AFCRL 94

STUDY OF POLARIZATION MODULATION TECHNIQUES

Joseph E. Ferris

Final Report

Contract No. AF 19(604)-4988

February 1961

Prepared
for

Electronics Research Directorate
AIR FORCE CAMBRIDGE RESEARCH LABORATORIES
OFFICE OF AEROSPACE RESEARCH
UNITED STATES AIR FORCE
BEDFORD, MASSACHUSETTS

MELPAR INC
A SUBSIDIARY OF WESTINGHOUSE AIR BRAKE COMPANY
3000 ARLINGTON BOULEVARD FALLS CHURCH, VIRGINIA

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ABSTRACT

Several methods are described for generating a multipolarized wave for ECM jamming purposes. These methods generally involve the excitation of two orthogonally oriented linear antennas with a varying phase and amplitude. Discussions and mathematical analyses of the performances of these systems are presented and some experimental data is included. Curves of theoretical relationships are plotted to illustrate performance characteristics.

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1.0 INTRODUCTION

The purpose of this program has been to conduct both an analytical and experimental investigation of techniques for rapidly varying the polarization of airborne transmitting antennas.

Some of the techniques investigated deal with the generation of a polarization-modulated wave by varying the orientation angle of a linearly polarized wave at a rapid cyclic rate. A polarization-modulated wave offers considerable advantage over conventional fixed polarizations, when applied to jamming systems. A linearly polarized jamming signal can be effectively rejected by using a cross-polarized linear receiving antenna; likewise, a circularly polarized jamming signal can be rejected by using an oppositely sensed circularly polarized receiving antenna. A polarization-modulated jamming signal, however, contains components of several polarizations to at least some of which a receiver having a fixed-polarization is vulnerable.

In addition to techniques for rapidly varying the polarization, techniques were also investigated which permitted the antenna polarization to be rapidly adjusted. As a result of this investigation, it was felt that future work should be directed toward obtaining improved electronic techniques as they are discussed in the body of this report.

2.0 GENERAL DISCUSSION

2.1 Elliptical Polarization

A plane electromagnetic wave at one frequency may have any polarization; this polarization must always be some form of elliptical polarization. The wave is elliptically polarized when the extremity of the electric vector describes an ellipse in a plane perpendicular to the direction of propagation, making one complete revolution during one period of the wave. If the electric vector (viewed approaching the observer) appears to rotate clockwise, the polarization ellipse is said to have left-hand sense; if it appears to rotate counterclockwise, the polarization ellipse is said to have right-hand sense.

The polarization of a wave is specified when its axial ratio (AR), tilt angle (Ψ), and sense (right- or left-hand) are known. Several examples are:

Vertical Linear Polarization

AR = 0

Ψ = 90°

sense = indeterminate

Right-Hand Circular Polarization

AR = 1

Ψ = 90°

sense = right hand

Right-Hand Elliptical Polarization

AR = .707

Ψ = 45°

sense = right hand

In general, the polarization of a wave may be specified when both the relative amplitude and phase of any two cross-polarized elliptical field components are known. However, it is simpler, in practice, to consider only orthogonal linear or oppositely sensed circular fields (which are extreme cases of an elliptically polarized wave) to specify the polarization of a wave. If the polarization is to be specified in terms of linear fields, two perpendicular reference axes OX and OY are chosen in the plane of the polarization ellipse and are perpendicular to the direction of propagation (figure 1). The field components along these axes are specified by:

$$E_x = A \sin(\omega t + \phi_x)$$

and

$$E_y = B \sin(\omega t + \phi_y). \qquad (2.1)$$

where "E_x" and "E_y" are the instantaneous magnitudes of the "X" and "Y" field components, respectively, and ϕ_x and ϕ_y are the phase angles of the two components, respectively. Figure 2 shows the various states of polarization as a function of the amplitude ratio ($\frac{B}{A}$) and the relative phase difference ($\phi = \phi_y - \phi_x$) of the two orthogonal linear field components. Because the polarization can also be specified by two oppositely sensed circular polarizations, the field components can be specified by:

$$E_{LH} = C\, e^{-j(wt + \phi_{LH})} \qquad (2.3)$$

and

$$E_{RH} = D\, e^{j(wt + \phi_{RH})}\ . \qquad (2.4)$$

Figure 3 shows the various states of polarization as a function of amplitude ratio ($\frac{C}{D}$) and phase difference ($\phi = \phi_{LH} - \phi_{RH}$).

The power received by an elliptically polarized receiving antenna in the field of a plane elliptically polarized wave is obtained from:

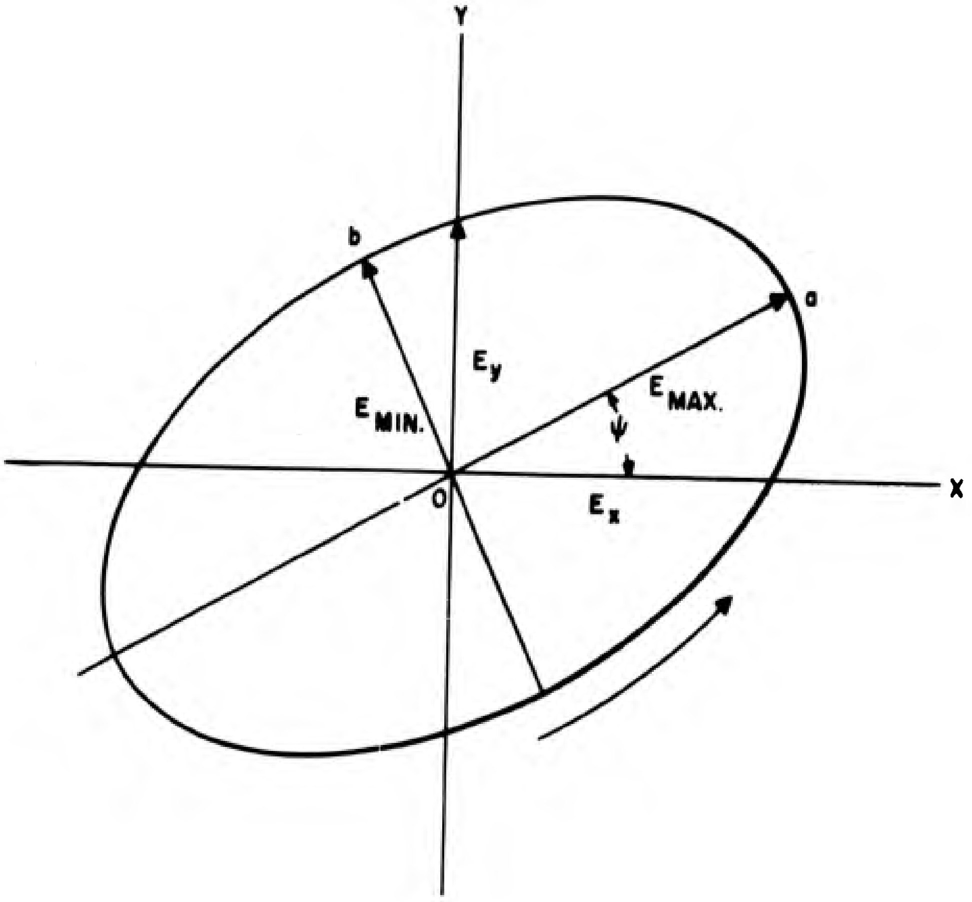

Figure 1. Polarization Ellipse

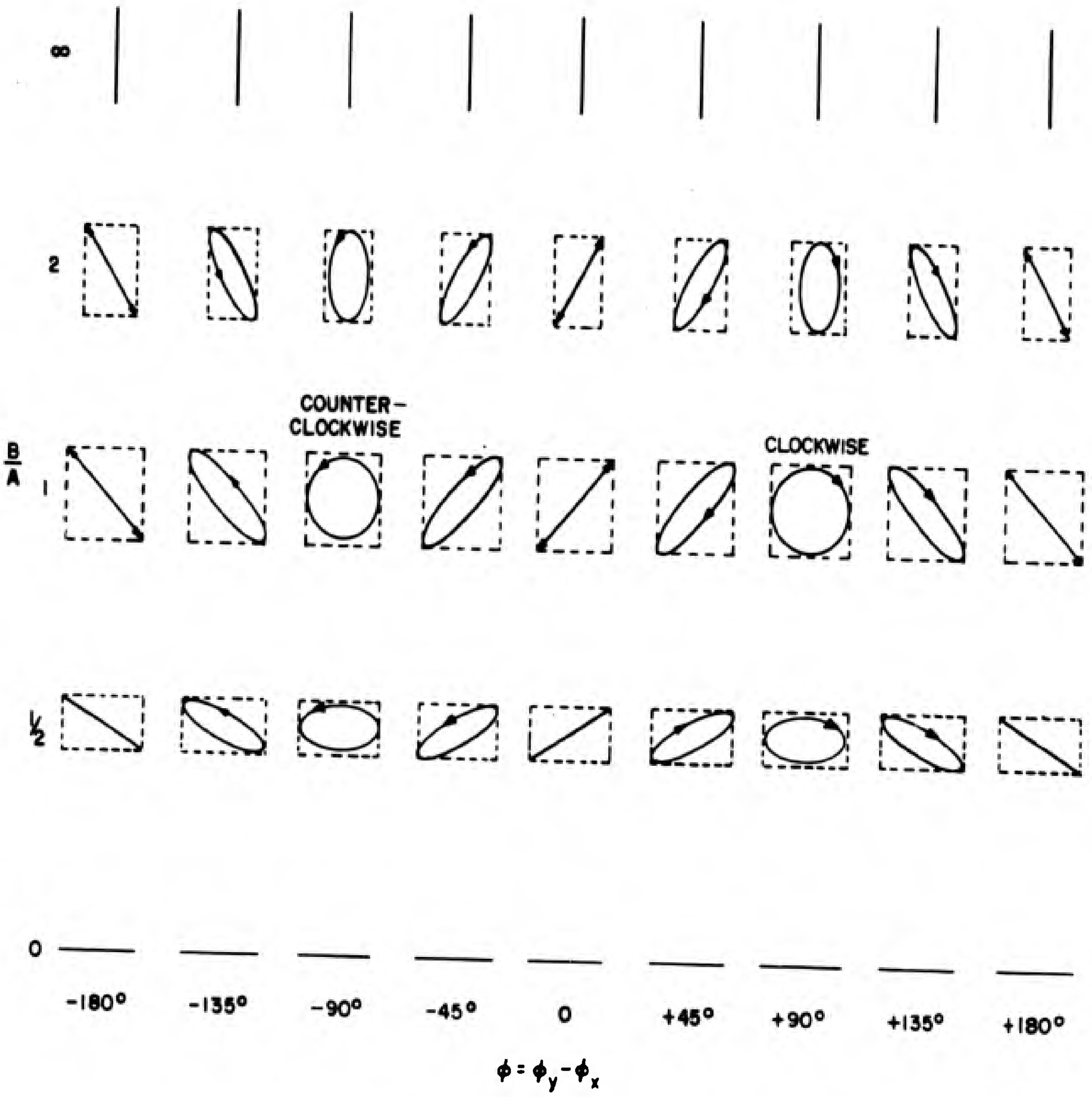

Figure 2. Polarization Ellipse as Function of Relative Amplitudes and Phases of Two Linear Field Components

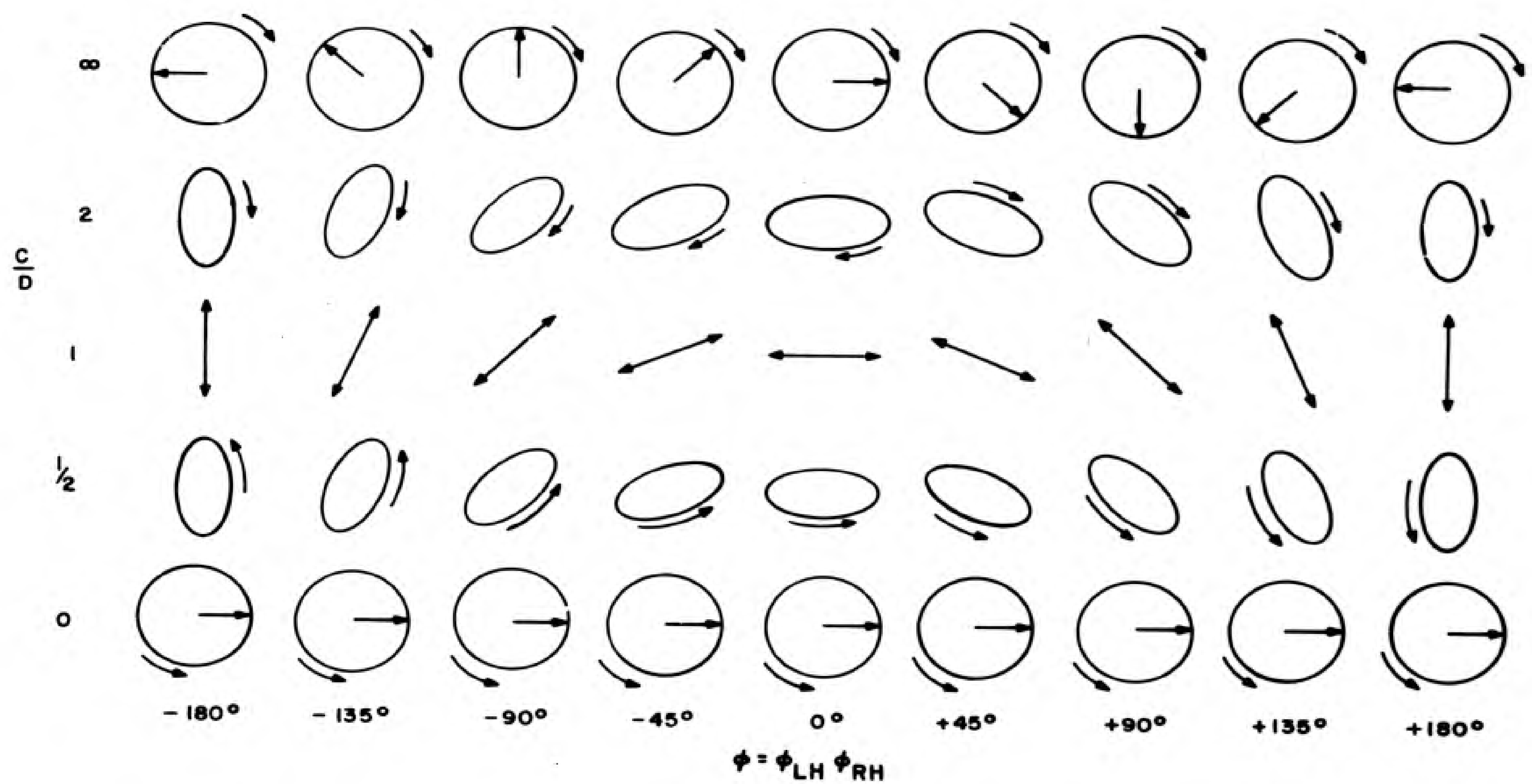

Figure 3. **Polarization Ellipse as Function of Relative Amplitudes and Phases of Two Circular Field Components**

$$W = KP \tag{2.5}$$

and

$$P = \frac{(1 \pm AR_1\ AR_2)^2 + (AR \pm AR_2)^2 + (1 - AR_1^2)\ (1 - AR_2^2) \cos 2\theta}{(1 + AR_1^2)\ (1 + AR_2^2)} \tag{2.6}$$

where

K = constant dependent upon the power density of the wave $\left(\frac{E^2}{240\pi}\right)$ and the effective aperture of the receiving antenna in the direction of propagation of the incident wave.

P = Polarization factor

AR_1 = axial ratio of elliptically polarized wave

AR_2 = axial ratio of elliptically polarized receiving antenna in the direction of the incident wave

θ = angle between maximum of wave polarization ellipse and maximum of the receiving antenna polarization ellipse.

The (+) sign is used when the sense of rotation of the wave polarization vector and the receiving antenna polarization vector are the same. The (-) sign is used if the two polarization vectors are of opposite sense.

A receiving antenna will receive maximum power, as a function of polarization, when the polarizations of the wave and antenna are identical in terms of axial ratio, tilt angle, and sense. If the polarizations are different, the antenna will receive less power, depending upon the particular values of AR_1, and AR_2, θ and sign ($\pm$). Several examples are:

(1) For optimum received power

$$AR_1 = AR_2 = AR$$

$$\theta = 0°$$

$$\text{sign} = + \text{ (same sense)}$$

$$\text{and } P = 2$$

Thus, the maximum value of $P = 2$.

Therefore, $W = K2$

(2) For crossed linear polarizations

$$AR_1 = AR_2 = 0 \text{ (linear)}$$

$$\theta = 90°$$

$$\text{sign} = \text{indeterminate}$$

$$\text{and } P = 0$$

Therefore, the minimum value of P is zero.

$$\text{and } W = K(0) = 0$$

(3) For linear and elliptical polarizations

$$AR_1 = AR \text{ (elliptical)}$$

$$AR_2 = 0 \quad \text{(linear)}$$

$$\theta = \theta_1$$

$$\text{sign} = \text{indeterminate}$$

$$\text{and } P = 1 + \frac{(1 - AR^2)}{(1 + AR^2)} \cos 2\theta_1.$$

(4) For elliptical and circular polarizations

$$AR_1 = AR \text{ (elliptical)}$$

$$AR_2 = 1 \text{ (circular)}$$

$$\theta = \text{indeterminate}$$

$$\text{and } P = \frac{(1 \pm AR)^2}{(1 + AR^2)}$$

Therefore, from this discussion, it can be seen that to decouple $(P = 0)$ two elliptical polarizations of the same axial ratio $(AR_1 = AR_2)$, they must be both of opposite sense (-) and orthogonally oriented $(\theta = 90°)$.

2.2 Variable Polarization

Methods of varying the state of polarization (with time) may be classed in terms of relative phase and amplitude differences between two cross-polarized linear or circular field components as:

1. Phase difference and amplitude ratio variable.
2. Phase difference variable, amplitude ratio fixed.
3. Phase difference fixed, amplitude ratio variable.

If the cross-polarized components are orthogonal linear polarizations $(AR = 0)$, the following conditions exist for each of the classes above:

Class I

(Phase difference and amplitude ratio variations) in general permits all possible states of polarization of figure 2 to be produced.

Class 2

(Phase difference variations) permits the ellipses of any one horizontal row of figure 2 to be produced.

Class 3

(Amplitude ratio variations) permits the ellipses of any one vertical column in figure 2 to be produced. If the cross-polarized components are circular of opposite sense (AR = 1) the following conditions exist for each of the above classes:

Class 1

(Phase difference and amplitude ratio variations) again permits all possible states of polarization of figure 3 to be produced.

Class 2

(Phase variations) permits the ellipses of any one horizontal row of figure 3 to be produced.

Class 3

(Amplitude ratio variations) permits the ellipses of any one vertical column in figure 3 to be produced.

Each of the classes above has advantages and disadvantages which are discussed below:

Class 1

Variations are useful in situations where it is desirable to optimize the polarization of a receiving or transmitting antenna. In this application, a pair of oppositely sensed circular antennas offer advantages over cross-polarized linear antennas. In the circular case, the tiit angle of the resultant polarization ellipse is a function of relative phase difference alone and axial ratio is a function of relative amplitude alone, while in the linear case, tilt angle and axial ratio are functions of both relative phase and amplitude (see figures 2 and 3).

Class 2

Variations give rise to an interesting state of polarization when the cross-polarized components are equal in amplitude and relative phase continuously varies thru 360°. During one complete cycle of phase change, polarizations shown in the center row of figures 2 and 3 are generated. It has been proven experimentally, as part of this study, that a signal transmitted with this type of polarization would contain components of all polarizations and a receiving antenna would receive some signal no matter what its polarization.

Class 3

Variations are used in situations requiring limited range of polarization modulation. One use, in particular, would be switching between two crossed polarized antennas, here relative amplitude and, thus, transmitted polarization is changed instantaneously. In general, however, the variations of class 2 or 3 are more applicable to practical systems intended for polarization modulation.

3.0 HIGH-SPEED ROTATING QUARTER-WAVE PLATE

3.1 Polarization Variation With A Dielectric Quarter-Wave Plate

The use of a dielectric quarter-wave plate inserted in a waveguide to convert linear to circular polarization is well known. The dielectric plate creates a relative phase difference between two orthogonal modes and, in particular, the differential phase shift is 90 degrees for a quarter-wave plate.

By mounting such a quarter wave plate in a circular waveguide propagating the TE_{11} mode, at an angle β * to the incident linearly polarized wave, two orthogonal components are generated which are parallel with and perpendicular to the axis of the plate. These two components have amplitudes proportional to $\cos\beta$ and $\sin\beta$, and a phase differential of 90°. An elliptically polarized wave results whose ellipticity and sense are dependent on the angle.

A theoretical and experimental investigation was conducted to determine the tilt angle and axial ratio of the resultant polarization ellipse using a linearly polarized wave incident on a quarter-wave plate, as functions of the angle β (between quarter-wave plate and incident linear polarization). The theoretical

* Is positive in the ccw direction when viewed so the direction of polarization is toward the observer.

investigation utilized the polarization chart described in appendix A. The results are shown in figures 4 and 5. The ellipses center along a line parallel with the linear polarization (incident on the quarter-wave plate) and range from right-hand circular through the original linear to left-hand circular.

A more versatile system can be obtained by feeding the circular waveguide with orthogonal modes of equal amplitude and 90° out of phase (circular polarization). The wave in the guide can be resolved into two equal amplitude components, one parallel and the other perpendicular to the quarter-wave plate. The parallel component is shifted 90° in phase, thus causing the two components to have either 0° or 180° phase difference. This results in a linear polarization oriented at $\pm 45°$ relative to the quarter-wave plate. By rotating the quarter-wave plate, the angle at which the linear polarization emerges from the waveguide is made to change thru 180°. The results of experimental measurements, using a circular waveguide fed with circular polarization, are compared with the theoretical predictions in figure 6. The experimental measurements confirm the theoretical prediction that a rotating linearly polarized wave results when a circularly polarized wave is passed through a rotating quarter-wave plate.

The feasibility of generating a satisfactorily modulated polarization, using a rotating quarter-wave plate, has been shown; therefore, further investigations were centered about rotation techniques and development of a quarter-wave plate with satisfactory electrical characteristics.

3.2 Rotational Techniques

The accepted method is to have the quarter-wave plate fixed with respect to a section of waveguide and to rotate the waveguide mechanically.

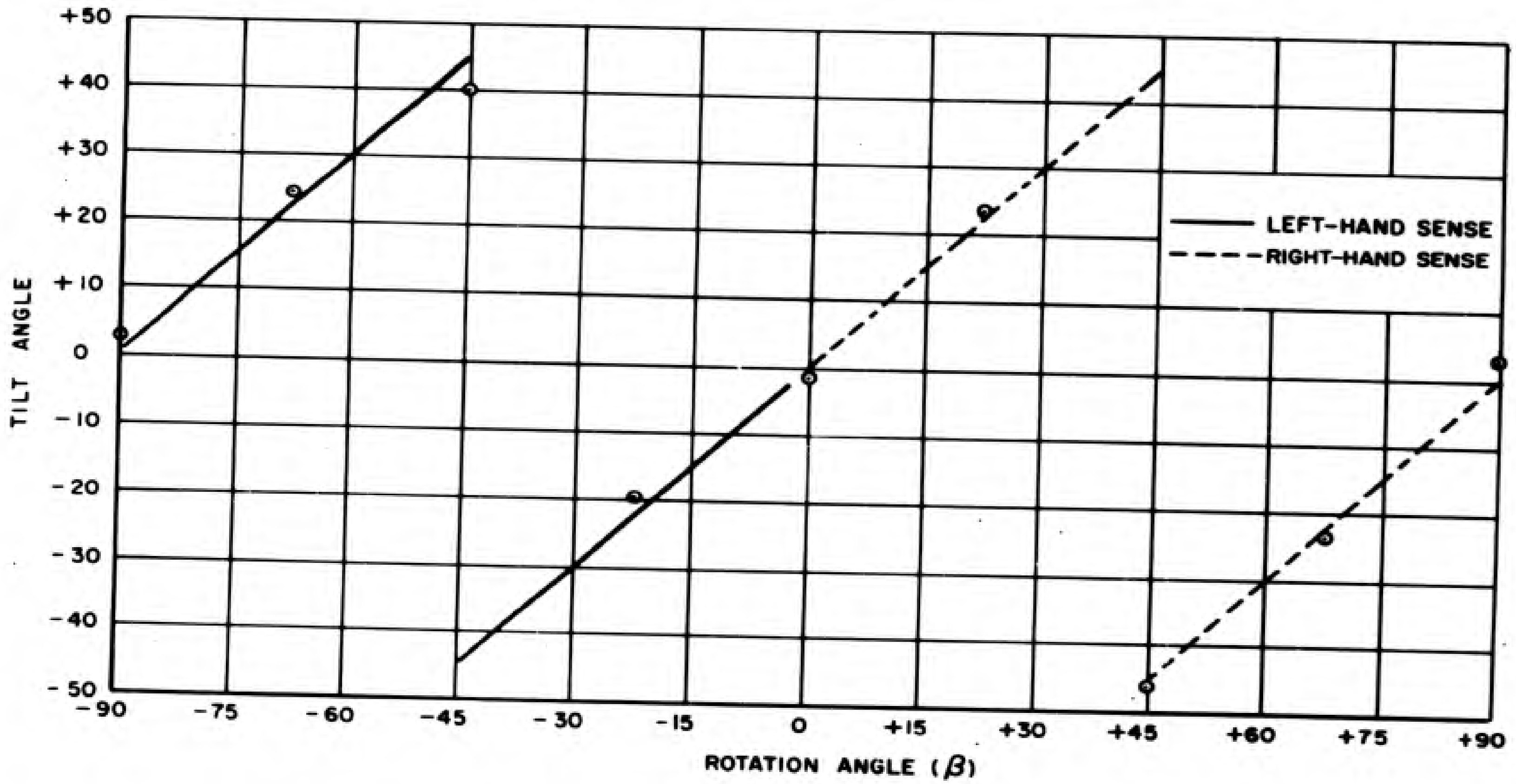

Figure 4. Tilt Angle (ψ) of Polarization Ellipse Generated as Function of $\lambda/4$ Plate Angle (β) with Experimental Data Compared Against Predicted Values (Linear Input)

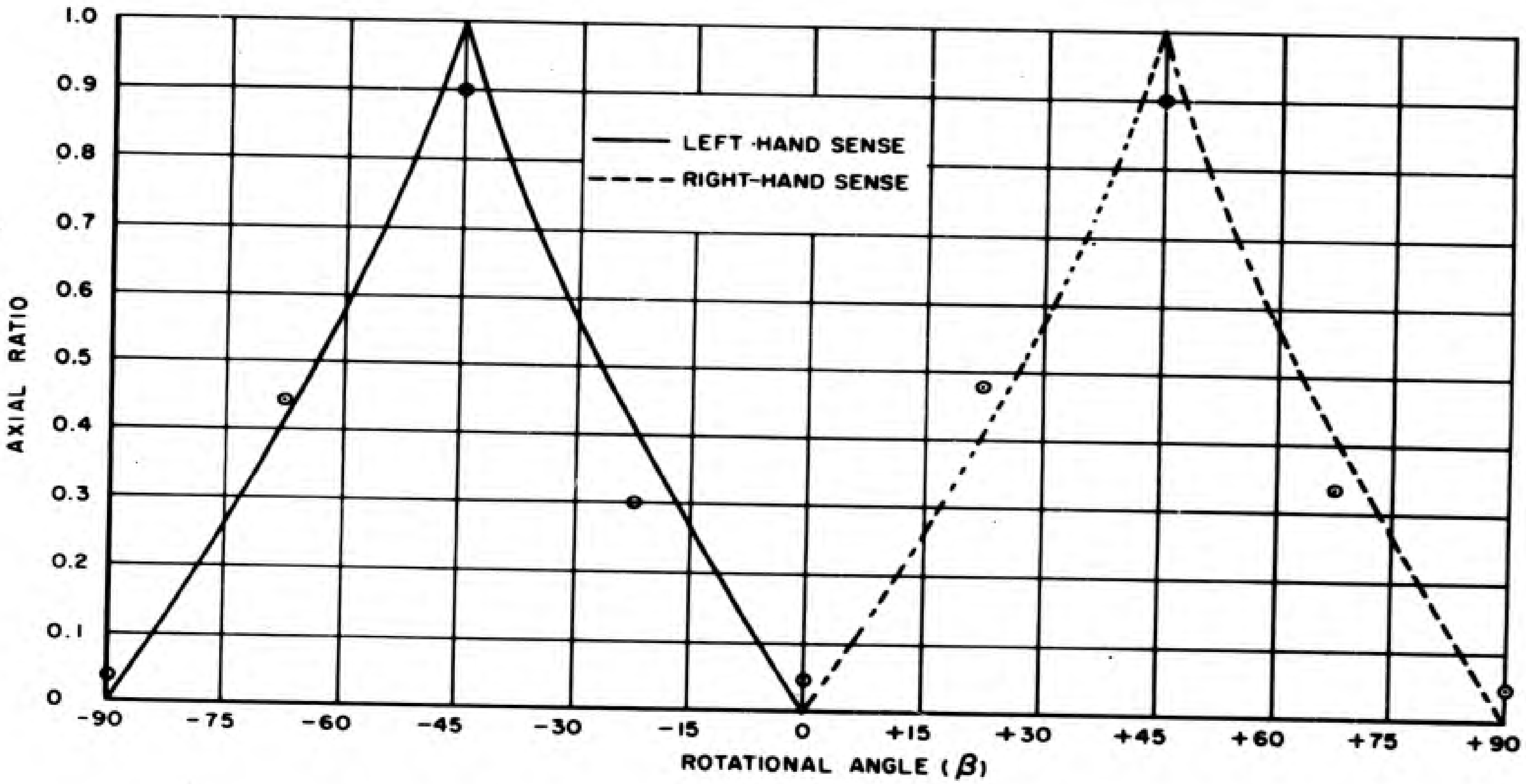

Figure 5. Axial Ratio of Polarization Ellipse Generated as Function of $\lambda/4$ Plate Angle (β) with Experimental Data Compared Against Predicted Values (Linear Input)

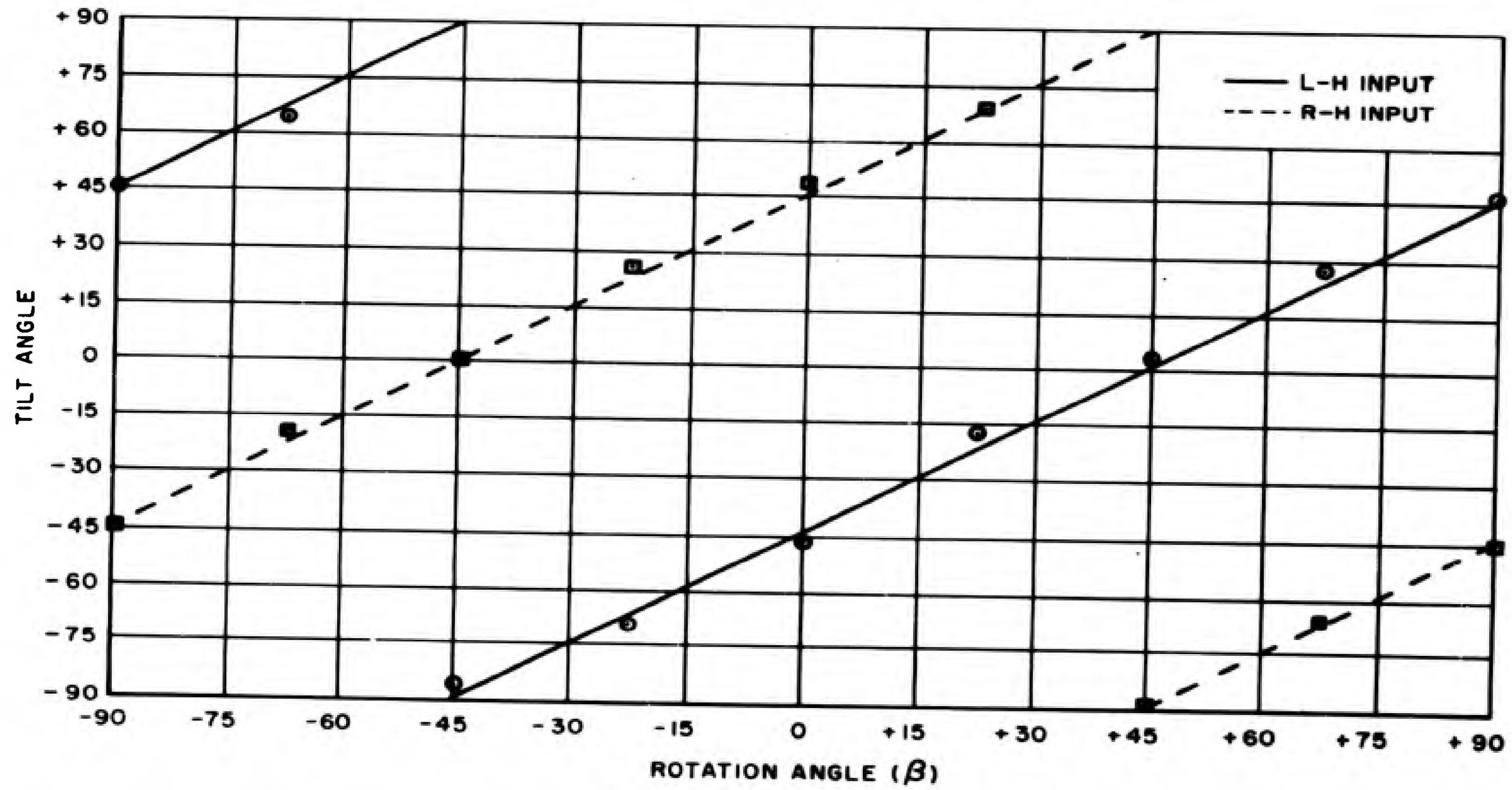

Figure 6. Tilt Angle (ψ) of the Polarization Ellipse Generated as Function of $\lambda/4$ Plate Angle (β) with Experimental Data Compared Against Predicted Values (Circular Input)

However, this requires a non-contacting waveguide rotating joint which has limitations. To eliminate the problem of a non-contacting waveguide rotary joint, an alternate method of rotation was proposed. The dielectric plate was supported inside the guide by small dielectric posts, allowing it to be rotated. The small cross section presented to the wave passing by the posts had a negligible effect on the phase delay. Two techniques of rotating the plate were considered: (a) motor technique and (b) air rotation technique.

3.2.1 Electric Motor Technique

Waveguide, generally made of a non-magnetic metal such as copper, aluminum or magnesium, will allow an externally produced magnetic field to penetrate through the walls. Therefore, if two small magnets are attached to the outside edges of the dielectric plate, it may be rotated by an externally applied magnetic field. A disassembled view of the experimental model is shown in figure 7.

3.2.2 Air Motor Technique

In the air motor technique, an air chamber was built completely around the circular waveguide. Small holes were drilled into the guide wall to introduce jet-like air streams into the guide, thereby rotating the plate. The amount of air pressure required to rotate the plate at a reasonable rate was 3 psi. Figure 8 shows the experimental model with the dielectric supports disassembled.

Both the electric and air motor techniques were investigated fully and found to be mechanically feasible.

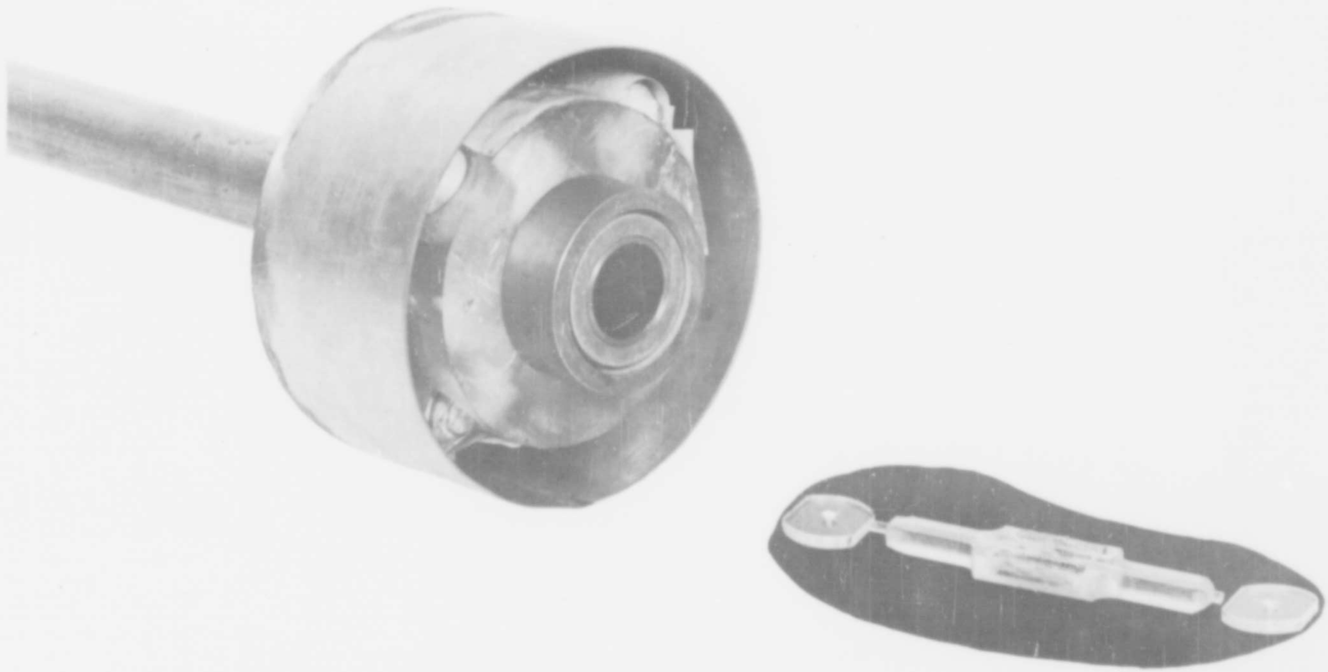

Figure 7. Magnetically Rotatable Quarter-Wave Plate (Disassembled)

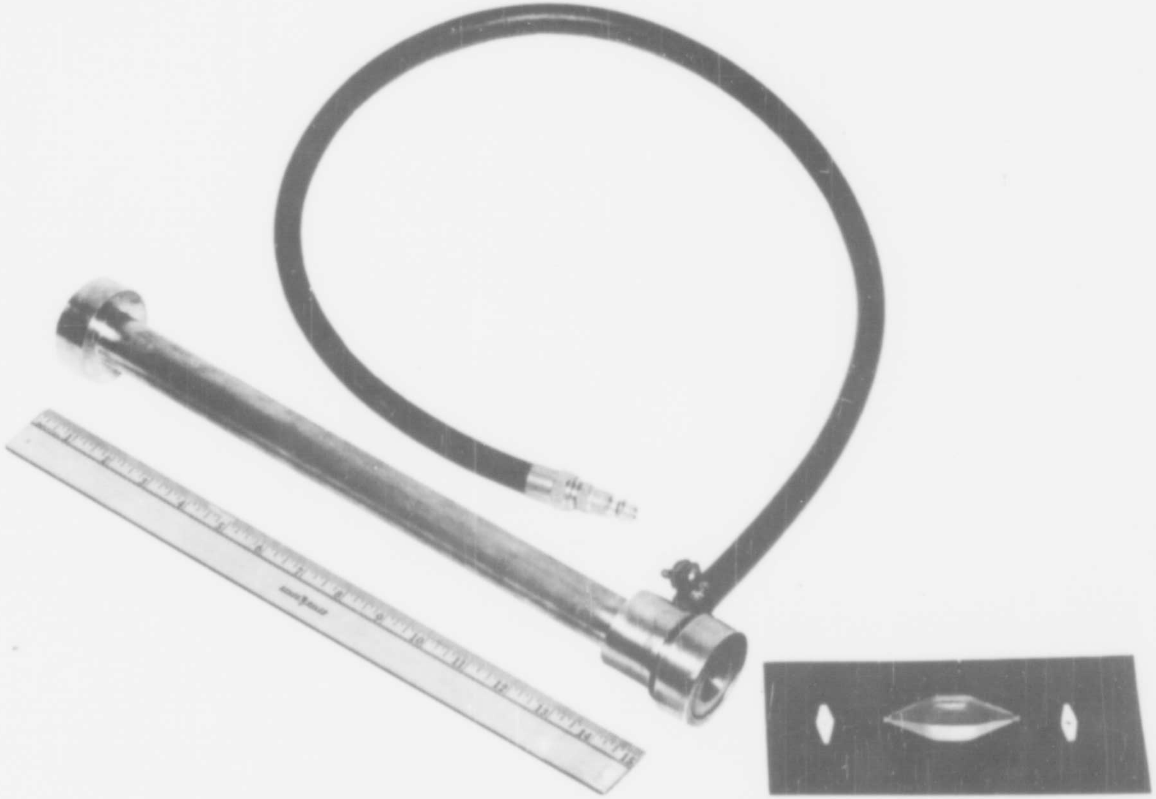

Figure 8. Air Turbine with Quarter-Wave Plate Disassembled

3.3 Electrical Considerations

Since the design of broadband quarter-wave plates is well covered in the literature, the electrical investigations of this study dealt mainly with match and power-handling capabilities. Experimental measurements were conducted over the frequency range 7.6 - 11.0 Kmc.

Attempts to secure a good match in the electric motor technique were unsuccessful because of the magnetic material attached to the dielectric plate. Tests were also conducted to determine the power-handling capabilities of the electric motor technique. The results of these tests indicated there was power breakdown in the region of the magnets when a signal of 15 kw of peak power was applied. As a result of these tests, the motor technique did not receive further consideration. The air-motor technique, however, yielded a VSWR better than 1.89 to 1 using the dielectric supports pictured in figure 9 to hold the quarter-wave plate. The VSWR is plotted as a function of frequency in figure 9. Tests were conducted to determine the power-handling capabilities of the air-rotation technique. These tests indicated no power breakdown when a signal of 21 kw of peak power was applied.

3.4 Quarter-Wave Plate Conclusions

In conclusion, the results of the tests conducted on the electric and air-motor techniques indicated that the air-motor technique was the superior method of achieving a varying polarization using a dielectric quarter-wave plate. Such a method, when used in jamming applications, offers a reasonable latitude for preventing counter-counter-measures by polarization modulation.

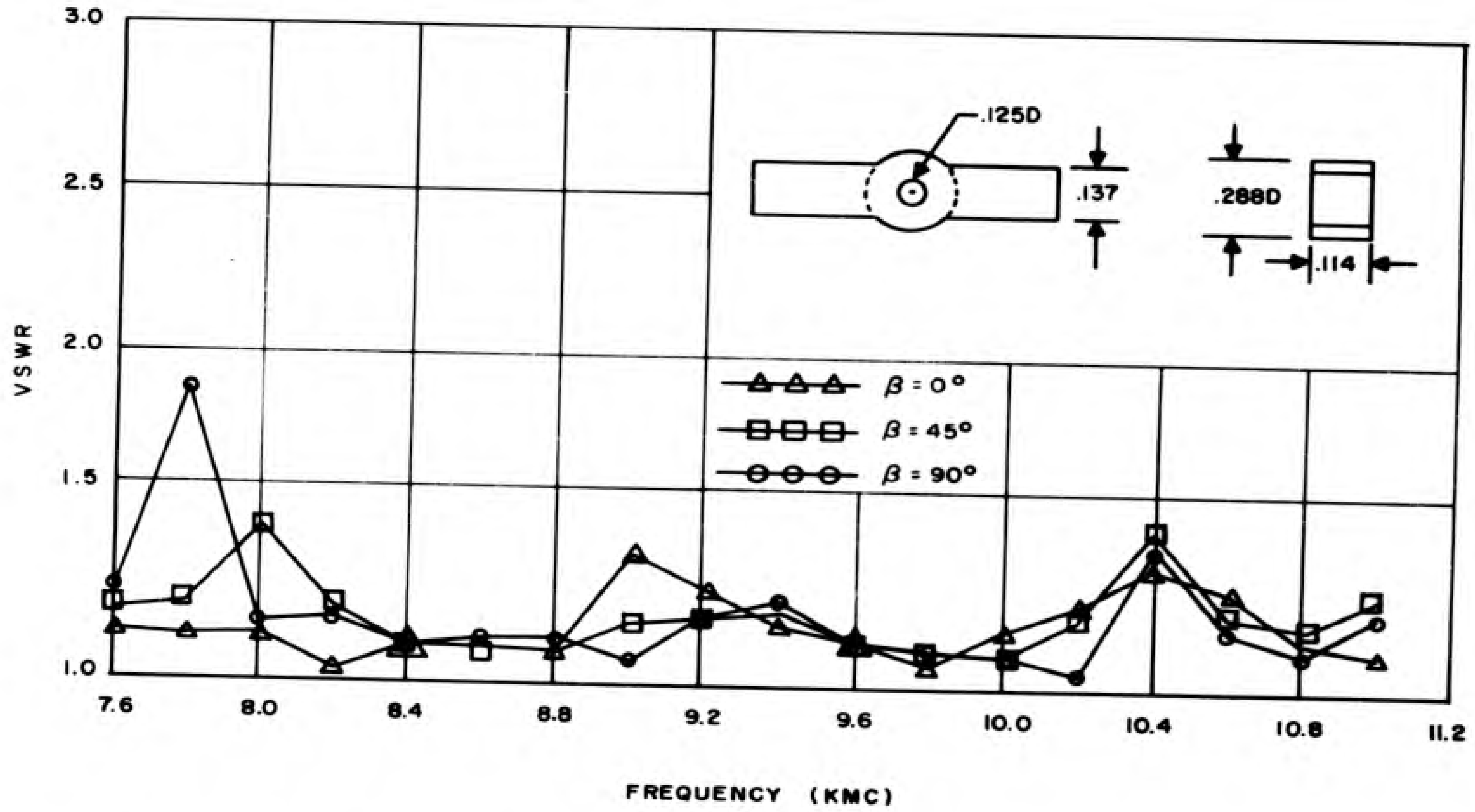

Figure 9. Experimental Dielectric Supports and Resulting Impedance Match

Additional jamming diversity could be obtained by modulating the rotational rate of the quarter-wave plate. Design problems associated with the system are centered around fabrication of suitable non-conducting bearing surfaces. In the experimental model, these surfaces showed signs of deterioration because of friction when air pressure was increased above 3 psi. It is believed that the laboratory tests show the feasibility of the technique, and that the detailed design and analysis required to further develop bearing surfaces would detract from the time to be spent on other specific techniques.

4.0 DUAL-FREQUENCY TECHNIQUE

4.1 Polarization Modulation by the Dual-Frequency Technique

A method of generating a polarization modulated wave which has received considerable attention during this study is the use of two orthogonal linear antennas excited at slightly different frequencies, and, because this difference can be in the r-f range, the instantaneous vector of the resultant wave will rotate at an r-f rate. The general result is an elliptically polarized wave. The axial ratio, tilt angle and sense of the wave are constantly and periodically varying; the period is dependent on the frequency difference.

Applying the proof of reference 1, the expressions for the currents in the two orthogonal antennas are:

$$E_x = A \sin \omega_1 t \tag{4.1}$$

$$E_y = B \sin (\omega_2 t + \phi_o), \tag{4.2}$$

where ω_1 and ω_2 are the angular velocities of the currents in the separate radiators, and ϕ_o is an arbitrary phase angle existing at $t = t_o$. The quantities ω_1 and ω_2 differ by ω_3, or

$$\omega_2 - \omega_1 = \omega_3 \quad \text{and} \quad f_2 - f_1 = f_3 \tag{4.3}$$

Equation (4.2) can be rewritten

$$E_y = B \sin \left[\omega_1 t + (\omega_3 t + \phi_o) \right] . \tag{4.4}$$

Expanding (4.4) in terms of the trigonometric identity for the sine of the sum of two angles, and substituting the value of sin $_1t$ and cos $_1t$ given in (4.1) yields

$$\sqrt{\frac{E_x 2}{1 - A}} \sin (\omega_3 t + \phi_o) = \frac{E_x}{A} \cos (\omega_3 t + \phi_o) - \frac{E_y}{B} .$$

Squaring both sides and rearranging

$$E^2_x \frac{1}{A^2 \sin^2 (\omega_3 t + \phi_o)} + E_x E_y \frac{2 \cos (\omega_3 t + \phi_o)}{A\, B \sin^2 (\omega_3 t + \phi_o)} \tag{4.5}$$

$$+ E_y^{\,2} \frac{1}{B^2 \sin^2 (\omega_2 t + \phi_o)} = 1.$$

If the quantities in brackets are taken to be coefficients, (4.5) is the general form of the equation of an ellipse. However, because the coefficients are not constant, the polarization state will constantly be changing, and the change will be periodic in ω_3. To illustrate this, consider the expression for the tilt angle of the polarization ellipse, which can be found by the substitution of rotational operators in terms of an angle ψ , into (4.5). The angle between the major axis of the ellipse and the X-axis is Ψ (figure 1). The cross-product terms in the resulting equation are made equal to zero, and the solution for them yields

$$\psi = \frac{1}{2} \tan^{-1} \frac{A\,B \quad \cos(\omega_3 t + \phi_0)}{A^2 - B^2} \qquad (4.6)$$

By similar derivations, the axial ratio and rotational sense can also be shown to be periodic in ω_3.

If f_3 is no greater than the bandwidth of the radar being jammed and if proper centering is accomplished, both components (f_1 and f_2) of the signal will be received and the jamming polarization at the radar antenna will be effectively modulated.

Figure 10 is the data measured by rotating (a) a linear antenna, (b) a left-hand circular and (c) a right-hand circular receiving antenna in a polarization modulated field. From 10(a), the unity circularity indicated would lead one to assume that the transmitted signal is circularly polarized; however, the rotational sense is not determinable. If the signal were a conventional circularly polarized one, the response of either a left-hand or a right-hand antenna to the signal would be zero, depending upon its sense. The data of 10(b) and 10(c) show that the polarization is modulated, and hence not a conventional circular wave.

It is clearly indicated that it is impossible to design or orient a conventional linear or circular antenna such that it would be irresponsive to such a polarization modulated wave. The question of degree of response

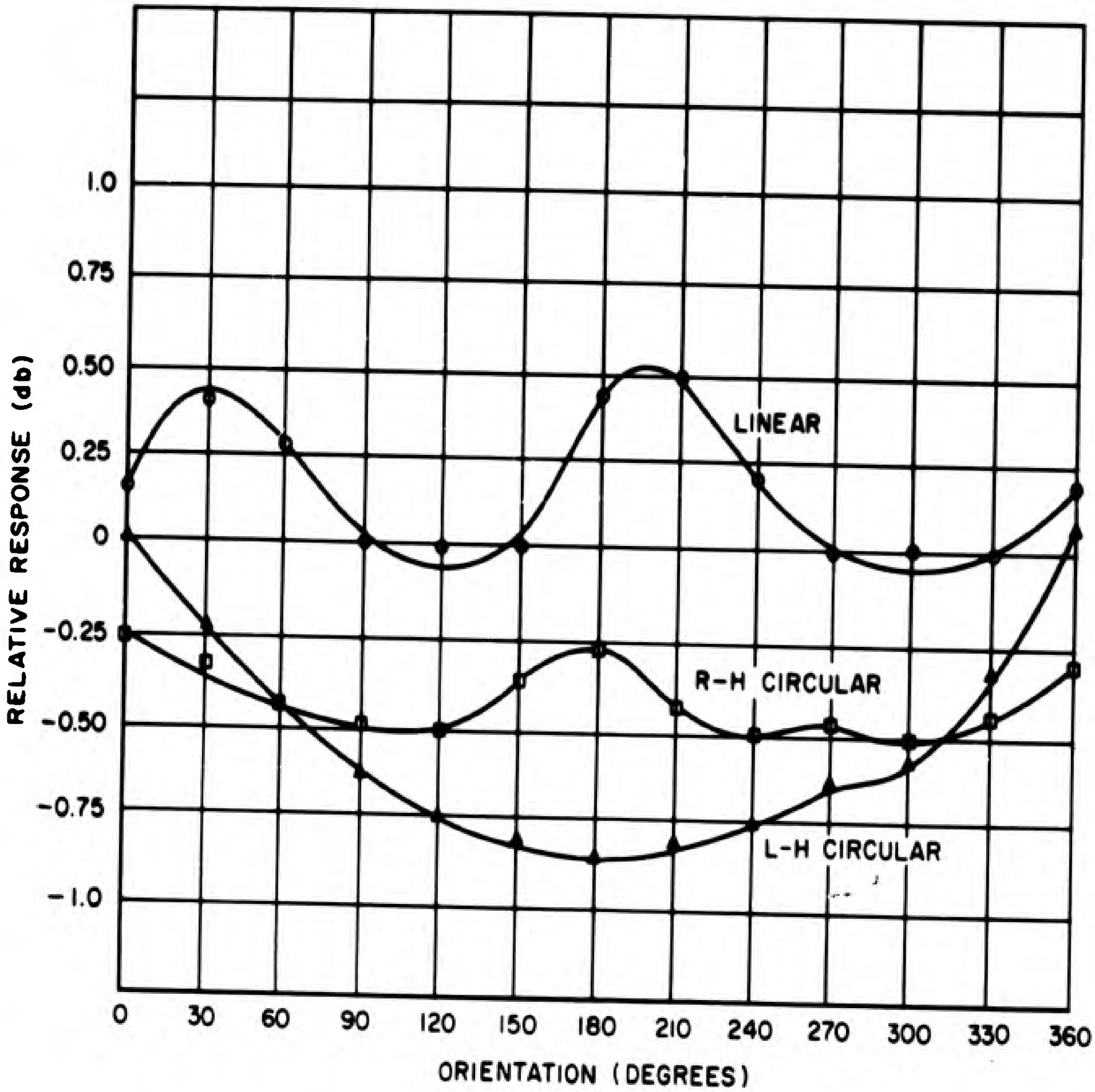

Figure 10. Measured Polarization Response Curves of Differently Polarized Antennas Showing Relative Response to Multipolarized Wave as Function of Orientation

remains, though.

If the jammer were, for example, vertically polarized and were being received by a vertically polarized radar receiving antenna and were receiving a certain amount of energy, it is important to know how much energy, in comparison, that antenna would receive if the same jammer power were transmitted in a polarization modulated field as described above.

In the case of a linearly polarized antenna, if it were aligned with one of the orthogonal components of the field, it is immediately apparent that half of the power would be lost, because none of the energy in the cross-polarized field would be received. If the same antenna were aligned 45 degrees to both orthogonal components, the voltage response to both components would be down 3 db or 0.707 of the maximum value. The effective value of the combination can be found by computing the square root of the sum of the squares, yielding again a 3 db loss.

For a circular receiving antenna, the condition would be the same, the square foot of the sum of the squares of the two 0.707 values. Experimental verification of this is shown in figures 11, 12 and 13.

In figure 11, the response of a linear-receiving antenna to both a linear field and a polarization modulated field, as a function of mechanical rotation of the receiving antenna about its phase center (in the normal fashion of a polarization pattern measurement). The familiar figure-of-eight-dipole pattern is apparent and it can be seen that the expected response to the polarization modulated field produces a constant response without the damaging nulls of the linear case.

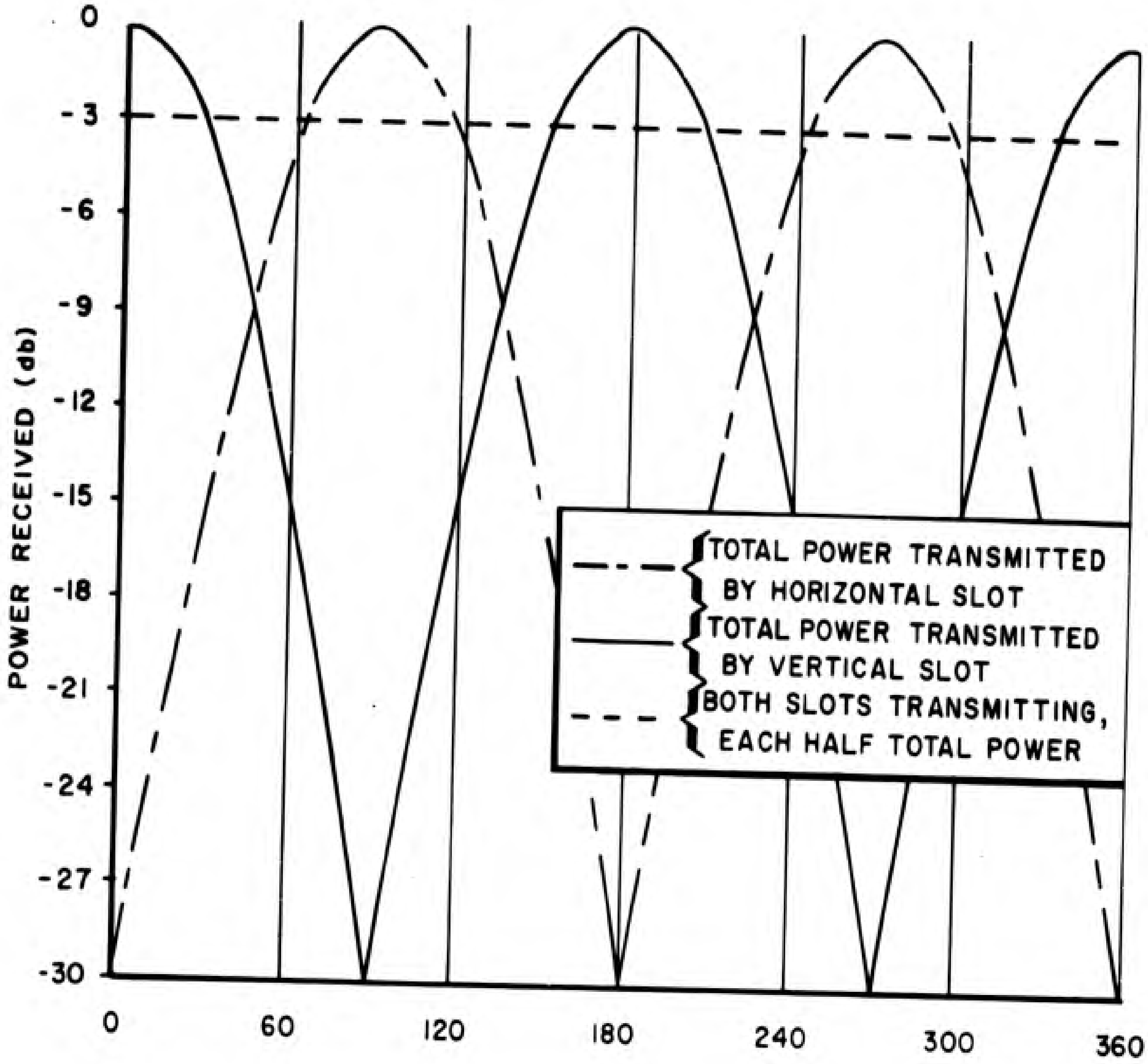

Figure 11. Power Received by Linearly Polarized Antennas as Function of Orientation

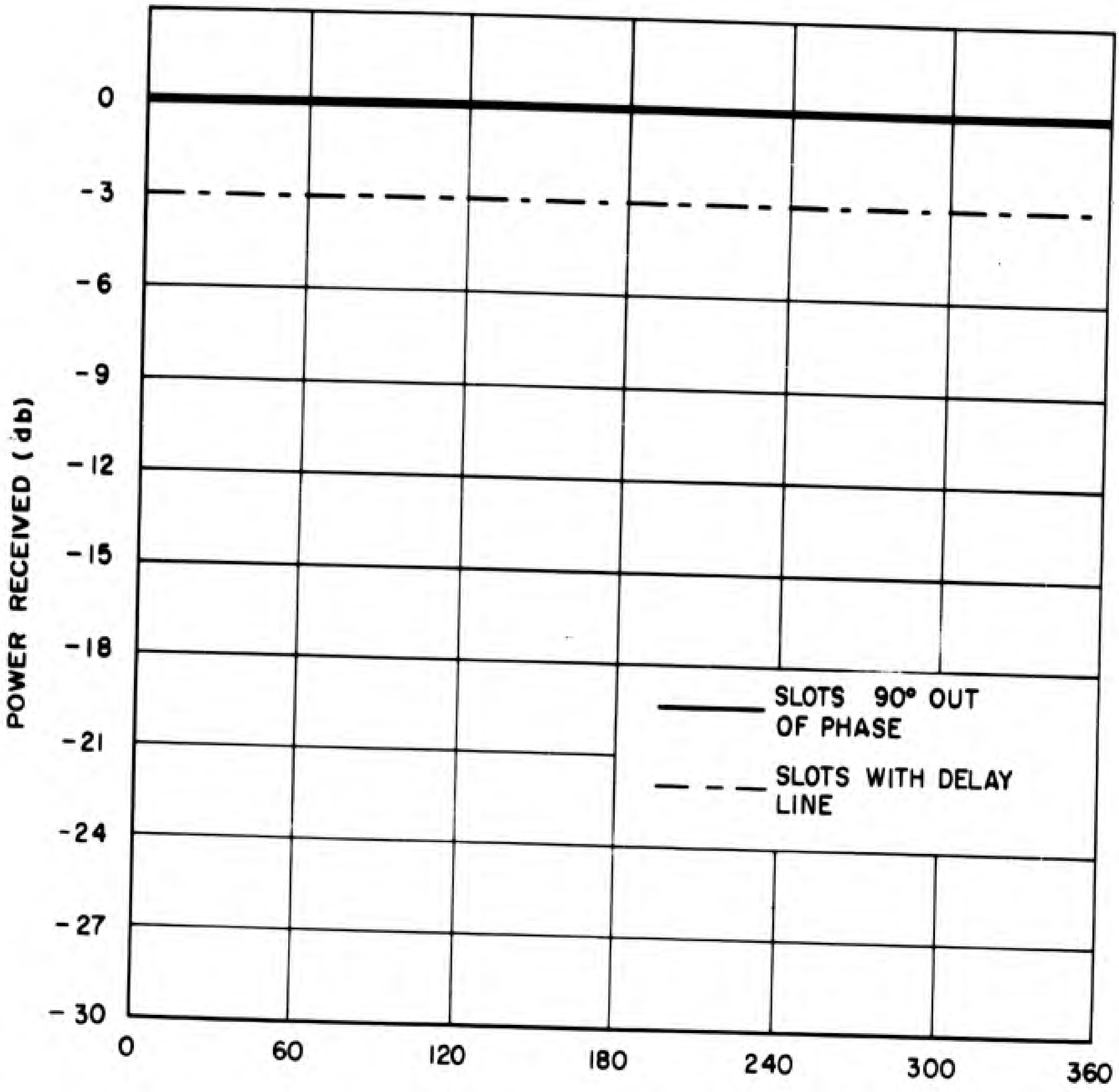

Figure 12. Power Received by Circularly Polarized Antenna as Function of Orientation (Right-Hand Sensed)

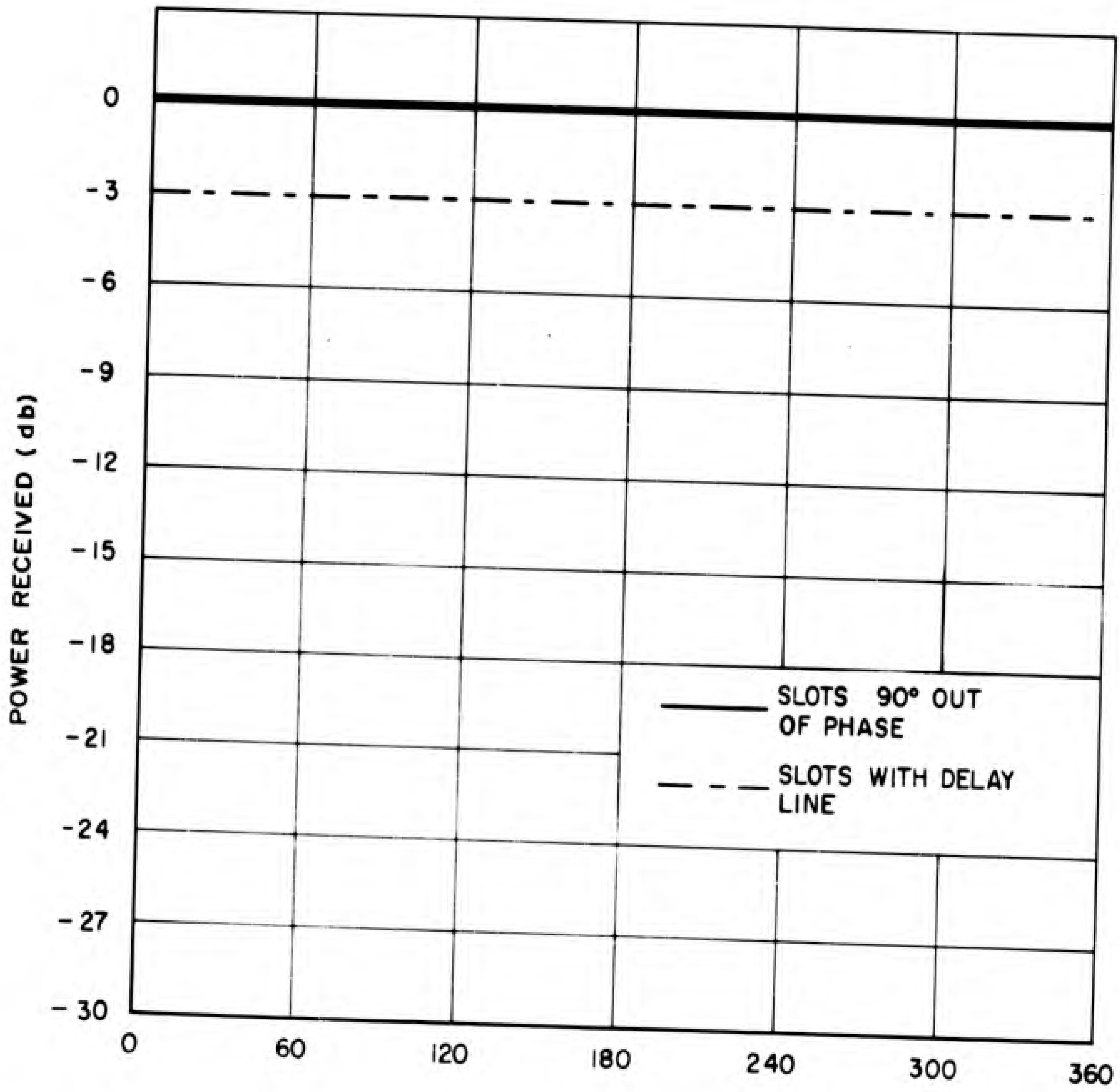

Figure 13. Power Received by Circularly Polarized Antenna as Function of Orientation (Left-Hand Sensed)

Figure 12 is the same measurement made with a circularly polarized antenna to both a right-hand circular field and a polarization modulated field. In figure 13 the same comparison is made for a left-hand circular antenna. In both figures 12 and 13 it is clear that the response of a circular receiving antenna to the polarization modulated field is only 3 db down from that of the proper sensed field, and is equal to the response of a linear receiving antenna.

Conventional jamming situations are vulnerable to exact cross-polarization and likely to produce no effect upon the radar receiver, whereas, a polarization modulated jamming antenna guarantees a response in all situations, regardless of the polarization of the radar antenna.

The essence of this scheme lies in the generation of a pair of r-f signals separated by a frequency f_3. The frequency difference must always be less than the bandpass of the narrowest receiver to be jammed. Because a filter is virtually impossible due to the inherently low values of separation frequency, several methods of generating the two carriers have been investigated in this study.

4.2 Two-Generator Method

One method of implementing the dual-frequency technique to achieve a modulated polarization is illustrated in figure 14. Two high-power oscillators, F and F + Δ F, are each connected to a linear radiator. A portion of the power from each transmitter is coupled off and combined in the mixer. The frequency $F_1 - (F_1 + \Delta F)$ is detected by an FM discriminator and this voltage is used to maintain the frequency difference Δ F between the transmitters.

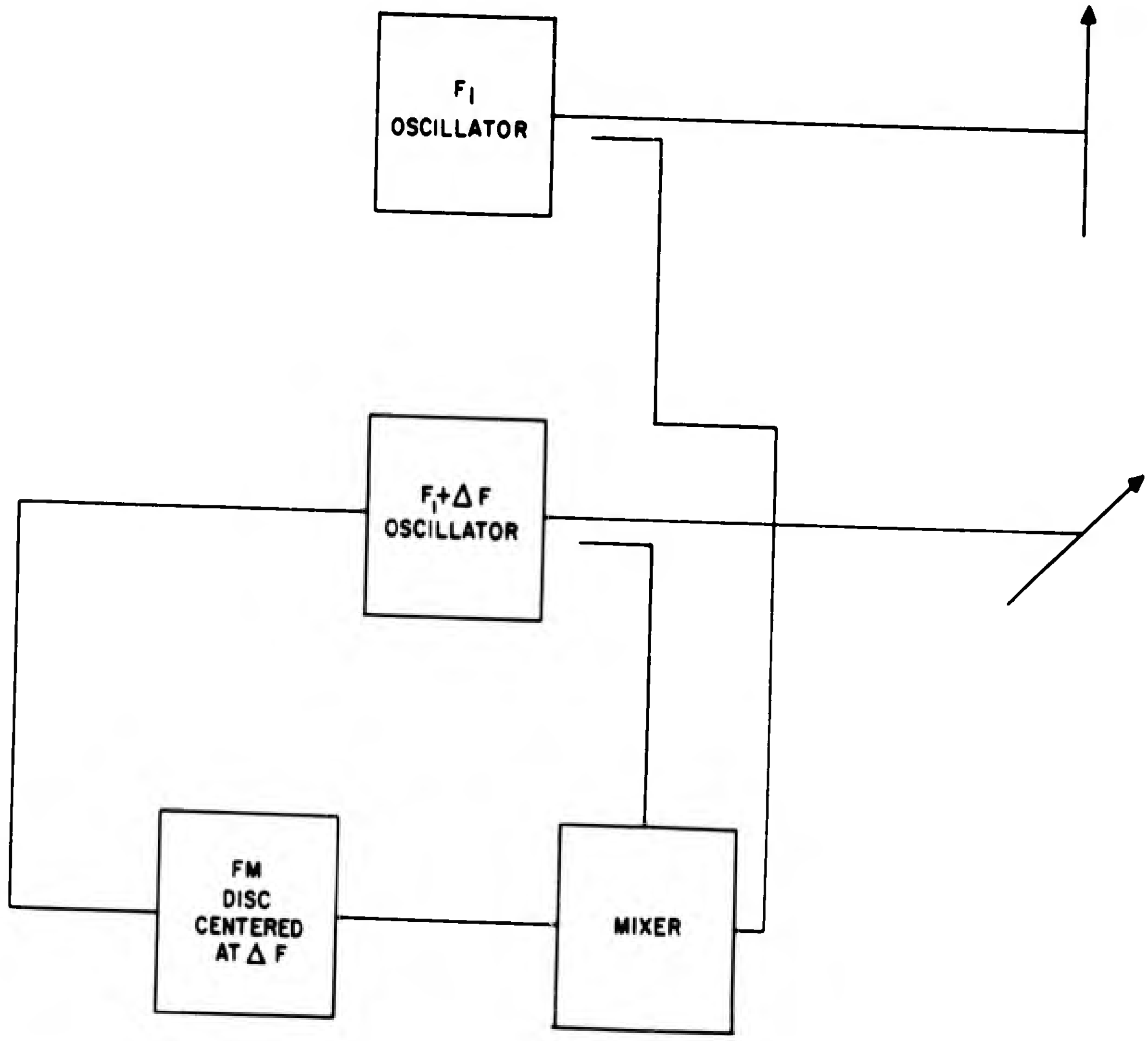

Figure 14. Frequency Control System for Two-Generator Method

Advantages of this system are its ability to generate a polarization that changes at an RF rate while remaining in the frequency range of a victim receiver with a very narrow bandpass. The primary disadvantage is the necessity of using dual transmitters.

4.3 Delay-Line Method

Another method of producing the two frequencies was devised under the program. It is shown in figure 15, where the output of a frequency modulated generator is divided and fed to two orthogonal antennas. A long delay line is placed in the line leading to one antenna, thereby causing the signal reaching that antenna to be slightly different in frequency from the signal reaching the other antenna. Figure 16 shows how the difference is derived. It shows that at the instant chosen, the frequency of the signal at "a" lags that at "b" by f'. The frequency difference can be found by multiplying the delay of the delay line (in time) by the derivative of the frequency:

$$f' = \frac{Ld}{Kdt} \quad (f_i) \qquad (4.7)$$

where K = velocity of propagation in delay line (feet/sec)

f_i = instantaneous frequency of the modulated signal at antenna "b"

L = delay line length (feet)

f_m = frequency of the modulating signal

$$f' = d/dt \, \frac{1}{2} \quad (f_o + \Delta f \sin \omega_m t) \quad L/K \qquad (4.8)$$

$$= \frac{\Delta f \quad L}{K} \quad f_m \cos \quad \omega_m t \qquad (4.9)$$

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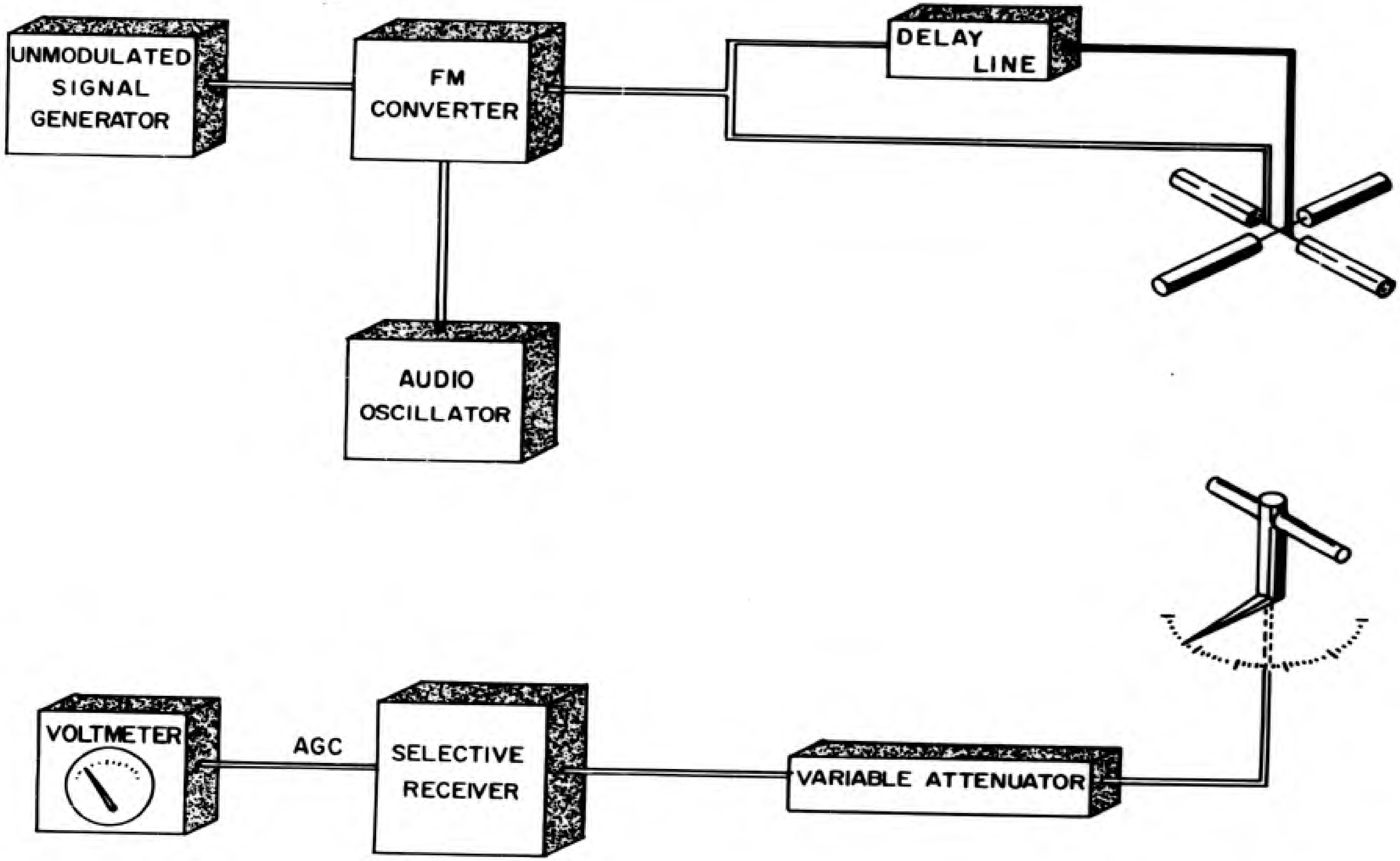

Figure 15. Block Diagram of Equipment Arrangement Using Delay Line

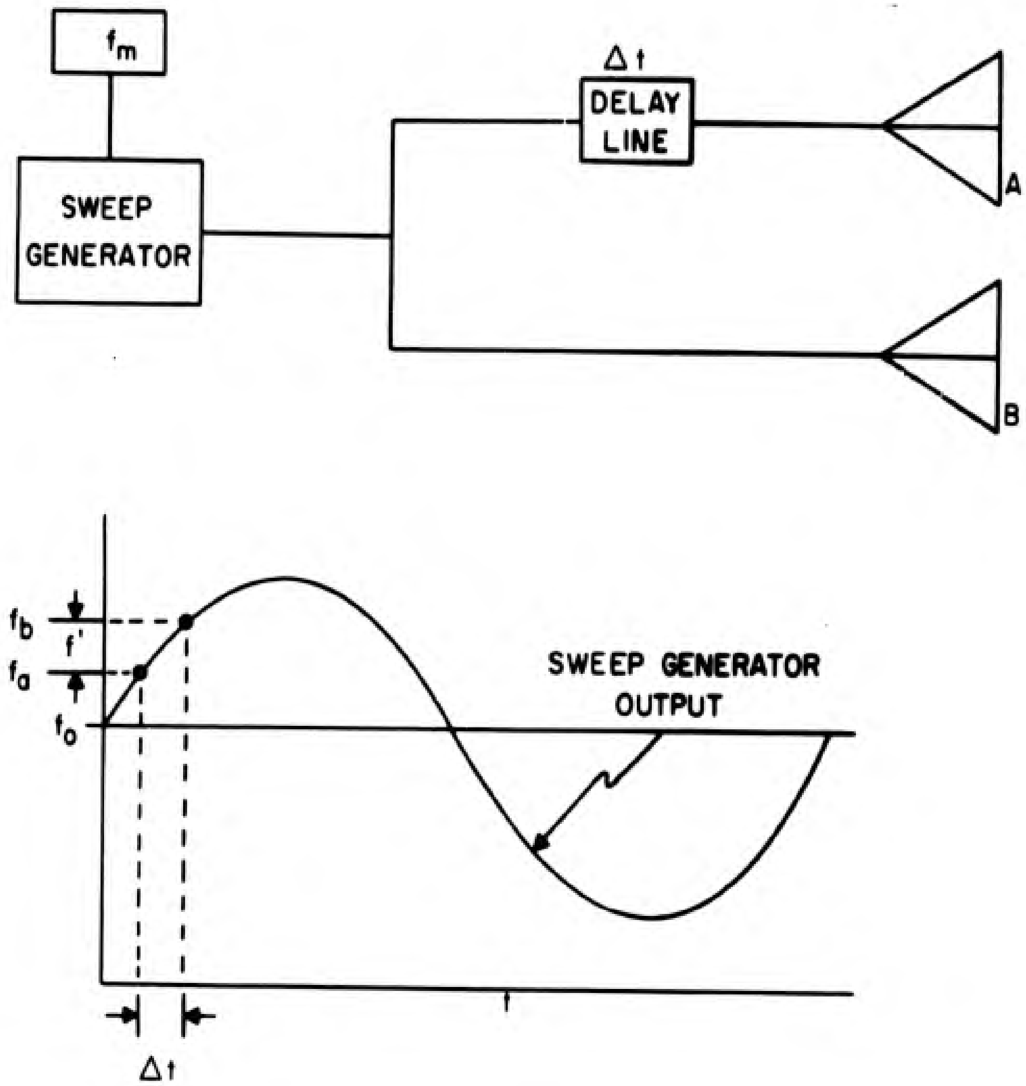

Figure 16. Equipment Arrangement and Graphical Description of Operation of Sweep Generator and Delay Line Method of Obtaining Dual-Frequency Output R0401

This is an approximation of the frequency difference because the slope will be different at the two points, but the difference is infinitesimal for practical values.

If the expression for the frequency difference given in (4.8) is substituted into equation (4.4), it appears that an expansion in a Bessel function would be possible, indicating the presence of a number of discrete polarization ellipses; however, a more direct approach is chosen, that of performing the Bessel expansion directly upon the expression of the generator output. In the usual textbook manner,[6,7] without including the details of the derivation, the generator output

$$E_b = B \sin\left(\omega_o t + \frac{\Delta f}{f_m} \sin \omega_m t\right) \tag{4.10}$$

is rewritten in its expanded form

$$\begin{aligned} Eb = B \Big\{ & J_0(m_f)\sin(\omega_o t) + J_1(m_f)\left[\sin(\omega_o + \omega_m)t - \sin(\omega_o - \omega_m)t\right] \\ & + J_2(m_f)\left[\sin(\omega_o + 2\omega_m)t + \sin(\omega_o - 2\omega_m)t\right] \\ & + J_3(m_f)\left[\sin(\omega_o + 3\omega_m)t - \sin(\omega_o - 3\omega_m)t\right] + \cdots\cdots \Big\} \end{aligned} \tag{4.11}$$

where $\Delta f/f_m = m_f$, called "modulation index."

It is seen, as expected for sinusoidal frequency modulation, that in addition to the carrier of frequency f_o, there is an infinite number of paired sidebands, symmetrical in position above and below the carrier, and separated by the modulation frequency, f_m. Their relative amplitudes can be found by referring to any table of Bessel functions.

Looking again at figure 15, and remembering that the signal going to antenna "a" experiences a delay, it is seen that the signals appearing at "a" and "b" will be identical except that all of the components of the FM spectrum at "a" will be delayed in time from those at "b". If the amount of delay for all of the sidebands were equal, then the phase angle between a given sideband at "a" and its counterpart at "b" would be the same for all such sidebands, or for all of the spectral components. However, because the delay line length in electrical degrees is not the same for all frequencies, there will be different phase angles for each frequency. Normally, the width of a jamming signal is very small in comparison to the center frequency and the n^{th} sideband is so slight as to be negligible. In the present situation, though, the delay line length is great enough to create a large phase difference change for a correspondingly small frequency change.

The difference in phase that will exist between a signal at "a" and the same frequency signal at "b" will be δ, such that

$$\delta_n = \frac{2\pi L}{\lambda_n} \tag{4.12}$$

where λ_n is the wavelength of the two signals being compared. To write an expression for the phase difference between any signal at antenna "a" and its counterpart at antenna "b", it is first necessary to define the quantity m_d, to be called "degree of modulation", where

$$m_d = \frac{f_m}{f_o} \tag{4.13}$$

From the relationship between frequency and wavelength and the separation of the sidebands by the modulation frequency, we can derive the following expressions for the phase angle of a given frequency signal.

$$\delta_{nu} = \delta_o \ (1 + nm_d) \tag{4.14}$$

$$\delta_{nl} = \delta_o \ (1 - nm_d) \tag{4.15}$$

In the expressions above, the subscripts nu and nl are meant to denote the n^{th} upper and n^{th} lower sidebands.

Choosing antenna "b" as a phase reference, we can now write an expression for the signal at antenna "a".

$$\begin{aligned} E_a = A \Big\{ & J_o(m_f) \ \sin(\omega_o t - \delta_o) \\ & + J_1(m_f) \ \left[\sin(\omega_o t + \omega_m t - \delta_{1u}) - \sin(\omega_o t - \omega_m t - \delta_1)\right] \\ & + J_2(m_f) \ \left[\sin(\omega_o t + 2\omega_m t - \delta_{2}u) + \sin(\omega_o t - 2\omega_m t - \delta_2)\right] \\ & + J_2(m_f) \ \left[\sin(\omega_o t + 3\omega_m t - \delta_{3}u) - \sin(\omega_o t - 3\omega_m t - \delta_2)\right] \\ & + \ldots\ldots \Big\} \end{aligned} \tag{4.16}$$

A comparison of equations (3.11) and (3.16) shows that there exists at either antenna a number of separate signals, each different in amplitude and frequency, and each separated from its neighbor by the modulating frequency, f_m. At the other antenna, the condition will be almost identical, that is, the same number of signals of the same frequencies, all having a common, proportional amplitude relationship to the signals at the first antenna, and each signal differing in phase with respect to its counterpart at the first antenna by an amount which is different for each frequency. Remembering that the two antennas are in space quadrature, it is apparent that the field produced by a sideband will be some form of an elliptically polarized wave. Because the phase angle for the two signals at some other sideband frequency will be different, the polarization of the field produced at each frequency will also be different.

The amount of phase change as a function of frequency is an instrumental quantity in the exploration of this approach and its relationship to the delay line length should be noted. It is given in the expression below, where f_1 and f_2 are any two frequencies and the change in δ, which is called $\Delta\delta$, is proportional to the frequency difference, thus:

$$\Delta\delta = \frac{2\pi L (f_2 - f_1)}{K} \tag{4.17}$$

K is the proportionality constant between λ and f in the delay line.

To demonstrate the importance of the quantity $\Delta\delta$, and to verify the calculations presented, an experimental program was conducted. The equipment arrangement described in figure 15 was used, where the output of a sinusoidally frequency modulated generator was divided and fed to two orthogonal dipole radiators. A delay line was inserted in one line, which was long enough to provide a $\Delta\delta_{max}$ of 280 degrees. The modulation frequency was sufficiently great so that the Bessel components were separated to the extent that they could be individually tuned on a selective receiver. The receiver was connected to a rotatable dipole and both axial ratio and tilt angle were measured at each frequency for all the detectable sidebands. This data is shown in figures 17 and 18, along with the calculated performance of the system.

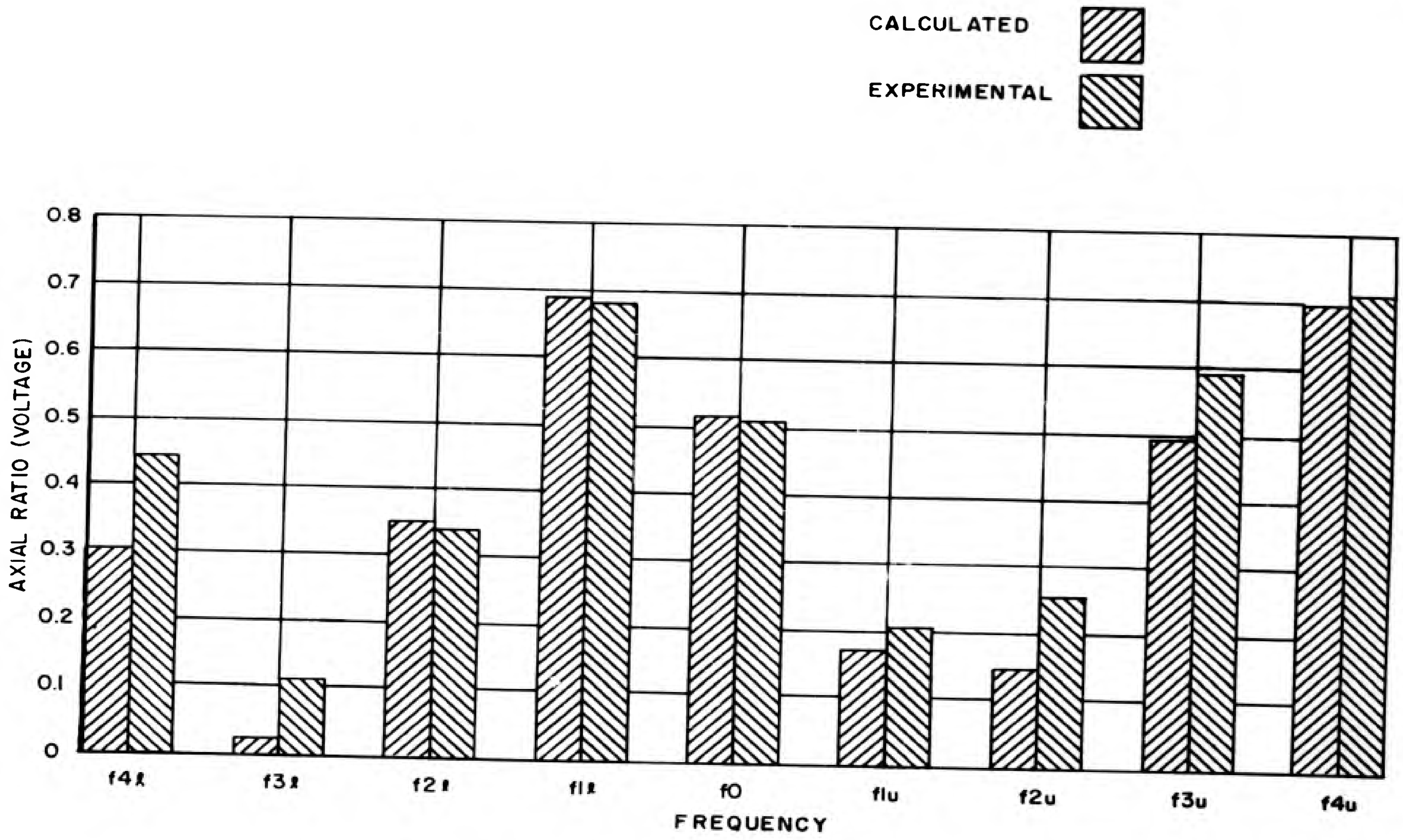

Figure 17. Axial Ratio vs. Frequency of Side Bands

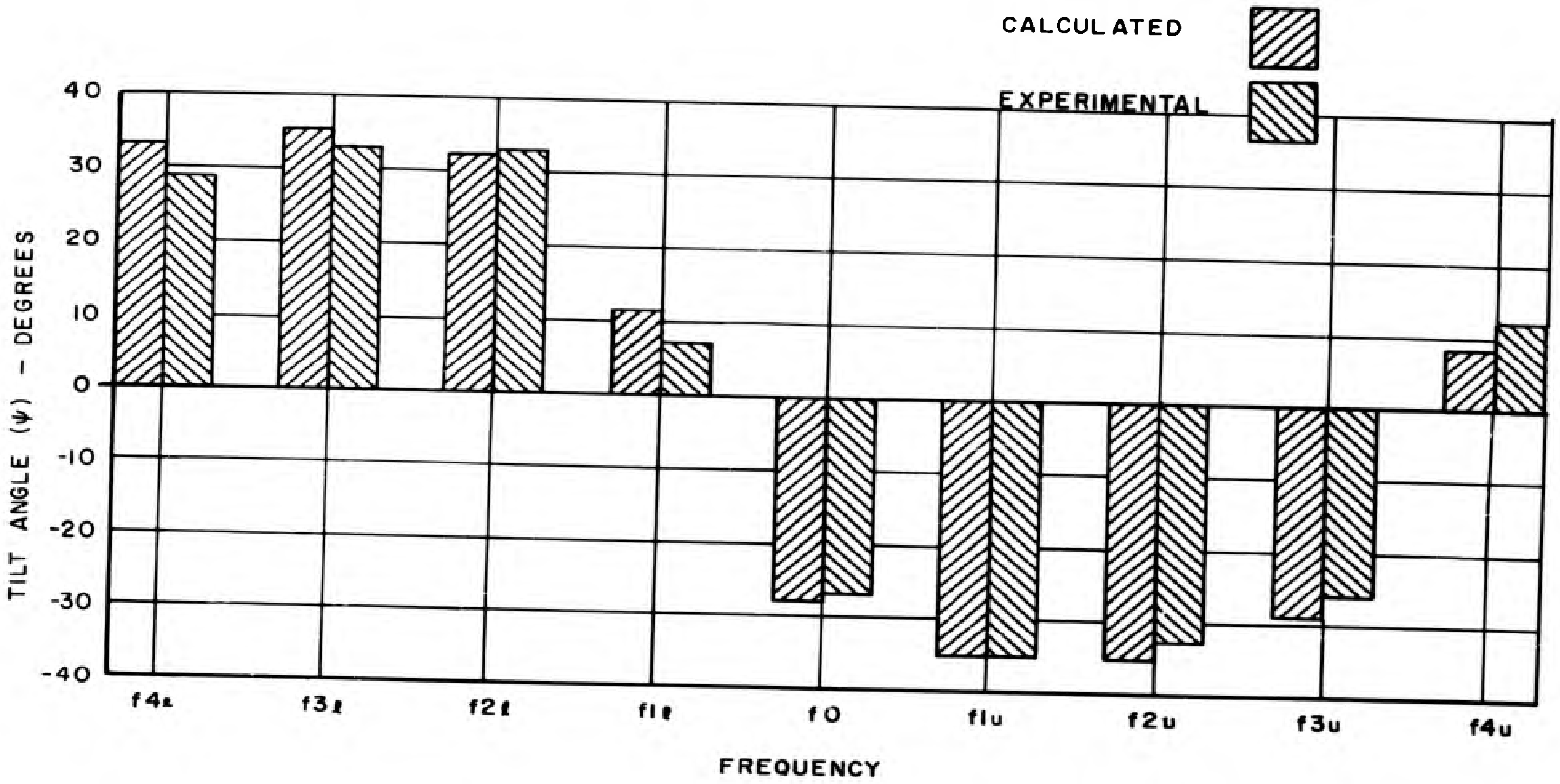

Figure 18. Tilt Angle vs. Frequency of Side Bands

Sinusoidal modulation is only of interest for research purposes, however, and the projection of this work into a practical situation would immediately involve other types of modulation. The most prominent of these would be broadband noise, where a continuous spectrum would effectively exist between the limits of the signal width, as opposed to the sinusoidal modulation case where only a finite number of discrete polarization ellipses exist. The condition of a continuous spectrum offers some design convenience. It is possible to reduce the design process to a simple relationship between L and signal width, such that a complete variation from one polarization state, through all other possible polarizations, back to the original polarization - or a $\Delta\delta$ of 2π radians.

then, by equation (4.17)

$$2\pi = \Delta\delta = \frac{2\pi L\,(f_u - f_L)}{K}$$

and $$L = \frac{K}{(f_u - f_L)} \tag{4.18}$$

where f_u and f_L are respectively the upper and lower frequency extremes.

Using the equipment of figure 19, data was measured on a continuous-spectrum signal. The signal generator of the figure is a BWO which provides a nominally constant output with varying frequency, and is capable of simultaneous amplitude and frequency modulation. Sawtooth frequency modulation and 1000-cycle square-wave amplitude modulation were used and the amplitude component was detected by a crystal and measured on a bolometer amplifier. Additionally, an oscilloscope whose beam was horizontally deflected by the

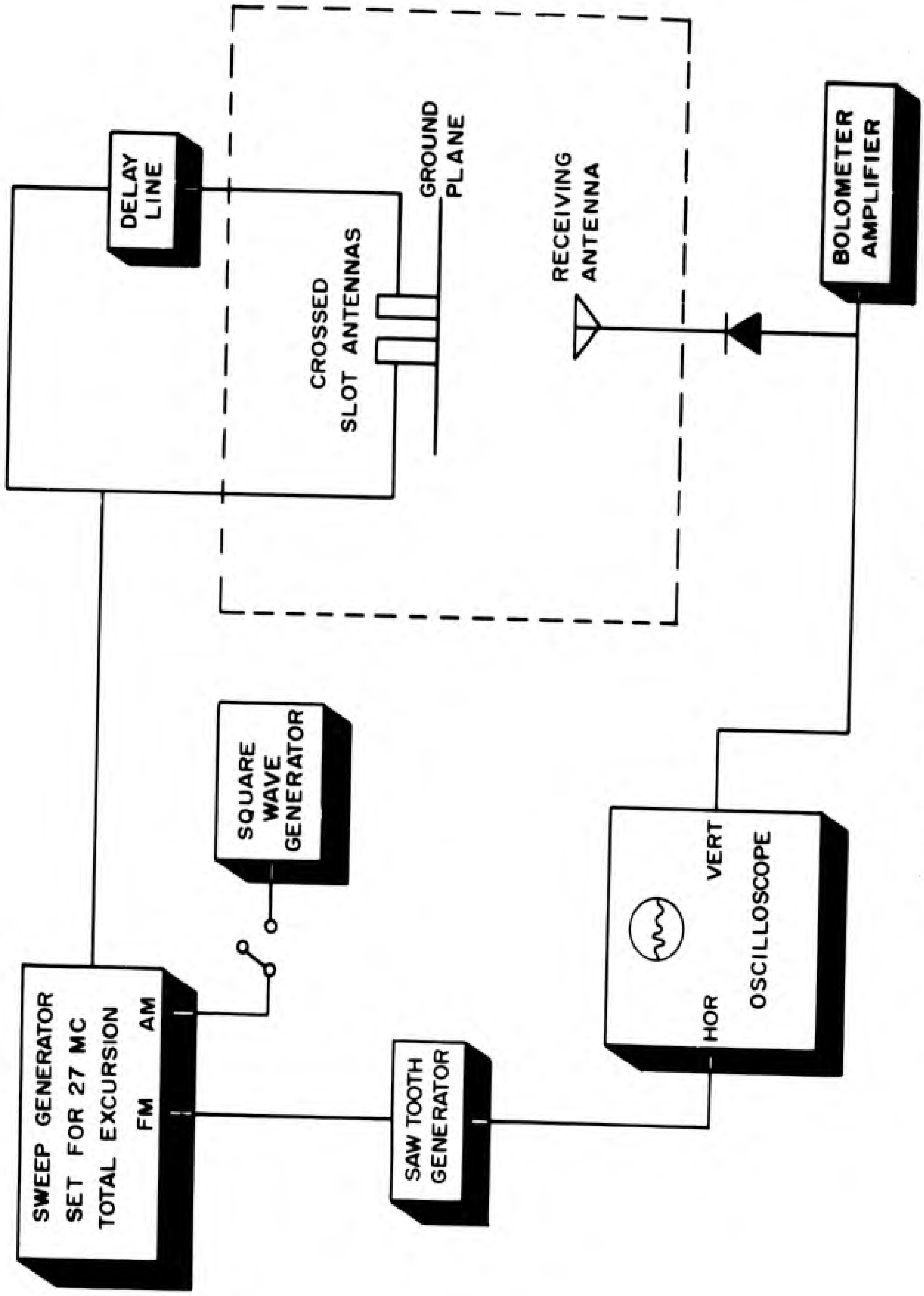

Figure 19. X-Band Delay Line Test Setup

sawtooth voltage used for FM, and vertically deflected by the crystal output of the receiving antenna; produced an amplitude vs frequency curve for four differently polarized antennas. Figure 20 contains oscillograms of these measurements and the expected results are seen to be verified. In 20(a) and 20(b) are shown frequency vs response curves for a linear receiving antenna which was aligned separately with the two orthogonal components of the transmitted field. These indicate that the component signals are of equal amplitude. In 20(c) the linear receiving antenna was rotated to an orientation of 45 degrees to each of the orthogonal components, and shows that the wave was linearly polarized and aligned with the receiving antenna at a point half way between the lower end of the signal and the center frequency. It also shows that the wave was linear and cross-polarized to the receiving antenna at a point half way between the center and the high end of the signal. The oscillogram of 20(d) corroborates this, because for (d) the linear receiving antenna was rotated 90 degrees from its position in (c).

The oscillograms of 20(d) and 20(e) were measured using identical left-hand and right-hand circularly polarized antennas, respectively. It is indicated that the wave is left-hand circular at the center frequency and right-hand circular at the extremes.

4.3.1 Implementation of Delay-Line Method

The most important factor in utilizing the delay line method is the delay line length. Theoretically, this parameter has much design flexibility; however, in a real situation the losses incurred through the delay

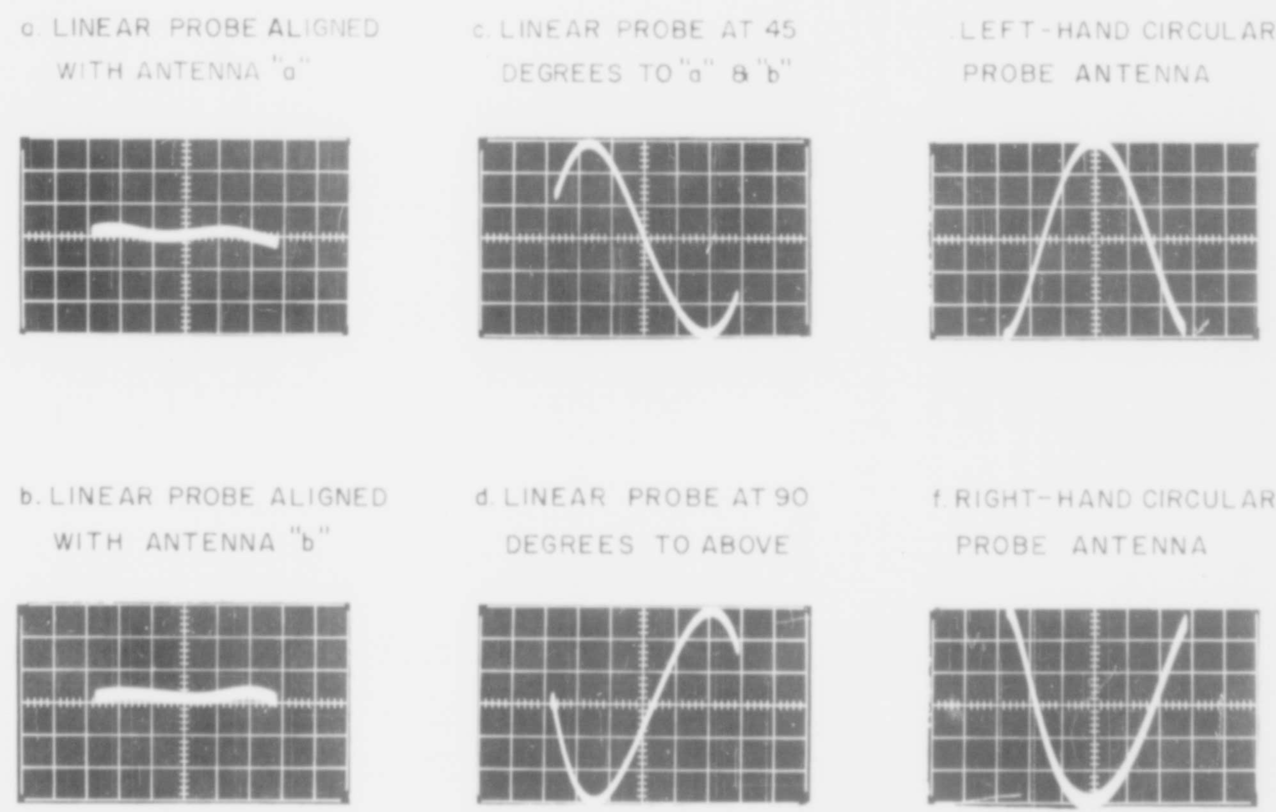

Figure 20. Oscillograms of X-Band Non-Sensed Antenna

line become very important.

A choice must be made between the degree of polarization diversity, the transmission losses in the delay line, and the bandpass of the victim receiver.

By equation (3.18) the length of delay line necessary to generate 360° phase change is exactly equal to $\frac{K}{f_u - f_L}$. Since $(f_u - f_L)$ must be less than the bandpass of the intended receiver if all polarizations are to be generated at the receiver, the shortest usable delay line would be one wavelength long at the band-width frequency of the victim receiver. For instance, a delay line (K/4 mc) long would be 100 percent effective against a receiver with a 4 mc bandpass.

The bandpass of the intended receiver since not known, must be considered in the worst case. Probably a communications receiver would have the narrowest bandpass to be jammed. The type of delay line would, in general, be determined by the frequency of the jamming equipment. At frequencies up through the UHF band coaxial cable or even artificial delay lines are practical and provide a satisfactory solution to the very long lengths of delay line necessary. However, at the higher frequencies the losses incurred in coaxial cable become prohibitive and waveguide components must be used. One compensating factor is the normally wider bandpass at the higher frequencies.

Several ideas have been investigated in the pursuit of either an extremely now loss delay line or one that provides a large amount of delay in a reasonably short length of guide. Three types of waveguide delay line have received attention in this study. Conventional rectangular waveguide operating in a TE_{10} mode has the advantage of being easy to construct; however, the extremely long lengths necessarily create losses which make this type of delay line impractical. Circular waveguide operating in a TE_{11} mode has the desirable feature that its length can be reduced by a factor of 2 if it is short circuited and a three part transition is used as shown in figure 21. This only reduces the length of line necessary and does not reduce the loss. The third type of waveguide delay line that received attention is a circular type of guide operating in a TE_{01} mode. This type of waveguide has the advantage of very low loss; however, the excitation of modes other than TE_{01} makes it very critical to irregularities in the guide. In applications such as this where very long lengths are required, circular guide in the TE_{01} mode become cumbersome and impractical.

To summarize, it would seem the delay line method offers a satisfactory solution to the equipment problem, but is limited by the types of delay lines available. The technique is practical at reasonably low frequencies or in applications where the victim receiver has a wide bandpass.

4.4 Heterodyne Phase Shifter

In addition to the methods discussed above, a variation of the delay line technique has been investigated. The main advantage of this method is that the delay line which provides the relative frequency difference can be used at an intermediate frequency, perhaps 30 mc, where delay line techniques are more realizable. This method is shown in figure 22. The output frequencies

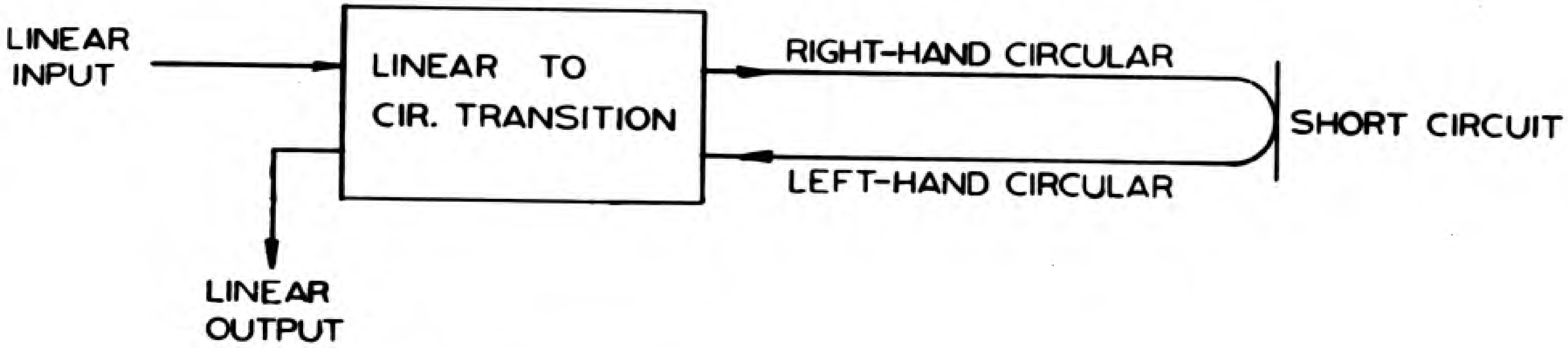

Figure 21. Delay Line Using Shorted Waveguide

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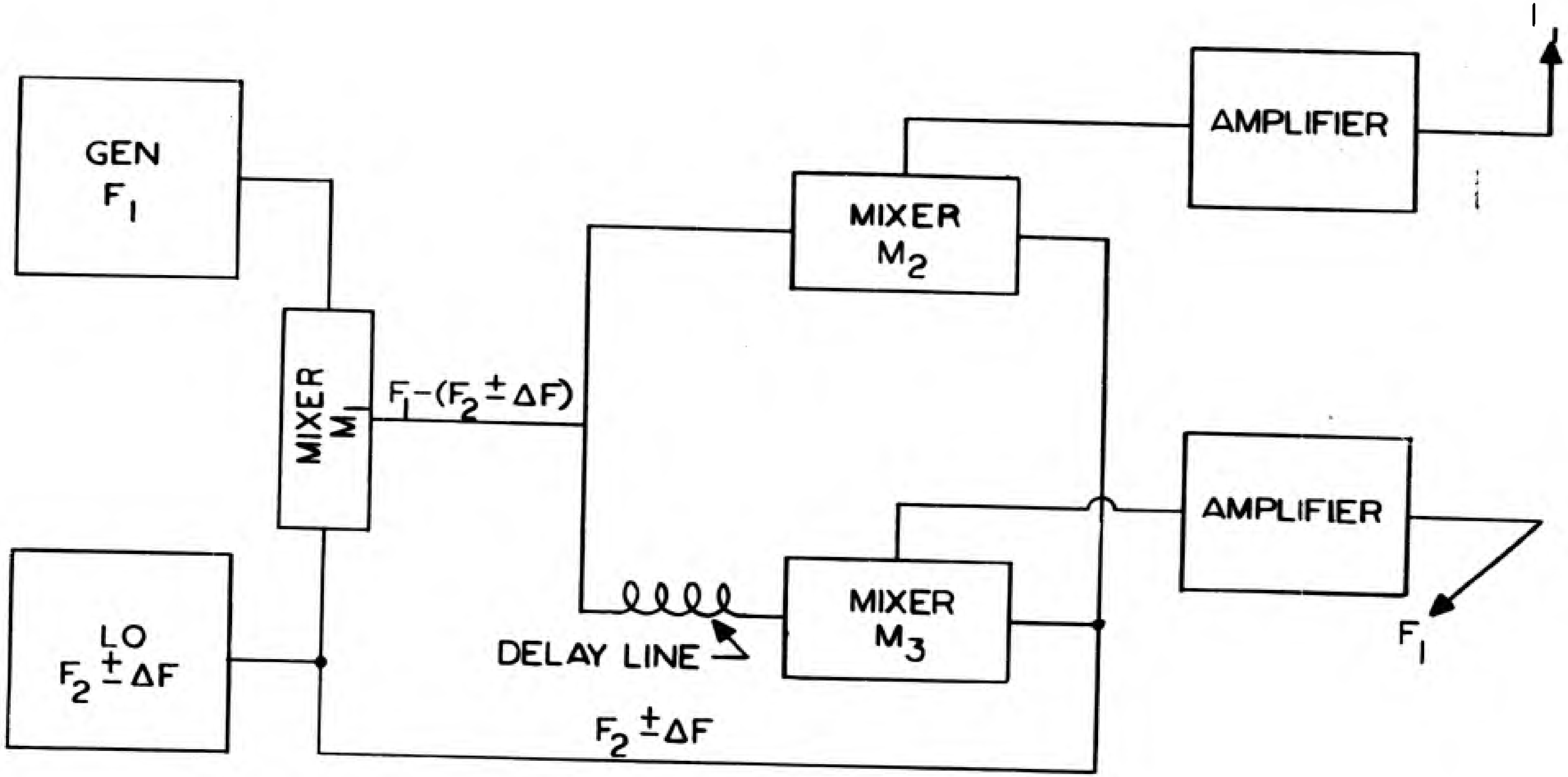


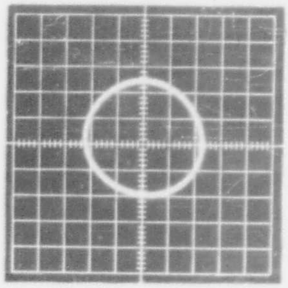
Figure 22. Polarization Modulation Using Frequency Modulated Local Oscillator

of transmitter (F_1) and a local oscillator ($F_2 \pm \Delta F$) are combined in the first mixer (M_1). The beat frequency $F_1 - (F_2 \pm \Delta F)$ is taken at the output of M_1 and the second part feeds through a delay line to a third mixer (M_3). The local oscillator frequency ($F_2 \pm \Delta F$) is then combined with the intermediate frequency on M_2 and M_3. Among the resultant output frequencies of M_2 and M_3 will be $F_1 - (F_2 \pm \Delta F) + (F_2 \pm \Delta F)$, which is exactly the frequency F_1. These two outputs are then amplified through separate amplifiers and connected to orthogonal linear radiators.

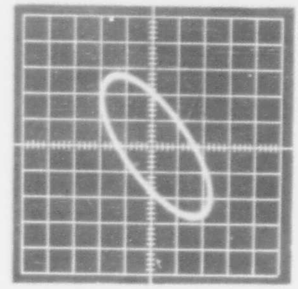
The frequency modulation of the LO provides a relative phase change between the two outputs which is determined by the modulation index and the length of the delay line. The output frequency of the system is changed by changing F_1.

The complete family of ellipses can be generated electronically by this method while keeping the transmitter output frequency within the bandpass of a sharply tuned receiver.

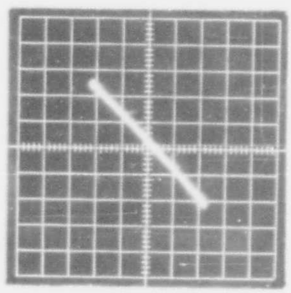
Bench measurements were performed using existing components and the technique outlined above. The outputs of mixer M_2 were connected to the Y axis and those of M_3 to the X axis of an oscilloscope. The frequency of the LO was changed manually through a frequency range so that the time delay of the delay line was equal to one period of frequency change. The resultant change in relative phase between the two outputs is shown in the oscillograms of figure 23.



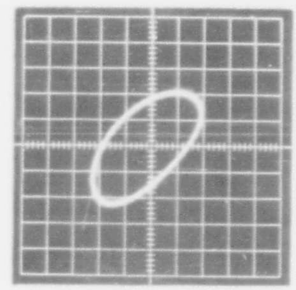
(a) +90° Relative Phase Difference



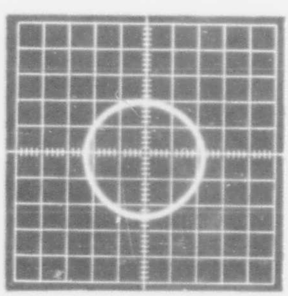
(b) +45° Relative Phase Difference



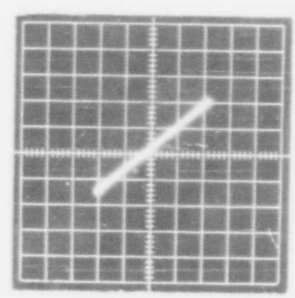
(c) 0° Relative Phase Difference



(d) -45° Relative Phase Difference



(e) -90° Relative Phase Difference



(f) -180° Relative Phase Difference

Figure 23. Polarization Modulation Using Frequency Modulated Local Oscillator

4.5 Traveling-Wave Tube Method

The traveling-wave tube, while offering many desirable features for use in airborne systems, has been to a great extent, restricted to ground-based applications because of the weight associated with its focusing structures. Recent development of periodic permanent magnet focusing structures[8] has done much to remove this restriction and currently many tubes, adaptable to airborne systems, are available.

A particular feature of the traveling tube that makes it especially useful in generating a polarization modulated wave is its ability to act as a phase modulator. This is usually accomplished by impressing a small modulating signal on the slow wave structure voltage. The general expression for a phase-modulated signal is given by:

$$E = \sin \left[w_o t + m_p\, g\,(t) \right] \tag{4.19}$$

where $w_o = 2\pi f_o$ and f_o is the unmodulated carrier frequency, m_p is the phase modulation index in radians, and g(t) is the modulating voltage function. A block diagram of a typical system for generating a polarization modulated using traveling wave tubes is shown in figure 24. When the modulating wave form of 24(a) is applied to TWT-1 and m_p is made equal to 2π radians, the signal appearing at antenna "a" is slightly different in frequency from that at antenna "b" and the transmitted signal is multi-polarized. To prove this analytically:

from equation (4.9)

$$E = \sin \left[w_o t + 2\pi g\,(t) \right]$$

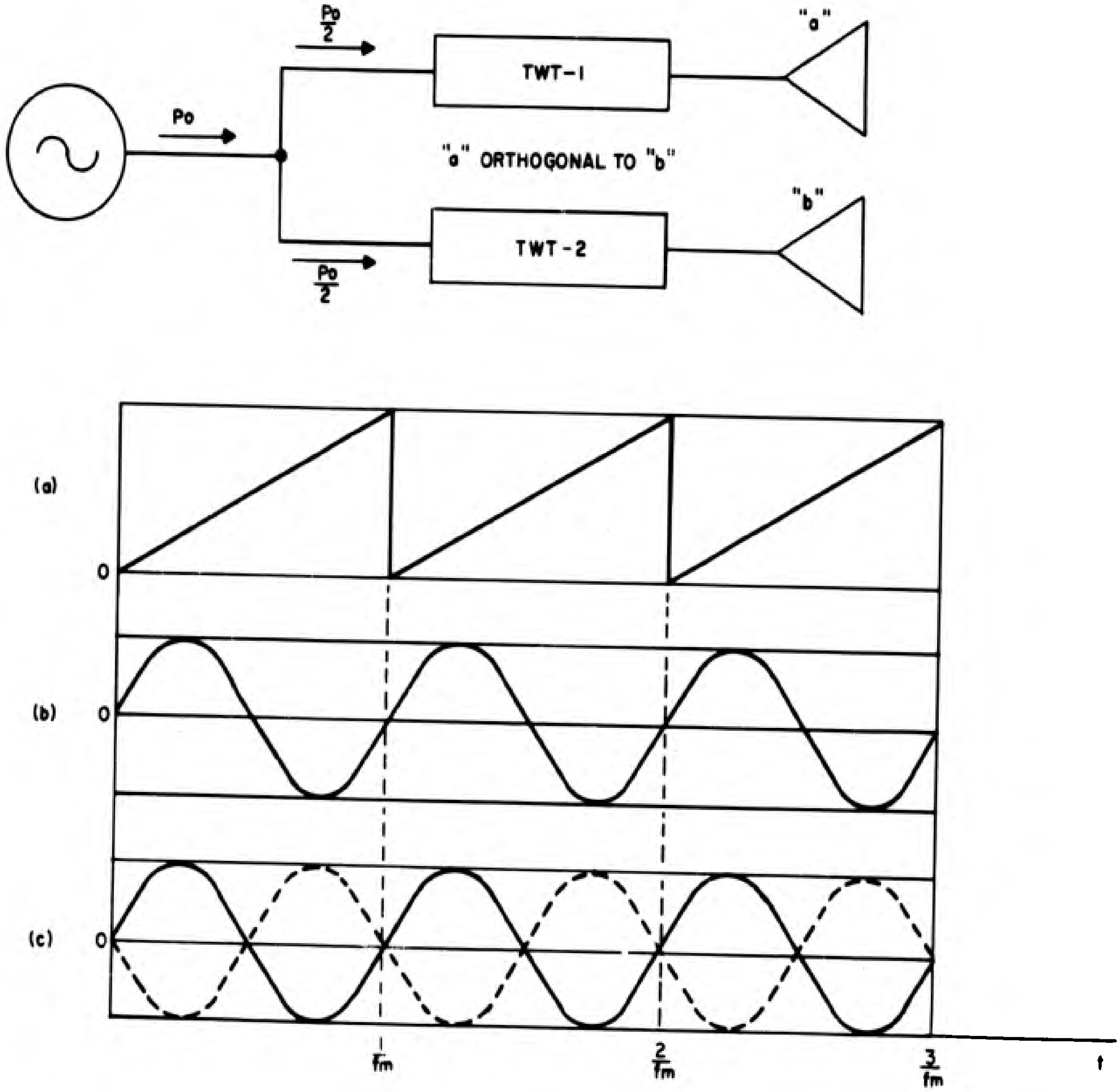


Figure 24. Block Diagram of Polarization System Using TWT's with Associated Modulation Functions

but g(t), during the first period of the modulating signal, is simply $f_m t$, where f_m is repetition rate or frequency of the sawtooth function.

Therefore, during the first period of the modulating signal:

$$E = \sin \left[w_o t + 2\pi f_m t \right] \tag{4.20}$$

or $$E = \sin \left[(w_o + w_m)\, t \right]$$

and frequency has been shifted from f_o to $f_o + f_m$. By similar analysis, it can be shownthat this is true for all subsequent periods of the modulating signal, and the frequency fed to antenna "a" does differ from that fed to antenna "b" by an amount f_m. Consequently, the polarization is made to change at the rate of the difference frequency f_m.

When the modulating waveform of .(24b) is applied to TWT-1 and m_p is again 2 radians, the relative phase between antennas "a" and "b" changes thru 360° and a polarization modulated wave similar to that described in section (4.3) is generated. Because bandwidth is directly proportional to the frequency of modulation, the rate of polarization change of this method is limited by victim receiver bandwidth. A good approximation[6] of this limiting factor is given by:

$$B_W \simeq 2 f_m (1 + m_p) \tag{4.21}$$

and in this case $$BW \simeq 2 f_m (1 + 2\pi) \tag{4.22}$$

where B_W is victim receiver bandwidth and f_m is phase modulating frequency.

A possible way of minimizing this bandwidth problem would involve modulating both TWT-1 and TWT-2. The function represented by the solid line of figure 24c would be used to modulate TWT-1 while the function represented by the dashed line would be used to modulate TWT-2. Modulation index of both tubes would be made equal to radians. The signals arriving at antennas "a" and "b" are respectively:

$$E_a = \sin\left[w_o t + \pi \sin w_m t\right] \qquad (4.23)$$

and

$$E_b = \sin\left[wot + \pi \sin (w_m t + 180°)\right] \qquad (4.24)$$

and the instantaneous phase difference between the signals would be

$$\Delta \phi = \pi\left[\sin \omega_m t - \sin (\omega_m t + 180°)\right]$$

or

$$\Delta \phi = 2\pi \sin \omega_m t \qquad (4.25)$$

Thus, in each half cycle of the modulating frequency, relative phase changes thru 360° and a polarization modulated wave are again generated. Limitation of bandwidth in this case is approximately.

$$B_w \simeq 2 f_m (1 + \pi)$$

4.5.1 Limitations to Traveling-Wave Tube System

Several design problems are associated with the implementation of the models above into practical systems. The problems and possible solutions are presented below.

In phase modulating traveling-wave tubes, some incidental amplitude modulation would be encountered, this modulation in most cases would not be serious, provided, the modulation index is below 2 π radians. In cases where this would be a problem two possible methods of resolution are (a) operation of the tubes near saturated signal level, thereby minimizing incidental amplitude modulation, and (b) compensory amplitude modulation could be introduced by means of a TWT control grid, or in the input RF signal.

A more difficult problem is encountered when it is desired to phase modulate over a frequency band, because the optimum slow wave structure voltage varies with frequency. This difficulty could be compensated for, by electronic control of the slow wave structure voltage, as a function of frequency.

4.6 Miscellaneous Dual-Frequency Techniques

Three other methods (ferrites, balanced modulators, and a doppler wheel) had serious limitations and are not recommended for polarization modulation systems.

The theoretical discussions of section (4.5), while concerned in particular with phase modulated traveling-wave tubes, are equally applicable to any phase-modulator device. In conjunction with this, a literature search was conducted into the possibilities of using ferrite devices in polarization modulation systems. It was found that three serious limitations would be placed on systems using ferrite devices: (1) High modulation power would be required, (2) only a narrow frequency band could be covered by a single device, and (3) there would be excessive attenuation at high-power levels over much of the spectrum.

The second method considered employed a balanced modulator to generate two signals of different frequency. It was concluded that the development effort required to achieve a usable balanced modulator at even low microwave frequencies would be prohibitive in regard to the scope of the contractural program. Another disadvantage was the high loss accompanying the generation of the two signals of different frequency, which would necessitate the use of power amplifiers. Because it was already demonstrated that power amplifiers (see TWT Section) alone could be used in generating a polarization modulated wave, the balanced modulator technique was not given further consideration.

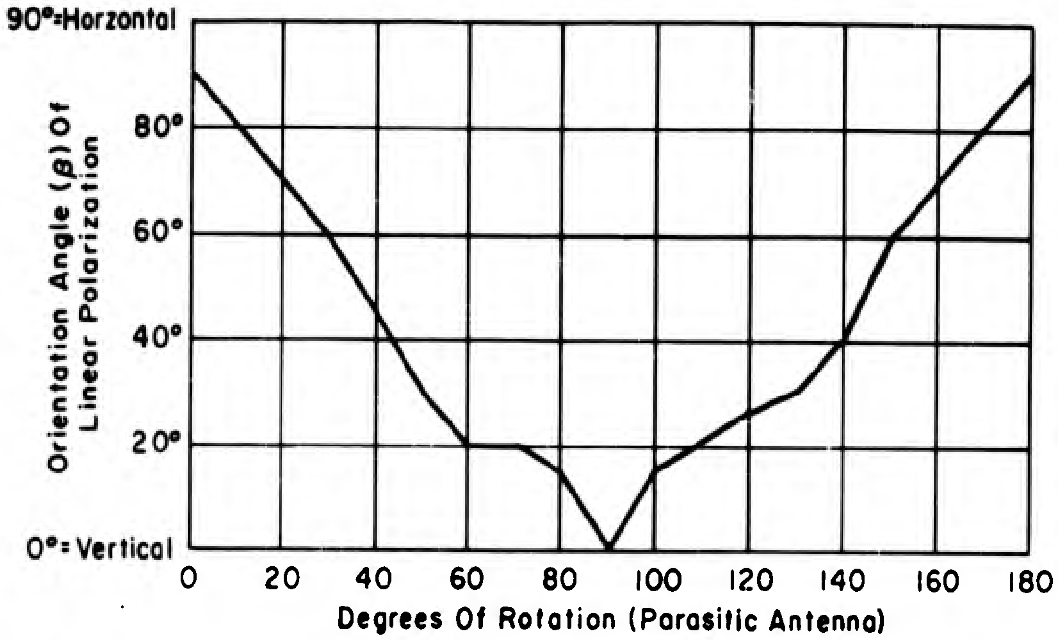
Another technique considers the use of a "Doppler Wheel" to create two signals of different frequency. This was the subject of some discussion in the sixth quarterly report. A review of work done by other laboratories at Melpar has shown that the critical mechanical tolerances required and the losses inherent in such a device would impose severe limitations on its use in polarization modulation systems. In addition, the technique described requires the use of two such devices, thereby doubling the problem involved. Based on this reasoning the "Doppler Wheel" is not considered practical for other than laboratory use.

5.0 PARASITIC SPIRAL TECHNIQUE

This technique employs two oppositely sensed spiral antennas (mechanically identical) placed so their axes coincide. In this technique, one spiral is active and is fed to produce the axial mode of radiation (maximum radiation along the spiral axis) and the second spiral acts as a parasite to the active element. Because the spirals are oppositely sensed and have a common axis, the resultant polarization of this configuration is linear. The resultant linear polarization is made to rotate by revolving one of the elements (active or parasitic) with respect to the other about the common axis. If the rotating element revolves at w radians per second, the linear polarization will rotate through all orientations of linear polarizations in 1/w seconds. Using this technique, experimental data was taken and is shown in figure 25 (top).

When the active element is fed to produce the normal mode of radiation (cone shaped radiation with the maximum displaced, approximately, $\pm 60°$ from the spiral axis), practical spirals do not have the necessary ellipticity and omnidirectivity to achieve a rotating linear polarization. No attempt was made during this program to determine analytically what polarizations could be obtained when radiating the normal mode of radiation. However, measurements of the resultant polarization, with the active element radiating the normal mode, were made at several frequencies; data is shown in figure 25 (bottom).

The peak power breakdown of several cavity-fed spiral antennas was measured experimentally at Melpar and found to break down at rather low power levels. The results of these tests are shown in figure 26. Thus, the low power-handling capability of the spiral is a limiting factor when using the parasitic spiral technique. The rate of polarization change acheivable is also a limiting factor, because either the active or parasitic element must be mechanically rotated. Because of the low power-handling capability of spirals and the above mechanical limitation, this technique is not recommended for a practical high power transmitting system of polarization modulation. However, this technique is applicable to low power transmitting and receiving systems where polarization modulation is desirable.



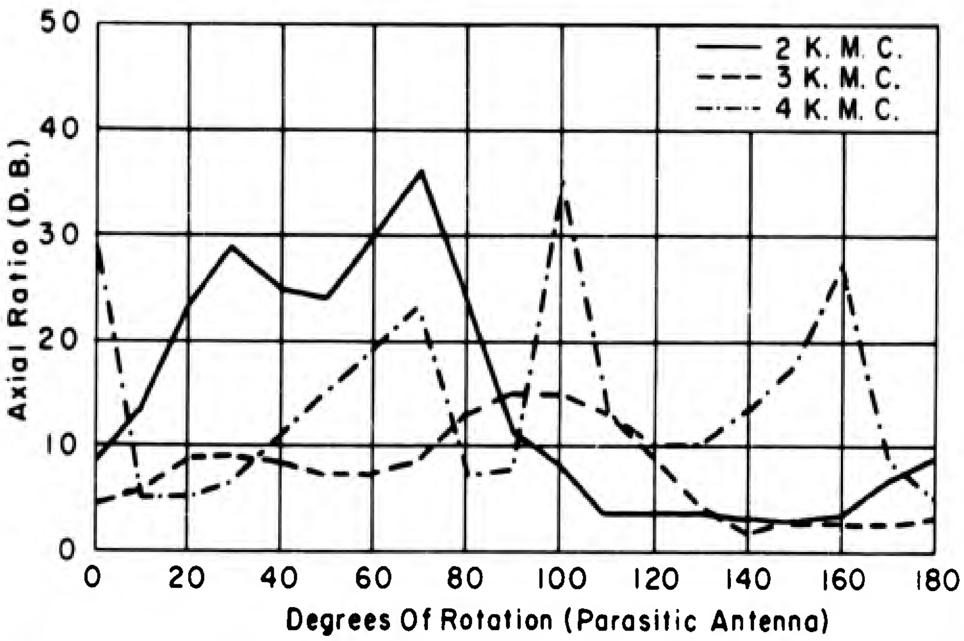


Figure 25. Rotation of Parasitic Spiral Showing Rotation of Linear Polarization for Axial Mode (top) and Variable Elliptical Polarization for Normal Mode of Radiation (bottom)

Breakdown Power (Average) Watts	No. Of Turns In Spiral	Type Of Dielectric	Duration Of Signal μ Sec	Signal Recurance Rate Cyc/Sec	Duty Cycle	Peak Breakdown Power (Kilowatts)
36.4	6	Insurok	2	1000	.002	18.2
37.0	6	Epoxy	2	1000	.002	18.5
45.0	6	Teflon	2	1000	.002	22.5
25.0	12	Epoxy	2	1000	.002	12.5

Figure 26. Typical Measured Spiral Peak-Power Breakdown Values

6.0 MISCELLANEOUS TECHNIQUES

Other techniques briefly investigated included: a) rapid switching between differently polarized radiators, b) physical rotation of an antenna about its axis of propagation, and c) a scanning technique in which an omnidirectional antenna with large variations in polarization as a function of pattern angle, is continuously rotated. In a) the switches must be capable of handling high r-f power at high switching rates for relatively long periods of time. In b) and c), complex drive mechanisms and high-speed rotary joints would be required. The associated design problems were considered beyond the scope of this program and, consequently, no further consideration was given to these methods.

7.0 CONCLUSIONS AND RECOMMENDATIONS

Several mechanical and electronic techniques for generating a polarization modulated wave have been investigated. There are advantages and disadvantages associated with each of these techniques; they are discussed below, along with recommendations.

The electronic technique which offers the most promise for implementation into an operational system is the dual-frequency technique. Of the four methods employing this technique (two generator, delay line, heterodyne and traveling-wave tube), the two-generator techniques, at present, seems most feasible because the associated control problem is not too difficult. The delay-line technique is hampered by the high loss associated with the long delay line involved. A recent survey of progress in development of high-frequency delay components[13] does not show, presently existing or soon to be developed, components that will substantially reduce the inherent loss. Future work on the delay-line technique should be restricted to low-frequency applications until suitable high-frequency delay components are developed. The heterodyne technique, because of the circuitry involved, should also be restricted to lower frequency applications, where proper circuitry is available. The traveling-wave tube technique shows much promise for future work. Future work should be concerned with the development of traveling-wave tubes having ideal parameters in terms of phase modulation and the design of switchable control circuitry.

Both of the mechanical techniques investigated (rotating quarter-wave plate and parasitic spirals) had a common disadvantage. This was the limited rotational speeds possible. Also, in the case of the rotating quarter-wave plate, it was learned that an improved bearing would be required to support and permit the quarter-wave plate to be rotated at a maximum rate.

Therefore, it was concluded, as a result of this program, that future work performed toward obtaining a rapidly varying polarization should be directed toward obtaining improved electronic techniques. It is also believed that this program has demonstrated that several techniques are available for obtaining a rapidly varying polarization, and that additional studies should be conducted to make those improvements that have been suggested in this report.

8.0 PERSONNEL

The following technical people performed work on the project:

H. H. Hibbs	Project Engineer
S. R. Jones	Consulting Project Engineer
E. F. Henry	Principal Engineer
M. N. Fullilove	Senior Engineer
J. R. Tomlinson	Senior Engineer
J. A. Jones	Engineer
S. Perrino	Engineer
H. C. Oliver	Engineer
J. E. Ferris	Engineer
H. K. Deloatche	Junior Engineer
C. A. White	Junior Engineer
E. Kelly	Technician

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APPENDIX A

Graphical Solution Technique

Polarization studies can be greatly simplified by use of a polarization chart. One such chart, figure 27, a projection of the Poincaré sphere[2], was used exclusively during this study. Each point on the chart represents a particular state of elliptical polarization and its equivalent of two orthogonal linear polarizations of the same frequency. Five quantities are associated with each point and are: 1) sense, 2) axial ratio, 3) tilt angle of the ellipse, 4) phase difference of the linear waves and 5) amplitude ratio of the linear waves. If the two linear waves are assumed of the form:

$$E_x = A \sin \omega t$$

$$E_y = B \sin (\omega t + \phi)$$

and are made parallel to the "X" and "Y" axes of a right hand XYZ coordinate system, with propagation along the positive Z axis, four parameters must be plotted (sense being specified by ϕ as shown below) and are uniquely defined as:

1. IRE Proceedings, Vol. 39, pages 540-544, May 1951

ϕ = phase angle by which E_y leads E_x; plotted as an absolute value; if positive, the generated ellipse has LH sense; if negative, RH sense.

$\gamma = \tan^{-1} \frac{B}{A}$ or $\tan^{-1} \frac{A}{B}$ so that $0 \leq \gamma \leq 45$, where $\frac{B}{A}$ and $\frac{A}{B}$ are the ratio of voltage maximums of E_x and E_y.

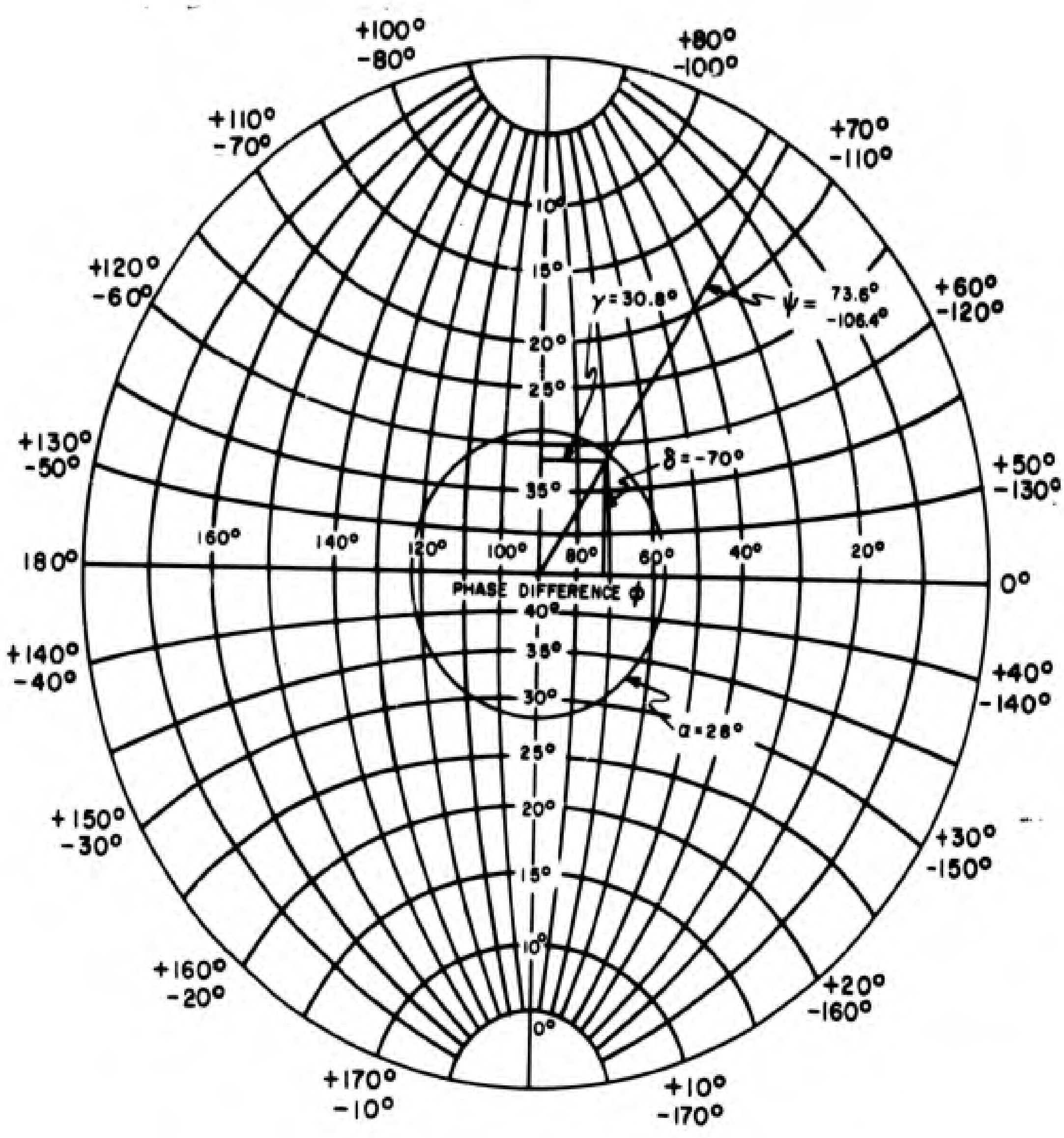

Figure 27. Polarization Chart

Ψ = angle major axis of polarization ellipse makes with +X axis*, because of symmetry $\Psi = \Psi \pm 180°$.

$\alpha = \tan^{-1} \dfrac{\text{ellipse minor axis}}{\text{ellipse major axis}}$, where $\dfrac{\text{minor axis}}{\text{major axis}}$ is the axial ratio of the ellipse.

* Is positive counter-clockwise when observed from the positive Z axis looking toward the origin.

APPENDIX B

Polarization Adjustment

In another study recently conducted at Melpar[14], a technique was demonstrated that rapidly adjusts the polarization of a two-element antenna. This study is summarized below.

As shown in previous sections of this report, it is possible, thru judicious variation of relative phase and amplitude between two cross polarized elements of an antenna, to generate all polarizations. It was also pointed out that an antenna consisting of two oppositely sensed circular elements would be more advantageous in this application, than an antenna consisting of two orthogonal linear elements. Thus, the antennas used in this technique were comprised of two oppositely sensed circular elements.

Incorporation of this antenna into a practical system is not necessarily a simple matter, because the two elements must be independently fed and their phase centers must be spacially close together. The several different antenna types investigated for their applicability to this technique included: 1) a pair of oppositely sensed spirals, 2) a dual feed Archimedean Spiral, and 3) a circularly polarized horn.

The basic system consisted of a variable phase shifter and an available power divider, incorporated into the antenna circuit as in figure 28. A second variable phase shifter, together with a hybrid junction, comprised the variable power divider. Any type of 180° or 90° hybrid junction could be used, but a more compact unit results when a short-slot junction as described by Riblet is used.

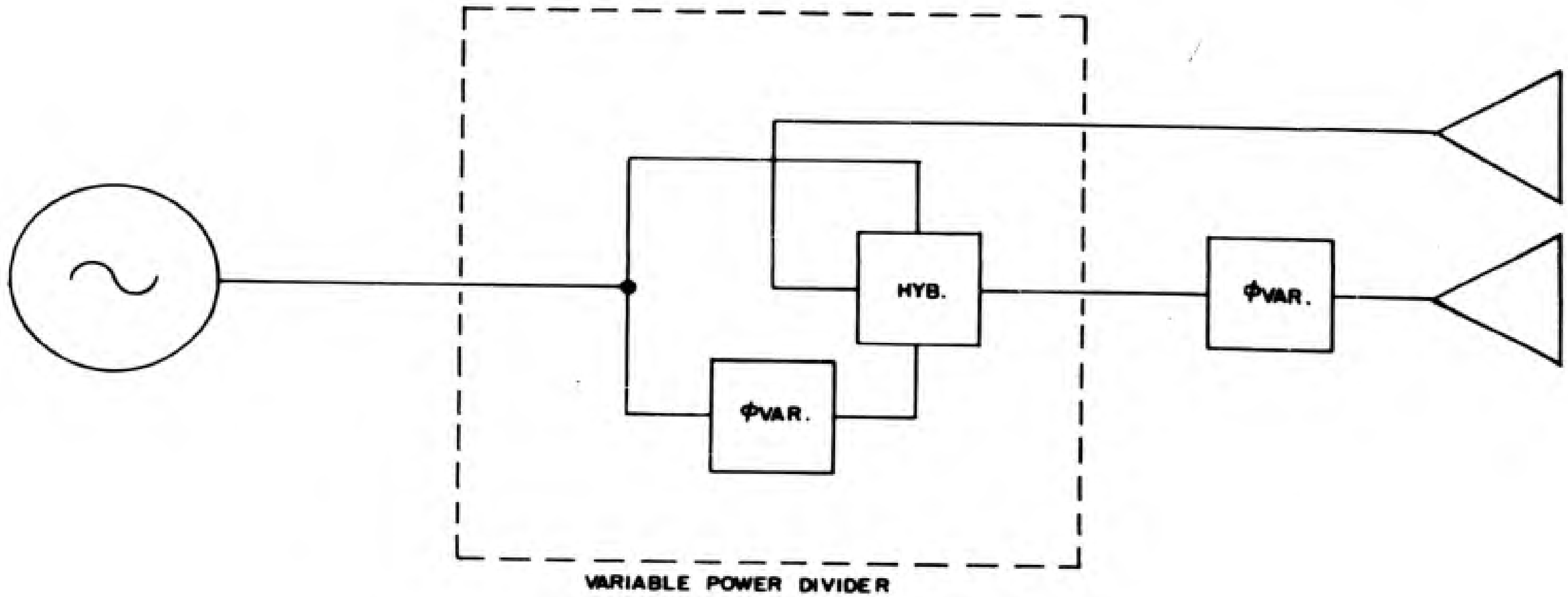

Figure 28. Block Diagram of Polarization Adjustment System

Attempts to use two oppositely sensed spirals, each fed separately, met with limited success. Although reasonably good performance was attained at some frequencies, suitable performance over a band was difficult to obtain because of the adverse effects of mutual coupling between the two elements.

The dual-feed Archimedian spiral was considered next. Similar radiation patterns exist whether the spiral is excited from the center or outer edge, but a change in feed point is accompanied by a reversal in sense. The two feed points are well isolated from each other and it is possible to introduce a signal at both points simultaneously. Polarization adjustment is made thru variation of relative phase and amplitude between the two signals. There was some difficulty encountered with radiation from the asymmetrically located outer feed, but this was resolved by locating a conducting ground plane very closely behind the spiral. Good performance was attained across a fairly broad frequency band.

Another type of antenna considered was a circularly polarized horn in which a dielectric quarter wave plate converted two orthogonal linear modes, incident upon it, to two cross-polarized circular modes. This was accomplished by orienting the linear modes at $\pm 45°$ to the quarter-wave plate. In evaluation of this method, a shunt and series fed circular waveguide horn was utilized. Performance was very good over a somewhat reduced band. The most severe bandwidth limitation was imposed on this method by available ways of coupling in the two linear modes and a suggested means for coupling in the two linear modes over a broader band is suggested. This alternate method uses a "Y" junction to feed in the two linear modes from two ridged waveguides. Further details of this technique can be found in reference 14.

The techniques presented offer improved efficient thru polarization control to many present-day antenna systems. Typical applications would be in countermeasure and counter-countermeasure systems or in communication thru the ionosphere where Faraday rotation can change the polarization of a wave. With little modification, the techniques could also be used as a polarization monitor system, where unknown polarization incident to a receiving antenna would be given by two phase measurements. One measurement giving tilt angle directly while the other would give axial ratio.

On the chart, constant ϕ is represented by arcs thru the ϕ axis; constant δ by arcs thru the δ - α axis (the portion of the arc above the ϕ axis is chosen if B > A and that below it if A > B); constant Ψ by radial lines from the chart center thru the circumference; constant α by circles which center at the chart center, thru the δ - α axis.

It should be noted that circular polarization is represented by the point $\phi = \pm 90, \delta = \alpha = 45°$ and linear polarization by points on the circumference.

An example will help clarify the plotting procedure.

(1) The polarization ellipse generated by the two orthogonal linear vectors

$$E_x = 3 \sin \omega t$$

$$E_y = 5 \sin (\omega t - 70°)$$

is found as follows:

(a) $\phi = -70°$ is plotted as an arc thru the horizontal ϕ axis.

(b) $$\gamma = \tan^{-1}\frac{A}{B} = \tan^{-1}\frac{3}{5} = 30.8°$$

is plotted as an arc thru the vertical $\gamma - \alpha$ axis, above the ϕ axis since B > A.

(c) Point (c), the intersection of arc (a) and arc (b), lies on the radial line $\Psi = 73.6°$ or $-106.4°$.

(d) A circle with center at the projection's center thru point C crosses the $\gamma - \alpha$ axis at $\alpha = 28°$.

$$AR = \frac{\text{minor axis}}{\text{major axis}} = \tan 28° = 0.532$$

Thus, the two linear vectors give rise to a polarization ellipse whose major axis is tilted 73.6° above the X axis, having an axial ratio of 0.532 and, as ϕ is negative, of RH sense.

Prepared by

J. E. Ferris
Engineer

Submitted by

H. H. Hibbs
Project Engineer

Approved by

R. W. Masters
Manager, Antenna Laboratory

Approved by

W. C. Purple, Jr.
Vice President
Engineering and Manufacturing

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