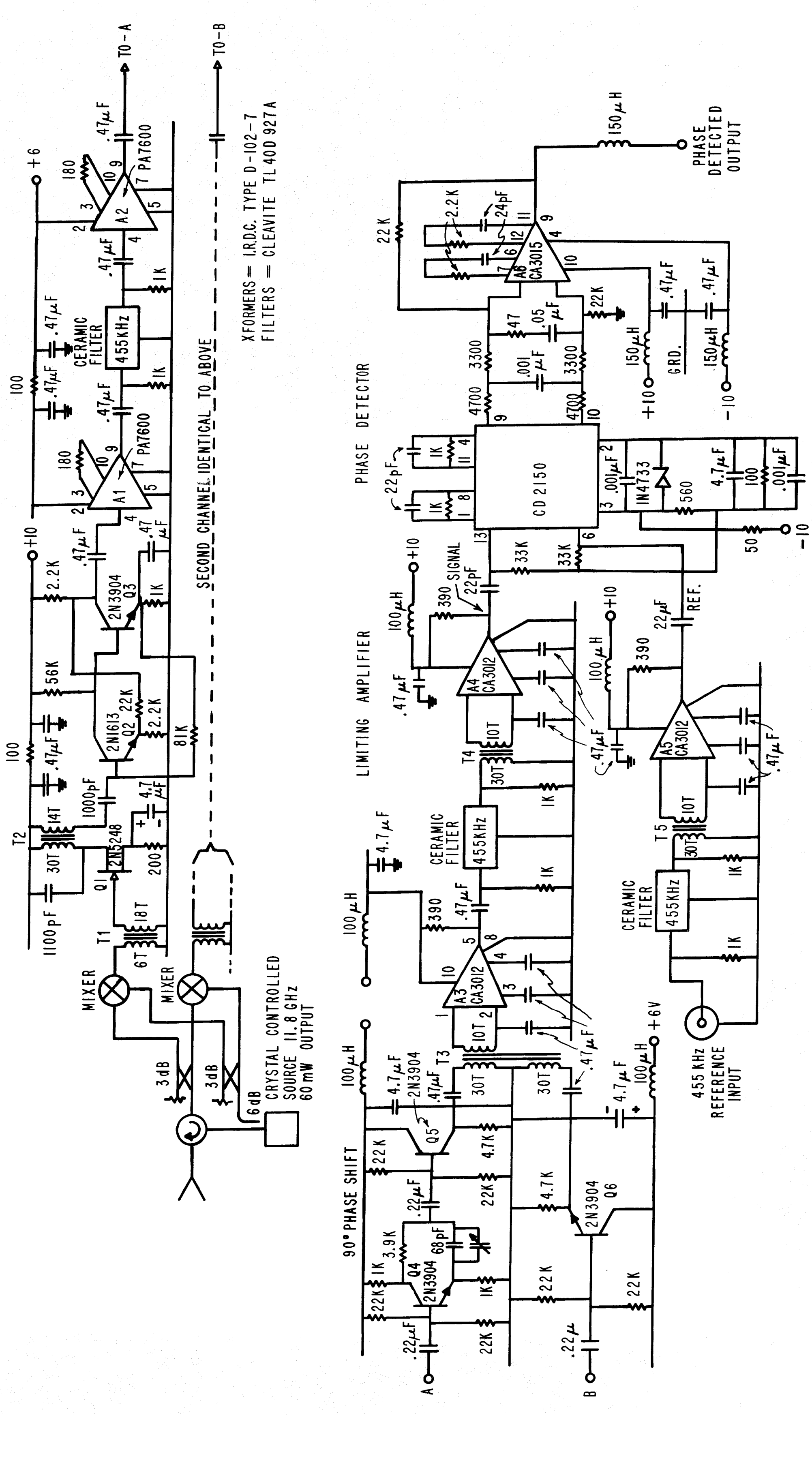
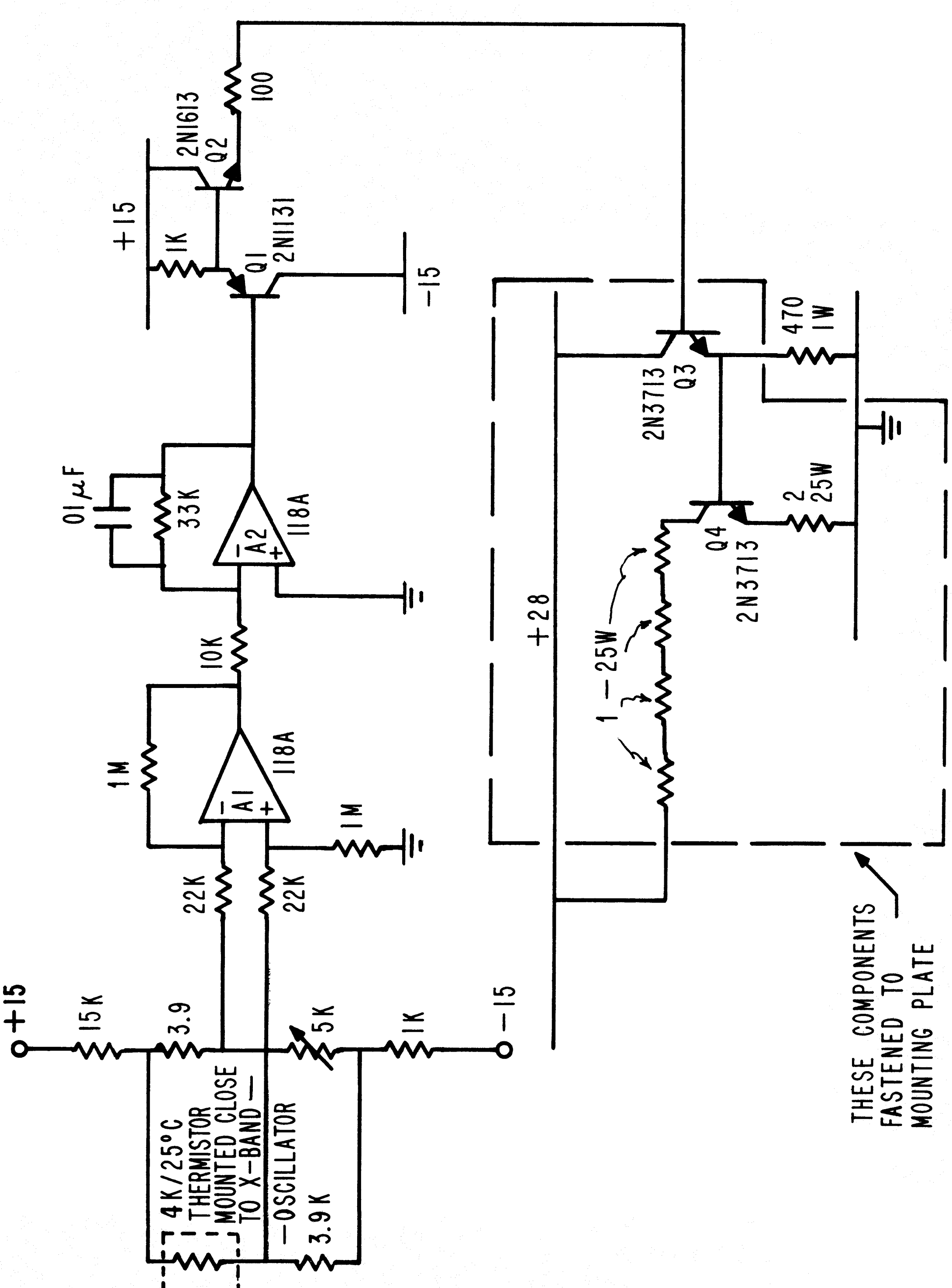

DISTANCE MEASURING INSTRUMENT
DISPLACEMENT
MAN
AUTO
RESET
ADVANCE
POWER
OFF
ON
DC VOLTS
INCHES
1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18 19 20 21
45° PHASE STEPS
0-45° PHASE ADJUST

Figure 3

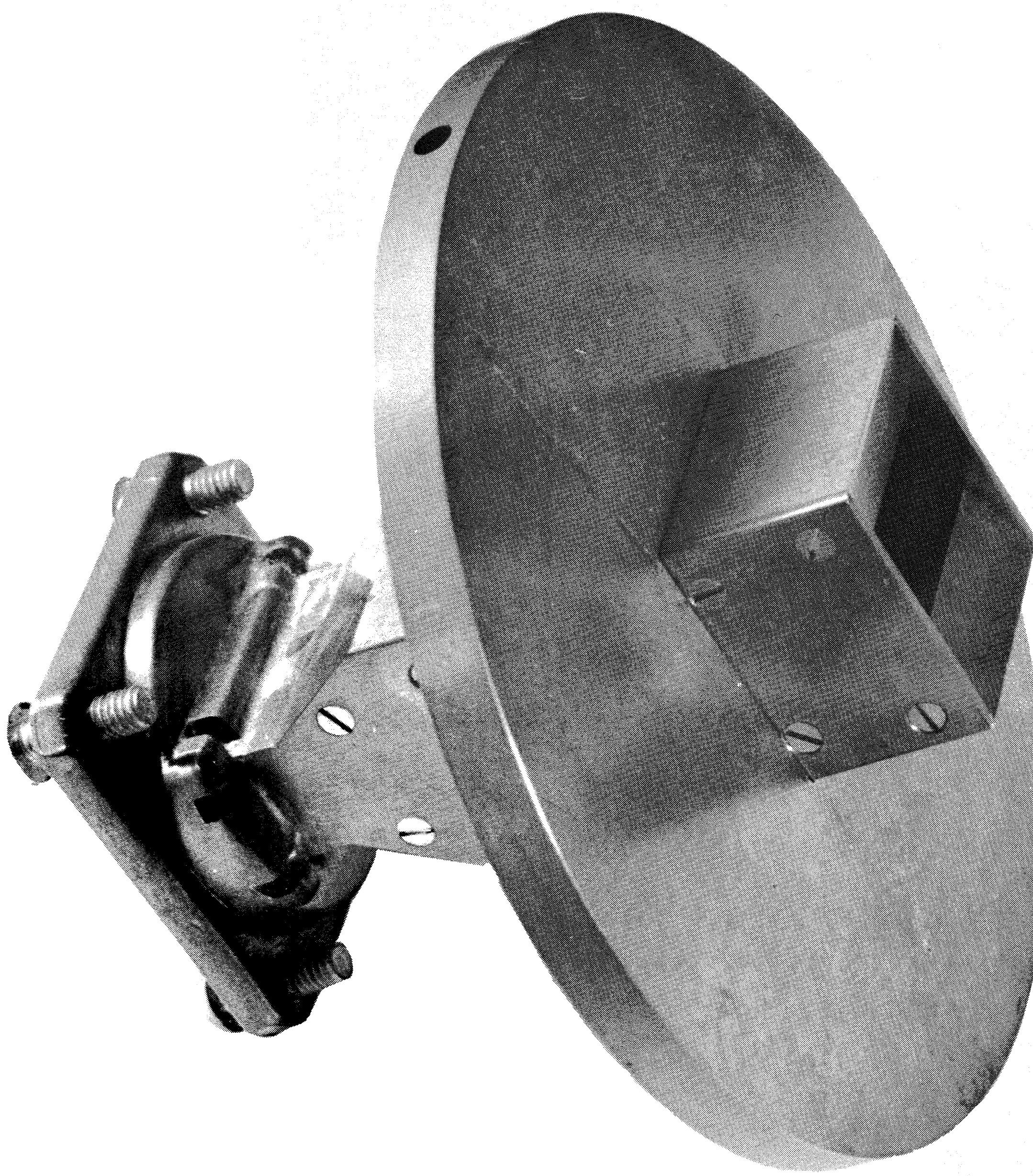


ANTENNA MEASURING SYSTEM
TRANSMITTER / RECEIVER
FIG. 4

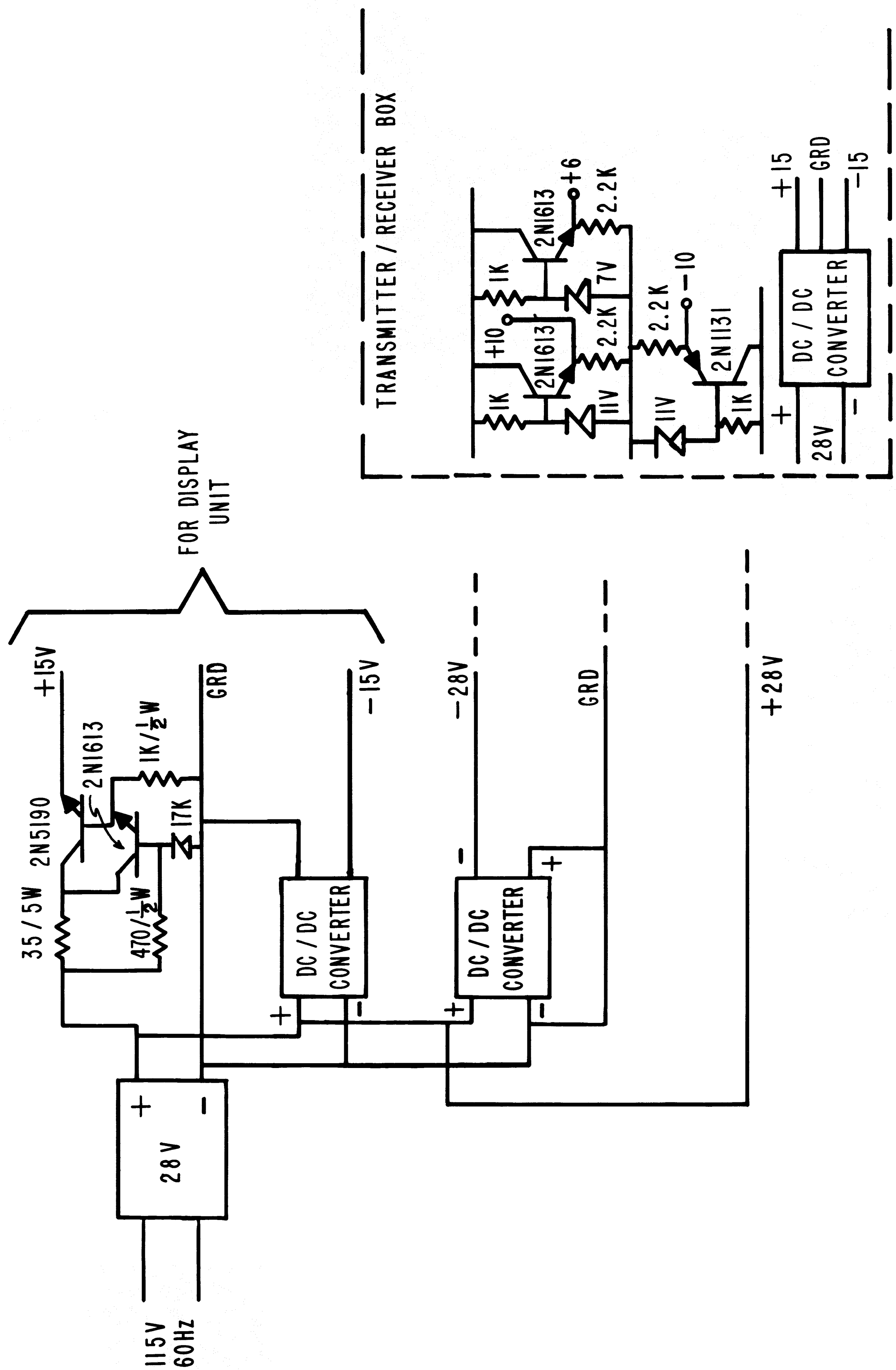


TRANSMITTER / RECEIVER TEMPERATURE CONTROL

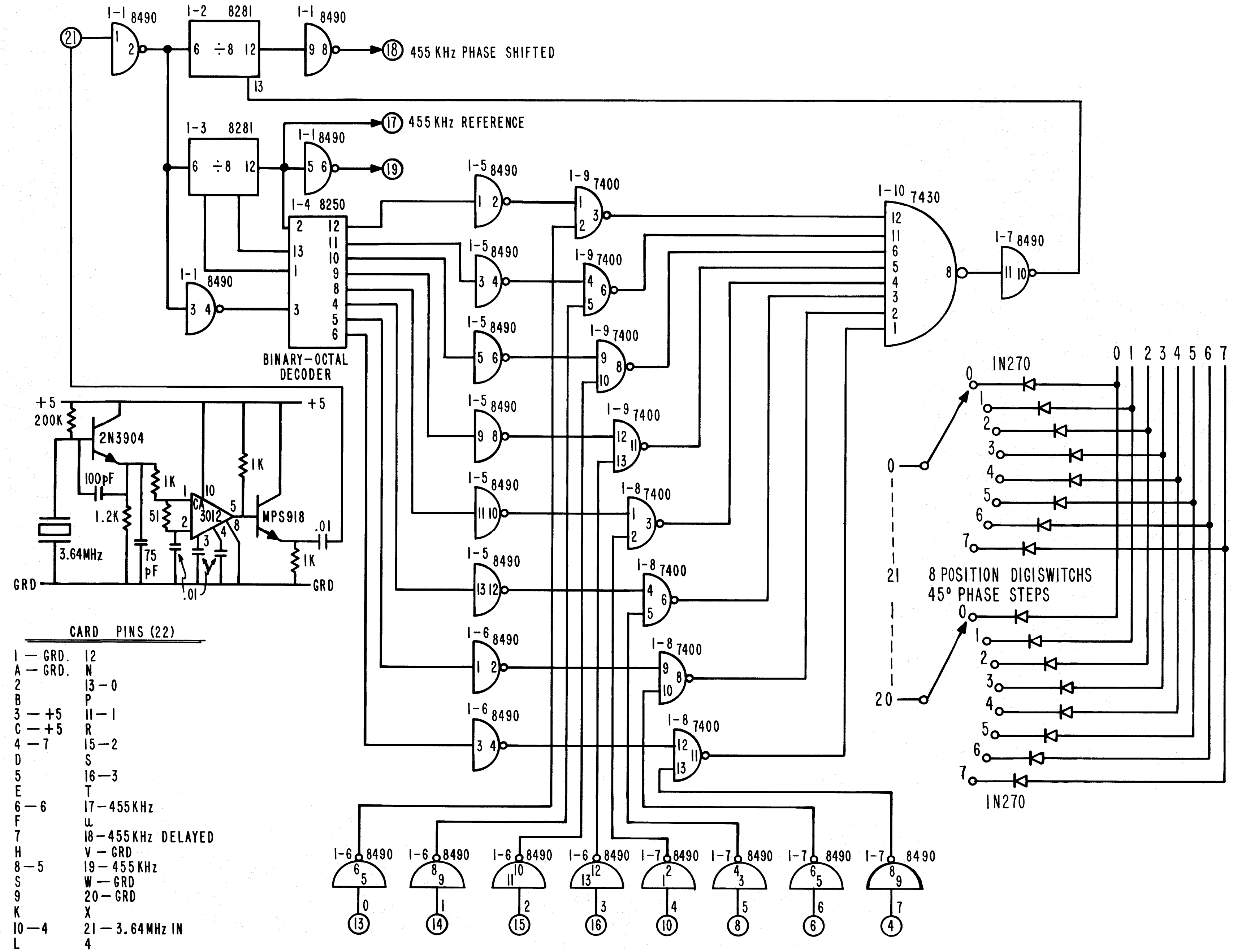
FIG. 5



HORN
FIG.6



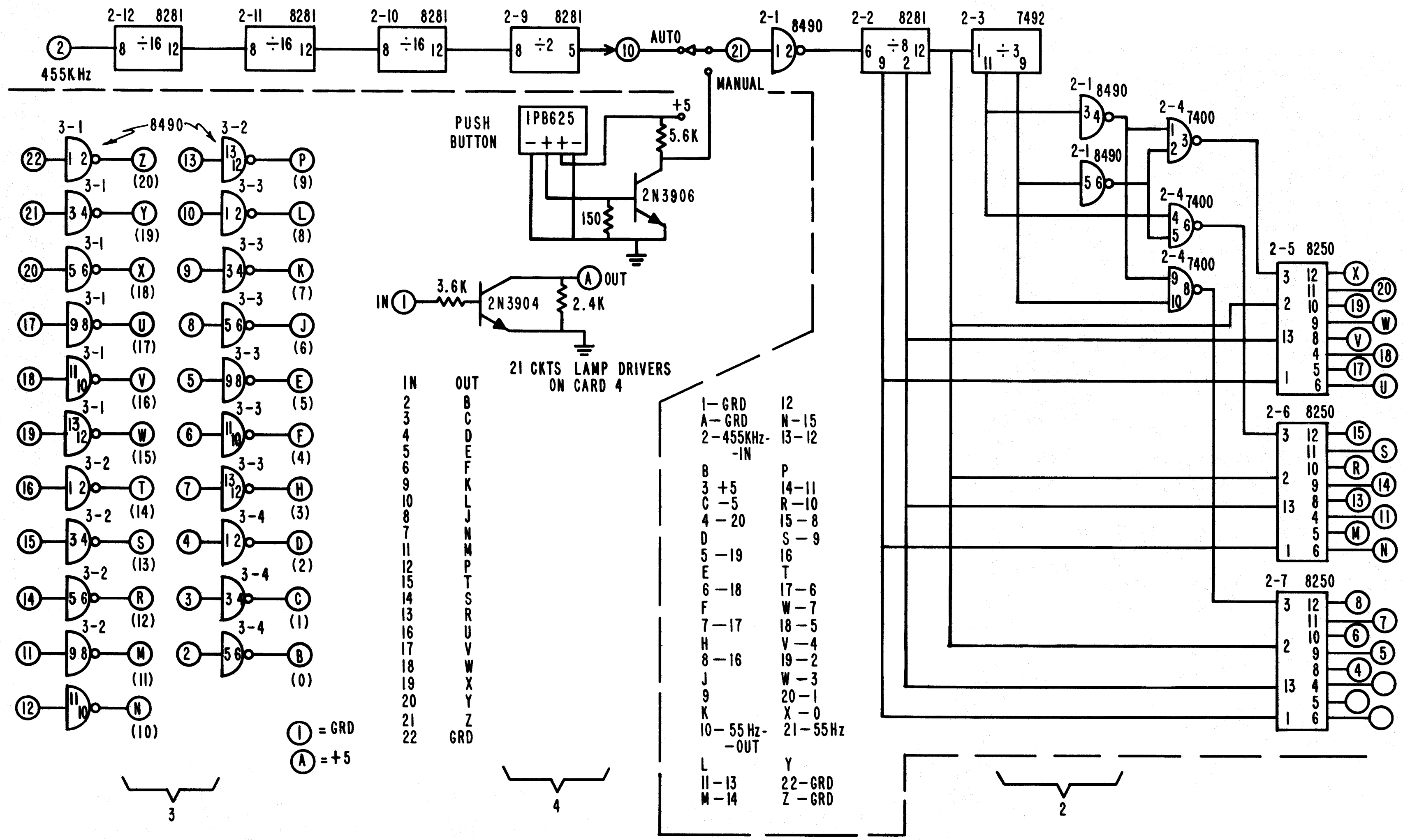
POWER SUPPLIES
FIG. 7



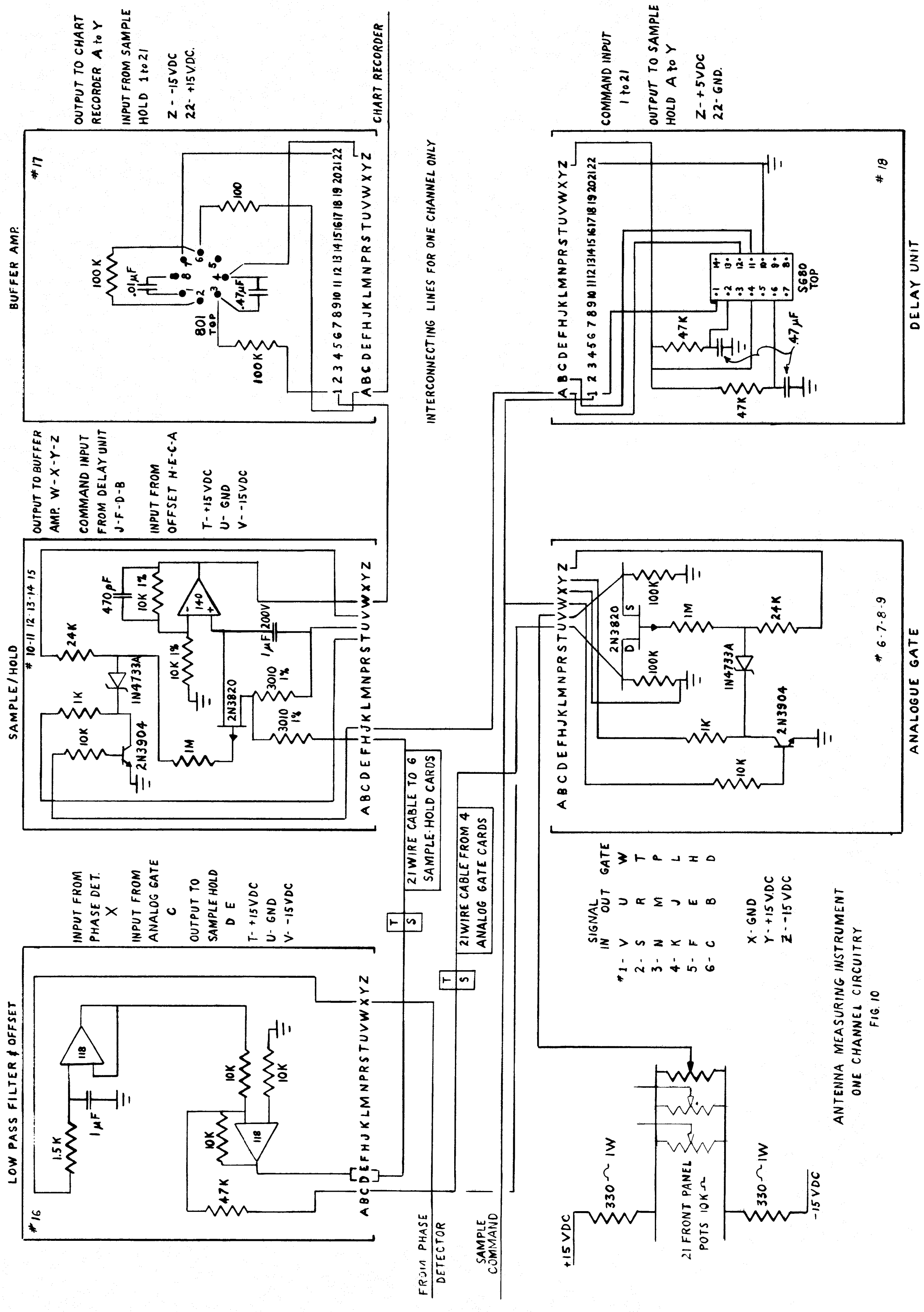
CARD PINS (22)	
1 — GRD.	12
A — GRD.	N
2	13 — 0
B	P
3 — +5	11 — 1
C — +5	R
4 — 7	15 — 2
D	S
5	16 — 3
E	T
6 — 6	17 — 455KHz
F	U
7	18 — 455KHz DELAYED
H	V — GRD
8 — 5	19 — 455KHz
S	W — GRD
9	20 — GRD
K	X
10 — 4	21 — 3.64MHz IN
L	4
11	22 — GRD
M	Z — GRD

CARD 1

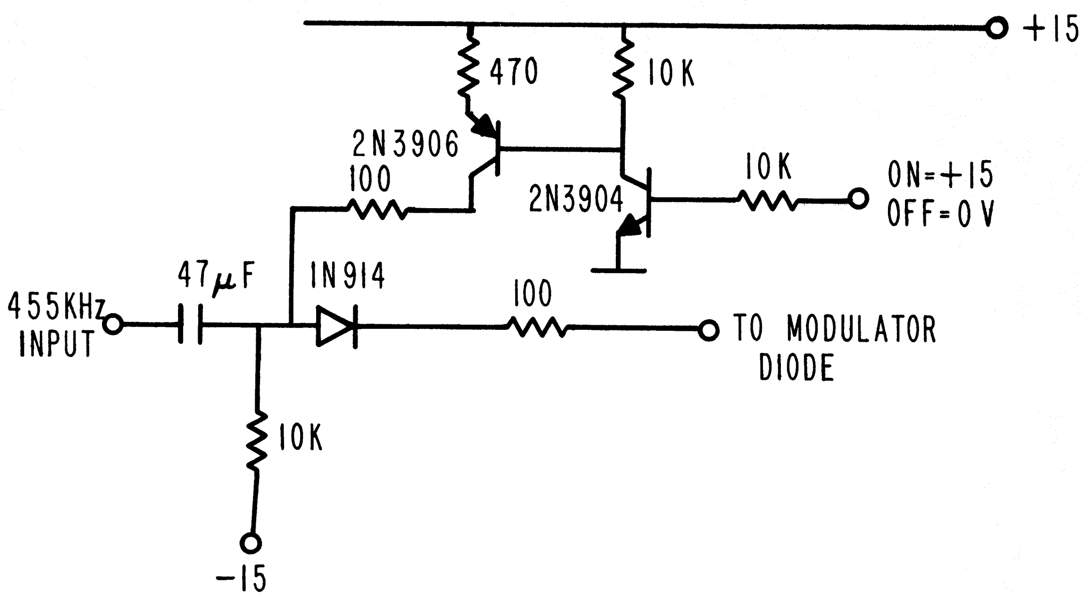
455 KHz GENERATOR
FIG. 8



55 Hz COUNTER & LAMP DRIVER
FIG. 9

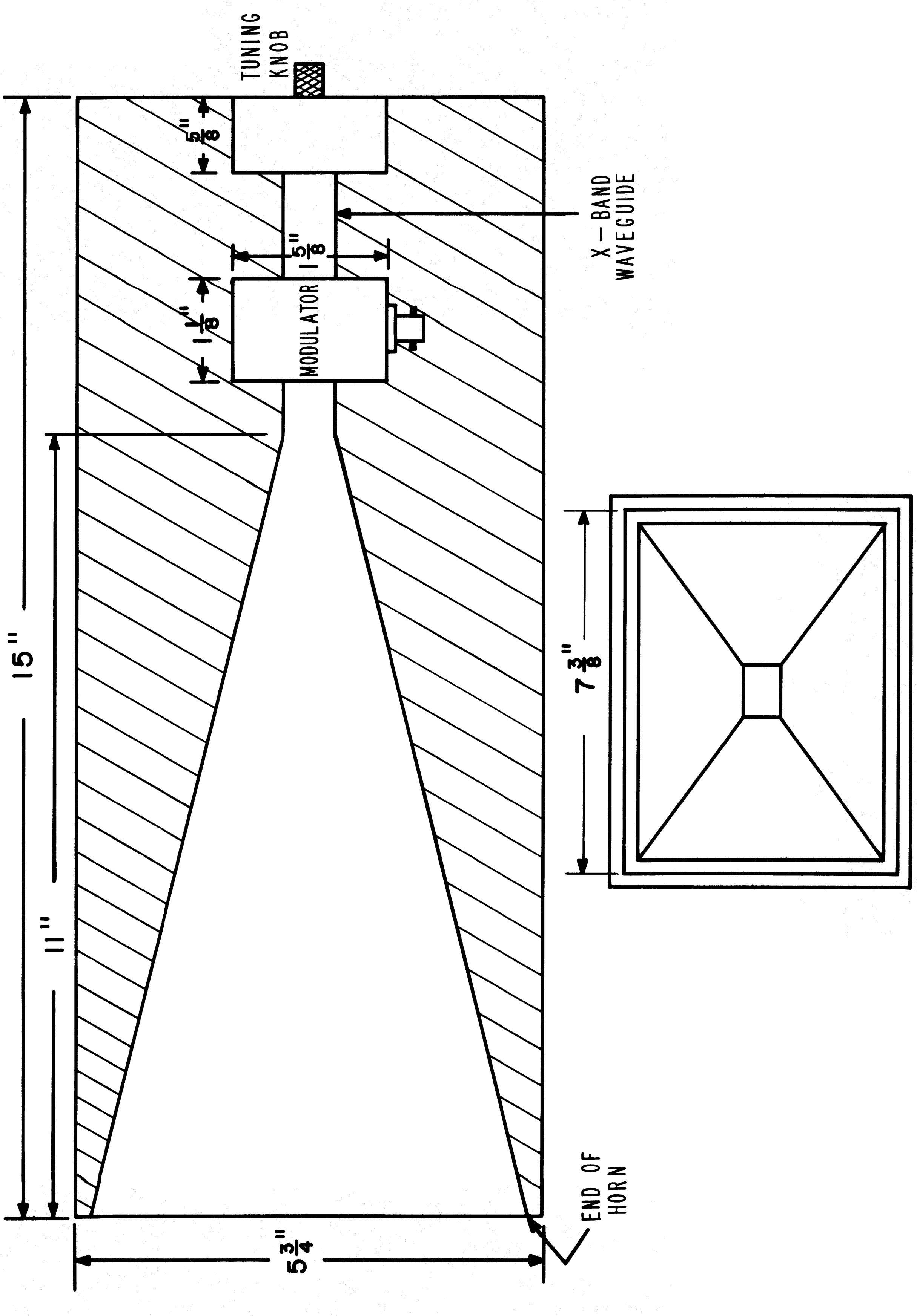


ANTENNA MEASURING INSTRUMENT
ONE CHANNEL CIRCUITRY
FIG. 10

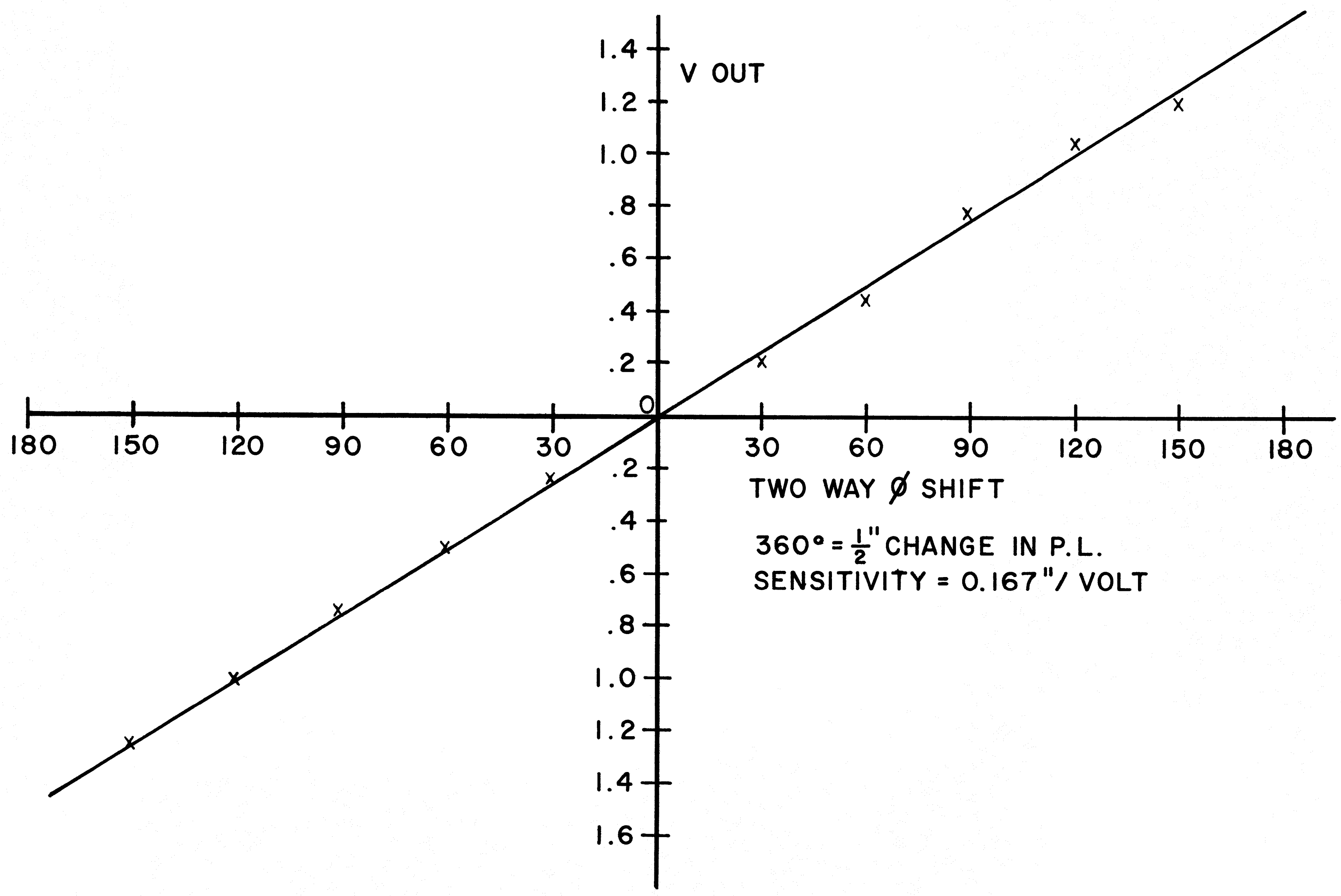


MODULATOR DRIVER CIRCUIT

FIG. 11



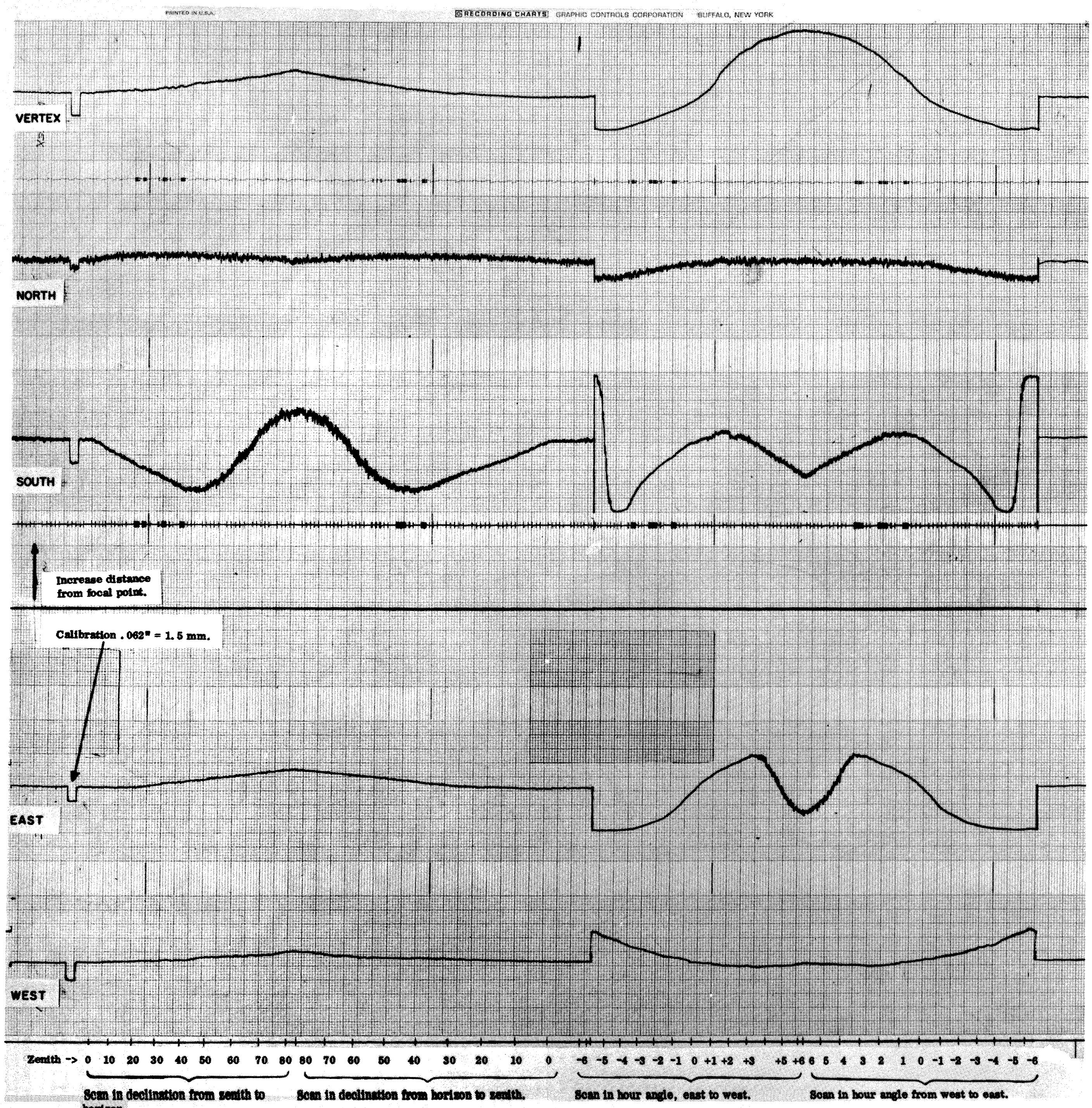
TRANSPONDER
FIG. 12



PHASE LINEARITY TEST
FIG. 13

ANTENNA MEASURING INSTRUMENT RESULTS ON 140-FOOT TELESCOPE — MAY 21, 1970

John M. Payne



RESULTS OF 140' TEST

FIG. 14