USER MANUAL
FOR THE MICROWAVE RANGE OF THE NATR
AT PAARDEFONTEIN

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for

GEROTEK

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ABBREVIATIONS, PREFIXES AND SYMBOLS

A	ampere
A	longitudinal amplitude taper
AF	antenna factor
AR	axial ratio
A_e	effective aperture
A_p	physical aperture
AUT	antenna under test
BW	beamwidth (half power unless stated)
COMINT	communications intelligence
CP	circularly polarized, circular polarization
CW	continuous wave
D	largest aperture dimension of AUT, m
DC	direct current
DF	direction finding, direction finder
d	diameter of source antenna, m
dB	decibel = 10 log (P_2/P_1)
dBi	decibels over isotropic
dBci	decibels over circularly polarized isotropic
dBli	decibels over linearly polarized isotropic
ELINT	electronic intelligence
E_i	incident electric field intensity, Vm^{-1}
E_n	normalized electric field intensity, Vm^{-1}
E_r	reflected electric field intensity, Vm^{-1}
F	force, N
f	frequency, Hz
G	giga = 10^9 (prefix)
G	gain
G_r	gain of test antenna
G_s	gain of standard antenna

g	gram
g_h	partial gain for horizontal polarization
g_v	partial gain for vertical polarization
H	horizontal
H	magnetic field, Am^{-1}
HPBW	half power beamwidth
HF	high frequency
Hz	hertz = 1 cycle per second
h_p	height of positioner above ground, m
h_r	height of centre of AUT above ground, m
h_t	height of centre of transmit antenna above ground, m
	apparent height of centre of transmit antenna above ground, m
ICR	integrated cancellation ratio
k	kilo = 10^3 (prefix)
k	ground reflection coefficient
kg	kilogram
kW	kilowatt
L	radar band, nominally 1 – 2 GHz
L	length of end-fire antenna, m
LHCP	left hand circularly polarized
LPDA	log-periodic dipole array
M	mega = 10^6 (prefix)
m	metre
m	milli = 10^{-3} (prefix)
N	newton
Nm	newton metre
NATR	National Antenna Test Range
P	polarization
P	power, W
P_i	power incident on antenna terminals, W
P_r	power reflected at antenna terminals, W
R	test range or test distance, m

RCS	radar cross section
RHCP	right hand circularly polarized
r	radius, m
rms	root mean square
RWR	radar warning receiver
S	radar band, nominally 2 – 4 GHz
SG	standard gain
SGA	standard gain antenna
SL	sidelobe
s	second
UHF	ultra high frequency
V	vertical
V	volt
VHF	very high frequency
VNA	vector network analyzer
VSWR	voltage standing wave ratio
V_i	incident voltage, V
V_r	reflected voltage, received voltage, V
W	watt
X	radar band, nominally 8 – 12 GHz
α	(alpha) subtended angle of AUT at range R
α_n	(alpha) depression angle from source antenna to base of test tower
Γ	(gamma) voltage reflection coefficient
ϵ_{ap}	(epsilon) aperture efficiency
λ	(lambda) wavelength, m
π	(pi) = 3.1416…
θ_{hp}	(theta) half power beamwidth

CAUTION – ELECTROSTATIC DISCHARGE (ESD)

ESD can damage the highly sensitive circuits in the VNA. ESD is most likely to occur as test devices are connected to or disconnected from the VNA test ports. The instrumentation can be protected by wearing a grounded static–discharge arm band. Alternately, the operator must ground himself by touching grounded metallic parts of the equipment (e.g. the outer chassis of the grounded VNA, the grounded test positioner, etc) before touching the VNA test ports or test cables connected to the VNA. Operators must avoid touching the centre conductors of test port connectors or test cables connected to the VNA unless they are properly grounded and have eliminated the possibility of static discharge. Ideally the centre conductors should not be touched if they are connected to the test equipment.

Outdoor measurements using a sensitive vector network analyser (VNA) or other test receiver are particularly susceptible to static and/or lightning damage to the test equipment. Users must exercise all reasonable caution to eliminate damage to test equipment. Long coaxial cables can build up significant static charge on the outer jacket of the cable because of wind friction. Once the antenna or system to be tested is installed, the coaxial cables at the VNA/receiver end must be short-circuited to discharge the cable. It is good practice to install a 3 dB fixed attenuator at the output of the test cable. This reduces the dynamic range by 3 dB but the attenuator has a DC path to ground which provides continuous bleeding of charge build up to ground. This advantage far outweighs the loss of 3 dB of sensitivity. An additional safety precaution will be to add a DC block between the 3 dB attenuator and the test port. Custom high inductance circuits which provide a DC path to ground can also be used. The DC block and inductance circuits may have an effect on the sensitivity of the VNA and should be tested over the frequency ranges of interest before starting the measurement programme.

WARNINGS

1. The user must ensure that the earth connections of his test equipment are in place and that there are no 'floating earths'. Connect all equipment via a 3-pin grounded power outlet. If a grounded 3-pin outlet is not available, use a conversion adapter and connect the equipment ground terminal to ground. If mains power is supplied without grounding the equipment there is a risk of equipment damage and/or electrical shock to operators.
2. Do not exceed the rated input RF power levels of the VNA, particularly when power amplifiers are being used. The maximum rated power levels and DC voltages are normally stated on labels near the test ports. Typical values are + 27 dBm max and 40 VDC max. If there is uncertainty, consult the manufacturer's operating manuals.

1. INTRODUCTION

The microwave range at the NATR is 500 m long and was originally designed as a test range for radar antennas operating above 500 MHz. The site at Paardefontein is ideally suited for a measurement facility with 500 m test distance. Undesired reflections in the azimuth plane can be controlled by having source antennas with narrow enough beamwidths so that reflecting objects are not illuminated. In the elevation plane reflections from the ground are always there because at long test distances the beamwidths of the source antennas cannot be made narrow enough to eliminate the ground reflections. It is possible to make an elevated range with large test distances by constructing the range across a deep valley. This is not possible at the Paardefontein site and the test range is designed to use the reflected wave. The microwave range is a "ground-reflection range" where the reflected field from the range surface is used together with the direct field from the source antenna to create the desired test field at the antenna under test (AUT). This is in contrast to outdoor "free-space ranges" where attempts are made to eliminate or reduce the ground reflection.

With proper design, the illuminating field in the test zone will have small essentially symmetric amplitude and phase tapers. The desired tapers are obtained by controlling the range geometry and source antenna characteristics. To obtain the largest possible test region at the receive tower, the height of the source antenna must be adjusted to ensure that the direct and reflected fields add constructively at the centre of the test zone. The source height is inversely proportional to frequency and as the frequency increases the source antenna must be moved closer to the ground. This can be done by means of a hand-held controller at the source tower which mechanically raises or lowers the source antenna. Because of its design parameters the Microwave Range can produce a test field with an amplitude taper of less than 0.5 dB over a 5.5 m high vertical aperture. By selecting the source antenna appropriately, the 0.5 dB amplitude taper is also achieved in the azimuth plane over an aperture at least 5.5 m wide. This 5.5 m x 5.5 m area is a very large test zone with such a small field taper.

This user manual discusses the range layout and the source antenna configurations to obtain the desired test field in amplitude and phase at the required test frequencies. The test field determines the size and frequency at which antennas can be measured for particular sidelobe and gain accuracies. Measurements of cross polarization of antennas and circularly polarized antennas can be done with good accuracy.

The design considerations and measured field distributions for the Microwave Range have been documented in reference 1. In his chapter on Antenna Measurement Ranges which is presented in reference 2, Dr. E. S. Gillespie states: "Reference 17 contains an excellent example of a ground-reflection range". His reference 17 is reference 1 of this user manual. The Microwave Range at the NATR is a world-class antenna test facility.

NOTE: In recent years the Microwave Range has also been used extensively to test VHF/ UHF antennas and antenna systems. The discussions in: "User manual for the VHF/UHF range of the NATR at Paardefontein" apply equally to the Microwave Range and are not repeated in this manual.

2. TEST RANGE CONFIGURATION

2.1 Test range lay out and principles of operation

A plan view of the microwave range is shown in Figure 2.1. The main receive tower is at the operations building and the main source tower is 500 m down range. The primary range surface is 100 m wide with a nominal rms surface accuracy of 25 mm. There are no nearby fences or reflecting surfaces which can introduce undesired reflections. The operations building is set an angle to the range axis so that scattering off the building is away from the receive tower.

The ground-reflected wave must add in phase with the direct wave from the source antenna. This is shown in Figure 2.2 (the scale is compressed in the horizontal axis for purposes of illustration). The height of the source above ground is adjusted to get the in-phase condition with the specular reflection. At angles of incidence near grazing (which is the case at the 500 m microwave range) the reflection phase for typical soil conditions is nearly 180° for both vertical and horizontal polarizations. Thus the height of the transmit antenna (h_t) is adjusted to add one half wavelength of path distance between the direct and reflected rays. It can be shown [3] and [4] that the transmit and receive antenna heights for the most uniform field distribution in the vertical plane are related by

$$h_t = \frac{\lambda R}{4 h_r} \qquad (2.1)$$

where

- h_t = height at centre of source antenna (m)
- λ = wavelength (m) = 0.3/f
- f = frequency in GHz
- R = range length = 500 m
- h_r = height at centre of AUT (m)
  = h_p + ½ D
- h_p = height of upper azimuth positioner = 11.2 m
- D = vertical aperture of antenna under test

Equation 2.1 shows that h_t is a function of frequency and ideally the height of the source antenna should be adjusted for each frequency. Measurements [1] have shown that a 0.5 dB amplitude taper in the vertical plane is achieved over a vertical aperture about 5.5 m high placed symmetrically around the centre of the test aperture. Thus, if the antenna aperture of the AUT is 1 or 2 m, the source height is not critical over a typical waveguide band.

The test field distribution in the horizontal plane is determined entirely by the pattern of the source antenna in the horizontal plane. For a 5.5 m horizontal aperture of the AUT, the total subtended angle of the source antenna at a distance of 500 m is only 0.63 degrees. Thus, as long as the source antenna gain is not too high, a 0.25 dB taper in the horizontal plane is easily achieved.

At the normal far-field range $R = 2D^2/\lambda$, it can be shown [3] that the diameter d of the source dish must satisfy

$$d \leq 0.74\, D \tag{2.2}$$

where D is the diameter of the AUT.

The above discussion has been based on the assumption that the reflection coefficient for both V and H polarizations off the range surface is -1. In practice, depending on the type of soil and moisture conditions, the magnitude of the reflection coefficient could differ from this. A value of around 0.9 has been observed at the microwave range in L and S bands [1]. This value shifts the symmetrical peak of the test field in elevation by a small amount. This is not critical unless the full test aperture of 5.5 m is to be used. For smaller antennas the effect is insignificant.

For a reflection coefficient of -1, the effective source position is on the ground (i.e. half way between the source antenna and its image in the ground plane). For a magnitude of the reflection coefficient of k, the apparent source height is directly below the source at an apparent source height $h_t^{'}$ above the range surface given by [5]

$$h_t^{'} = \left(\frac{1-k}{1+k}\right) h_t \tag{2.3}$$

If k = 0.9, $h_t^{'}$ = 0.053 h_t. At 3 GHz with h_r = 11.2 m h_t = 1 116 mm and $h_t^{'}$ = 59 mm. At 500 m this represents an insignificant angle; the effective source position is on the ground directly below the source antenna.

With h_t as in Equation 2.1 it can be shown that the total electric field in the test zone follows a cosine distribution. The normalised field strength E_n is given by [5]

$$E_n = \cos\frac{\pi}{2}\frac{x}{h_r} \tag{2.4}$$

where x is the displacement from the centre of the test aperture at height h_r.

Figure 2.3(a) shows the electric field over the test zone with the source at height h_t (solid line) and at height $3h_t$ (dashed line). It is clear that first field maximum is much wider than the second maximum at $3h_t$. For large antennas it is important that the source height is set to exploit the wider field distribution achieved at h_t. Figure 2.3(b) is a useful curve which gives the amplitude taper over an antenna of diameter D as a function of the ratio D/h_r. For 0.25 dB amplitude taper in the vertical plane we find that the tower height must be about 3.3 times the diameter of the dish. At 5 m diameter (a typical S-band radar antenna) this will require a 16.5 m tower. This was deemed excessive since a crane would be required for mounting even quite small antennas. A tower 11.2 m high gives a field taper of about 0.5 dB over a 5 m aperture which is acceptable. Extensive measurements have been made on the microwave range of the field distribution at L, S and X bands [1]. Excellent field uniformity is achieved.

2.2 Evaluation of test field quality

Extensive evaluation of the co- and cross-polarized fields in the test zone at the receive positioner were conducted using an aperture field probe (Scientific Atlanta model 59511) which is 5.5 m (18 ft) long giving ± 2.75 m (± 9 ft) of travel from the centre. The original equipment came from the USA so the distances of travel were read out in feet and inches (1.0 ft = 305 mm, 1.0 inch = 25.4 mm and 12 inches = 1.0 ft). The field probe is a rigid aluminium H-beam structure which mounts on the upper azimuth turntable of the az/el/az positioner on the receive tower. This is shown in Figure 2.4 where the probe is set to make a scan in the horizontal plane. The elevation axis of the positioner is tilted forward to about 91.3° so that the axis of rotation of the upper azimuth turntable points at the base of the source tower 500 m away. There is a polarization positioner mounted on the movable base of the field probe. A signal pick-up horn with a square (610 mm x 610 mm) of microwave radar absorbing material (RAM) directly behind it is mounted on the polarization positioner. The RAM is there to reduce reflections from the H-beam particularly when the incident polarization is parallel to the H-beam. The polarization positioner can be remotely controlled and gives a convenient means to set the probe horn polarization for V or H or even to rotate the probe polarization continuously.

The aperture field probes were made in L (1.6 GHz), S (2.9 to 3.9 GHz) and X (9.6 GHz) bands for V and H polarizations in the horizontal and vertical planes [1]. Co- and cross- polarized fields were measured at 2.9 and 9.6 GHz. Figure 2.5 shows a set of field probes at 9.6 GHz. Figure 2.5 (a) and (b) show that excellent test field symmetry is obtained in the vertical plane for both V and H polarizations for a source height of 0.35 m. The expanded plots directly below Figure 2.5 (a) and (b) are on a 0.5 dB scale. The field in the test zone is smooth to less than 0.1 dB. The black dots show the predicted field distribution using Equation 2.4 – the agreement is remarkable! Figures 2.5 (c) and (d) show the test field in the horizontal plane; again the field is extremely flat and uniform. A test field taper of 0.5 dB is achieved over an aperture of 5.5 m x 5.5 m. It is this quality of test field which gave rise to the comments in [2].

Cross-polarization measurements were made to assess the cross polarization in the vertical plane at 9.6 GHz. Figure 2.6 shows a vertical field probe with the reference trace at 0 dB being a trace for transmit H polarization and receive H polarization (H - H). The trace below - 40 dB is for transmit V and receive H (V - H) and the trace at - 35 dB is for H - V. Cross polarization of 35 dB is excellent for a ground reflection range. At 2.9 GHz similar results to those in Figure 2.6 were achieved. In addition a set of polarization patterns were measured at 2.9 GHz by setting the source for V polarization and then rotating the probe horn through 360°. The source horn polarization was then rotated in fixed steps of 22.5° and sets of polarization patterns taken as shown in Figure 2.7. The cross polarization is better than 35 dB for all source angles. This is an important result for axial ratio measurements of radar and other reflector antennas with circular polarization.

The field probe evaluations of the quality of the field in the test zone at L, S and X bands show that the excellent test fields are achieved over the 5.5 m x 5.5 m aperture in the test zone. The measured fields are in excellent agreement with those expected from the discussion in Section 2.1. It is clear that the ground-reflection principle works extremely well and that the results in [1] show that the Microwave Range will function well at other microwave frequencies provided the range is properly configured.

2.3 Standard configuration for antenna measurements

Section 2.1 has presented the requirements for the height of the source antenna to obtain the most uniform field in the test zone. Aperture field probe measurements as described in Section 2.2 and [1] have confirmed that excellent field quality is maintained over a 5.5 m x 5.5 m test zone. The discussion below is for measurements in the 1.7 to 2.5 GHz frequency band but the test set-up is applicable to other frequencies when the appropriate test equipment is used.

A set of measurements was made to evaluate the dynamic range when measuring the Scientific Atlanta SA 12 – 1.7 standard gain horn and using the Scientific Atlanta SA 22 – 8A parabolic reflector as the source antenna. In addition to the dynamic range measurements, a set of E- and H-plane patterns was also measured. These patterns are not presented here but are available at the NATR. Figure 2.8 shows the test range lay-out. The SA 12 – 1.7 horn (AUT) is mounted on the test positioner by means of a short fibre glass mast. The source antenna is the SA 22 – 8A (2.4 m diameter) reflector antenna with a 0.5 to 3 GHz prime-focus LPDA feed. The RF synthesized sweeper is at the source tower and the microwave receiver in the operations room requires a signal pick-up antenna at the receive tower to obtain a reference signal to phase lock the microwave receiver.

The geometrical parameters for this configuration are:

h_r = 12.2 m

R = 500 m

Because the horn antenna (AUT) is small, a centre frequency of 2.1 GHz was selected to obtain the ideal source height from Equation 2.1. This gives

h_t = 1.46 m

The depression angle $\alpha = 1.4^\circ$.

All the required interconnections are shown in Table 2.1. The microwave receiver operates in a similar way to a vector network analyzer except that the S-parameter test box is replaced by a frequency converter. The user selects how the connections to the frequency converter are made. In this case port a1 was selected as the reference port for phase lock and port b2 was selected as the test port. The microwave receiver is then set to measure b2/a1.

NOTE: The HP 8511A frequency converter is extremely sensitive to damage by electrostatic discharge (ESD). It is recommended that 3 dB fixed attenuators be placed at the RF input ports to the HP 8511A. These attenuators have a DC path to ground and assist in the continuous discharge of static build-UP on the test cables.

Table 2.1 Interconnections for microwave antenna measurements

Item number	Description
RF Sweeper	HP 83640 A, 10 MHz – 40 GHz synthesizer
Tx1	Transmit coaxial cable from sweeper
RF amplifier	Mini Circuits ZHL – 4240, 0.7 – 4.2 GHz
DC Power supply	Agilent N5766 A
Control	Fibre optic link from receiver
Tx2	Transmit coaxial cable to source dish
Source dish	Scientific Atlanta SA 22 – 8A reflector
Phase reference	Phase reference antenna, 0.5 – 3.0 GHz LPDA
REF1	Coaxial cable to port a1
AUT	Scientific Atlanta SA 12 – 1.7
Rx1	Receive cable from AUT to rotary joint
RJ	DC – 40 GHz rotary joint in upper azimuth turntable
Rx2	Cable from rotary joint to positioner
Rx3	Cable to step attenuator
Step attenuator	110 dB step attenuator
Rx4	Cable to test port b2
Receiver	HP 8530A microwave receiver
Frequency converter	HP 8511A frequency converter
Control	Fibre optic control link from receiver to source

Notes:

1. The HP 8511A frequency converter covers 45 MHz – 26.5 GHz.
2. Measurements can be done over the frequency range of the HP 8511A provided appropriate RF cables, amplifiers and source and reference antennas are used. This equipment is available at the NATR.

Table 2.2 shows the nominal set-up parameters for the measurements.

Table 2.2 Set-up parameters for test equipment

Microwave receiver HP 8530A	Measure Averaging Smoothing	: b2/a1 : 128 : 0%
RF Sweeper HP 83640A	Output power	: - 5 dBm (max. into ZHL-4240)

The dynamic range was evaluated for vertical polarization. Previous measurements have shown that the dynamic ranges for V and H polarizations are the same. The source antenna is shown in Figure 2.9. The vertically polarized LPDA feed is visible near the front of the tripod which supports the feed. The height of the source carriage is set to place the centre of the dish 1.46 m off the ground. The axis of the source dish is aligned to be parallel to the range surface.

The synthesized sweeper, DC power supply, fibre optic control link and Mini Circuits amplifier are shown in Figure 2.10. The Mini Circuits ZHL – 4240 amplifier is shown in Figure 2.11. The RF input is on the left and the RF output is on the right.

The SA 12 – 1.7 horn (AUT) and the phase lock antenna (LPDA 0.5 to 3 GHz) are shown in Figure 2.12. The 110 dB step attenuator, frequency converter and microwave receiver are shown in Figure 2.13.

The b2/a1 response was calibrated as a 'THRU' response with the AUT pointing at the source antenna at 500 m. This results in a 0 dB level on the display screen of the HP 8530A receiver. Attenuation was applied in 10 dB steps to -70 dB. The trace in Figure 2.14 was obtained by applying the 10 dB steps successively as the frequency is swept from low to high. The small 'glitches' in the trace occur as the mechanical attenuation knob on the attenuator is rotated. At the – 60 dB level there is about ± 0.5 dB noise on the trace. Figure 2.15 shows the noise at the – 70 dB level. A dynamic range of 70 dB ± 2 dB is achieved. The AUT has a nominal gain of 15.5 dBi. Typical radar antennas have gains which are 15 to 20 dB more than this value. In S-band the dynamic range is excellent. As the frequency increases to X-band the dynamic range degrades because of increased path loss and cable loss. However, a dynamic range of 40 dB can be achieved in a standard set-up to 18 GHz.

IMPORTANT NOTE FOR CIRCULAR POLARIZATION MEASUREMENTS: For linearly polarized measurements a linearly polarized phase reference antenna can be used. If only V and H source polarizations are to be used the phase reference antenna (horn or LPDA) can be set at 45°. However, if axial ratio measurements are to be made with a spinning linear source antenna the phase reference antenna must be circularly polarized. For rotating linear source, a linear phase reference antenna will be cross polarized at some angles and the phase lock may be lost.

There is a 305 mm diameter reflector antenna with a circularly polarized Archimedes spiral antenna feed which covers 1.5 to 18 GHz available at the Microwave Range. This provides a good phase reference signal. Helical antennas and conical spirals can be used for frequencies lower than this.

If the reference signal a1 fluctuates as a function of the source polarization angle, it may influence the ratio b2/a1. Under these conditions it may be beter to measure b1 as the test signal for the AUT. The operator must be aware that circularly polarized measurements require careful evaluation of the particular test set up.

Figure 2.1 Plan view of the microwave antenna test range of the NATR

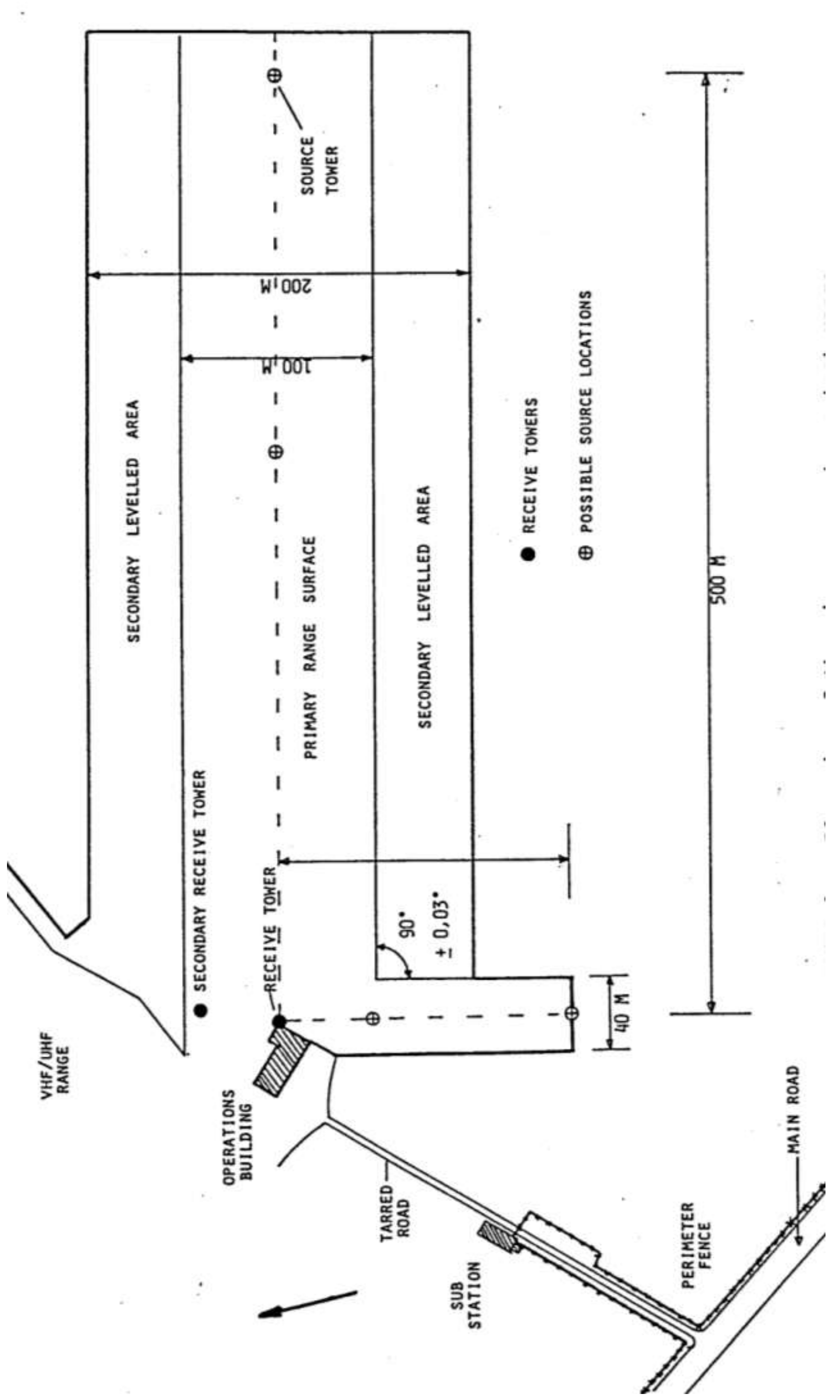


Figure 2.2 Geometrical ray tracing showing
(a) direct and reflected rays,
(b) formation of elevation test field by summation of direct and reflected waves

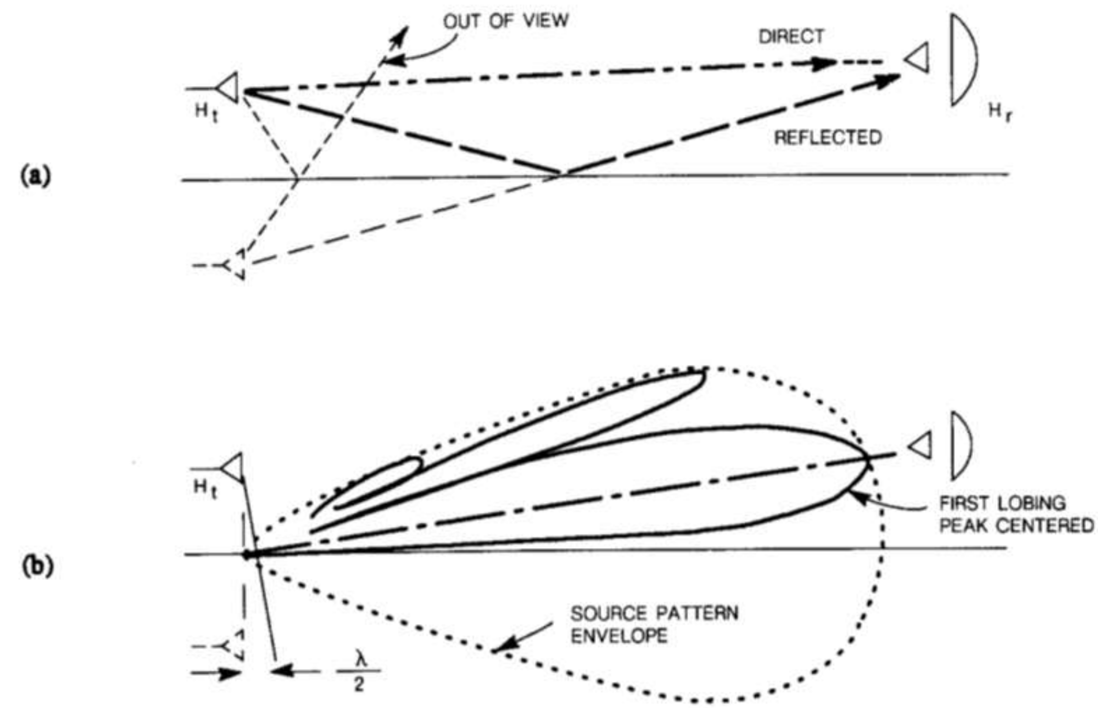


Figure 2.3 **Taper in ground reflection range test zone (a) cosine variation, (b) edge taper as a function of the ratio of test antenna size size and height at centre of test zone**

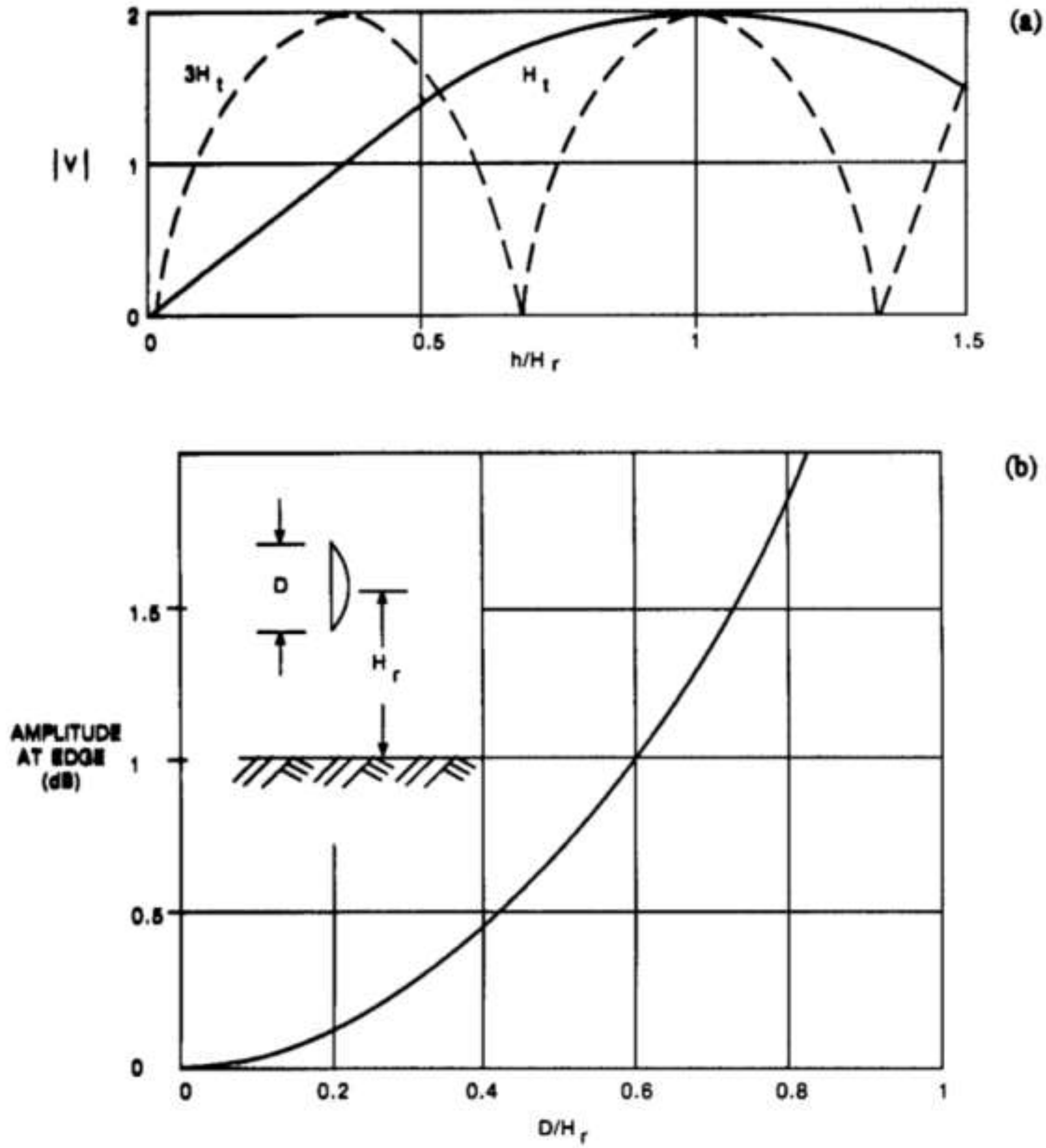


Figure 2.4 Aperture field probe mounted on the upper azimuth turntable tilted forward in elevation by 91.3° and set for a scan in the horizontal plane

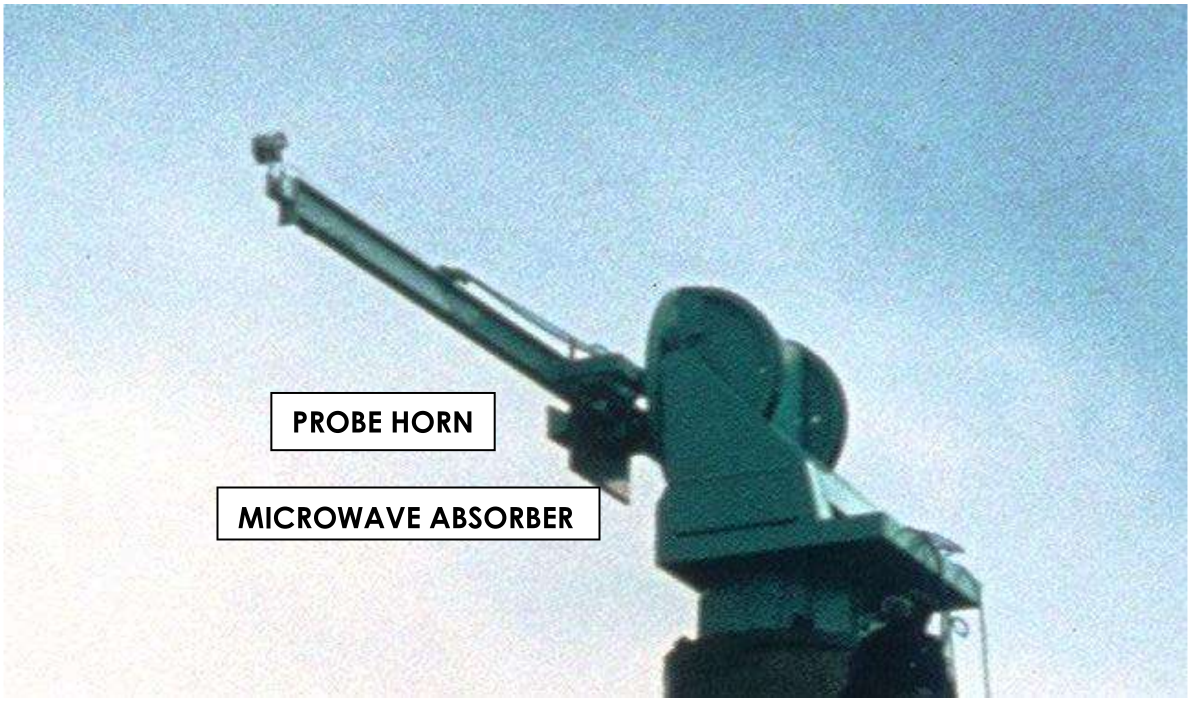


Figure 2.5 Aperture – field patterns at 9.6 GHz (a) Vertical plane for vertical polarization, (b) vertical plane for horizontal polarization, (c) horizontal plane for vertical polarization and (d) horizontal plane for horizontal polarization. The Y axis is in dB as shown and the X axis is in units of 1 foot (1 ft. = 305 mm). Measured (———), predicted (● ●).

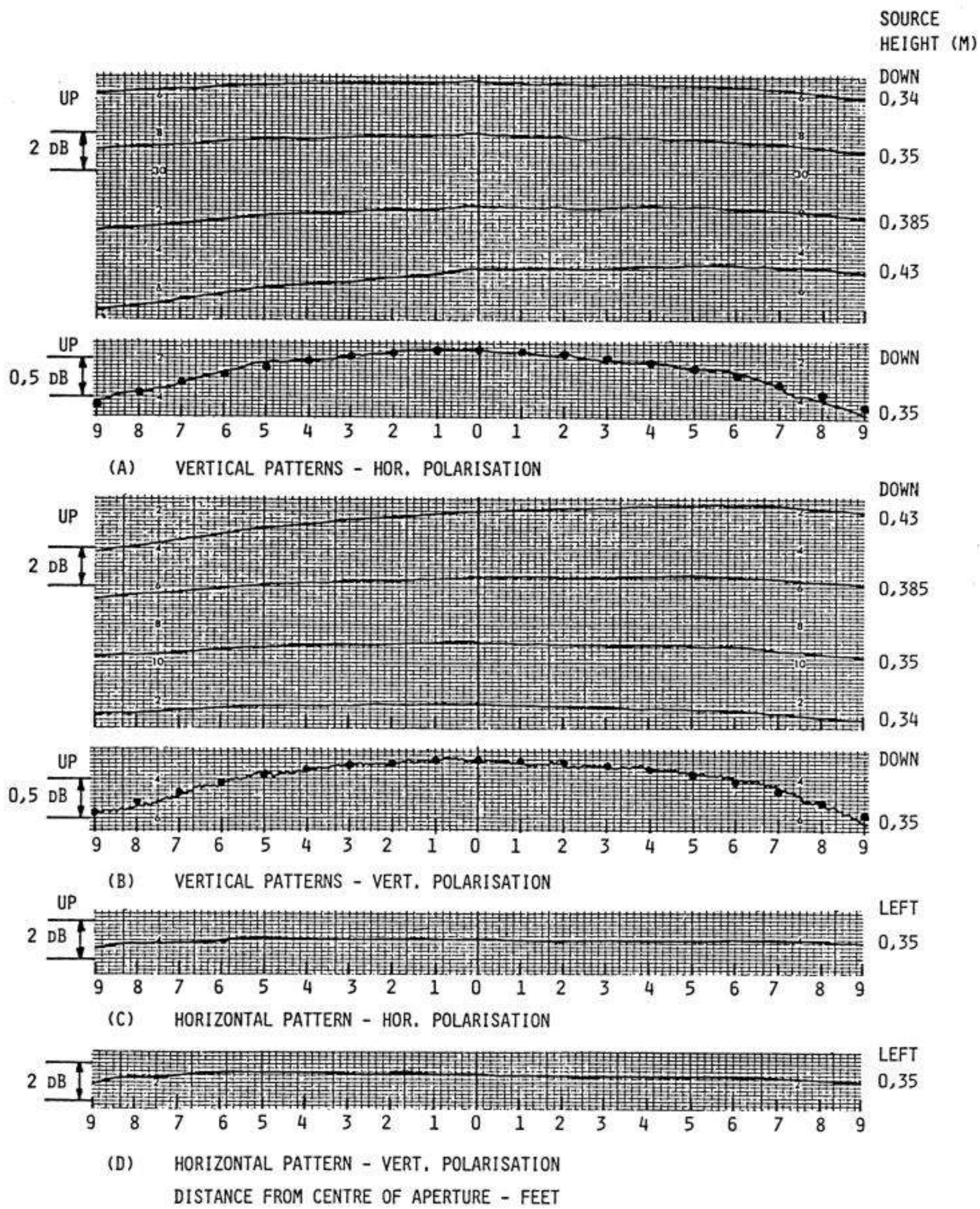


Figure 2.6 Measured cross polarization in the vertical plane over 5.5 m height at 9.6 GHz for source height of 0.35 m (1 foot = 305 mm)

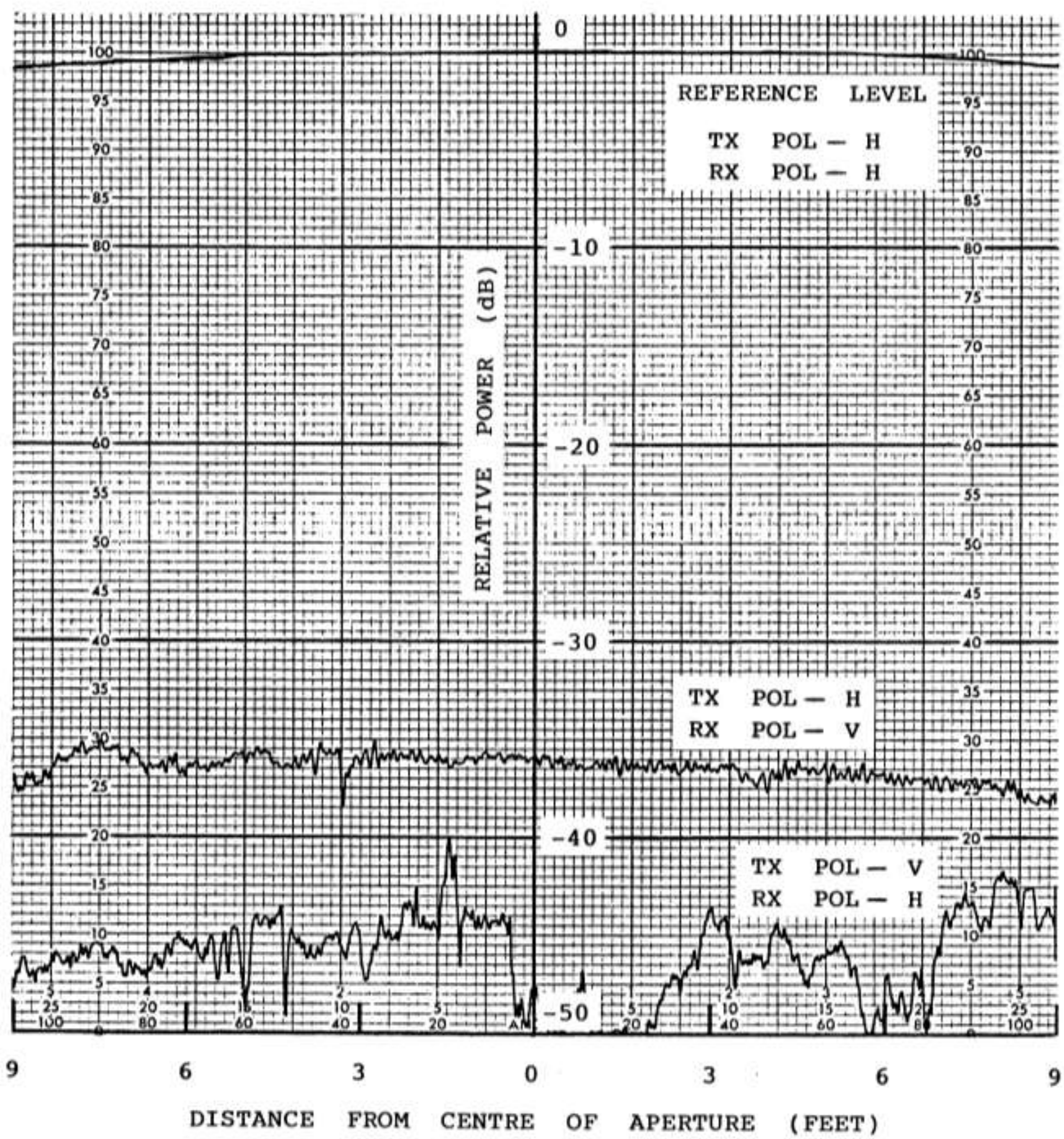


Figure 2.7 Family of polarization patterns as a function of source polarization angles at 22.5° angular spacing at 2.9 GHz for source height at 1.06 m

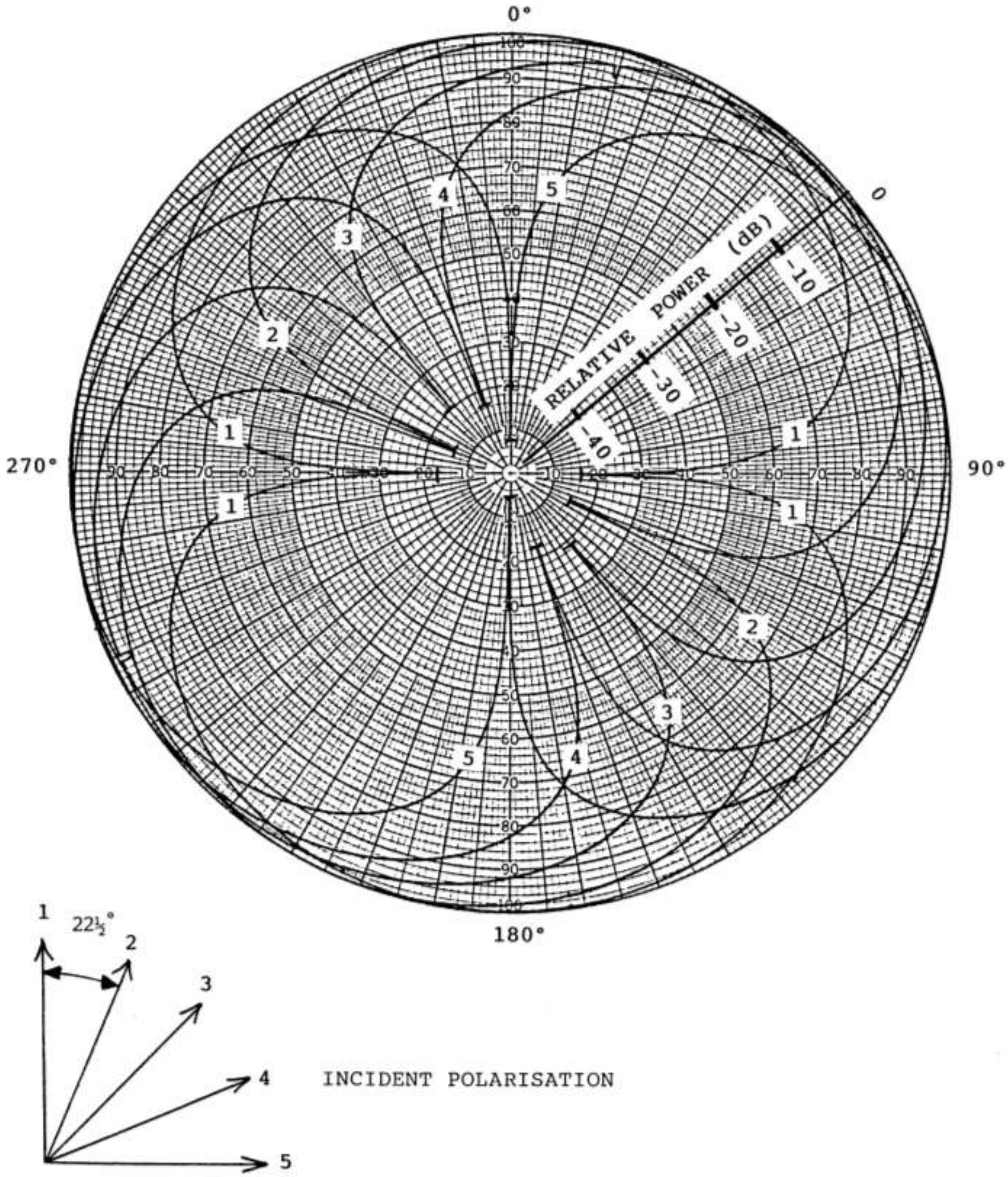


Figure 2.8 VHF/UHF test range layout with SA22-8A with LPDA feed set to optimise ground reflection at 2.1 GHz for SA12 – 1.7 horn as AUT

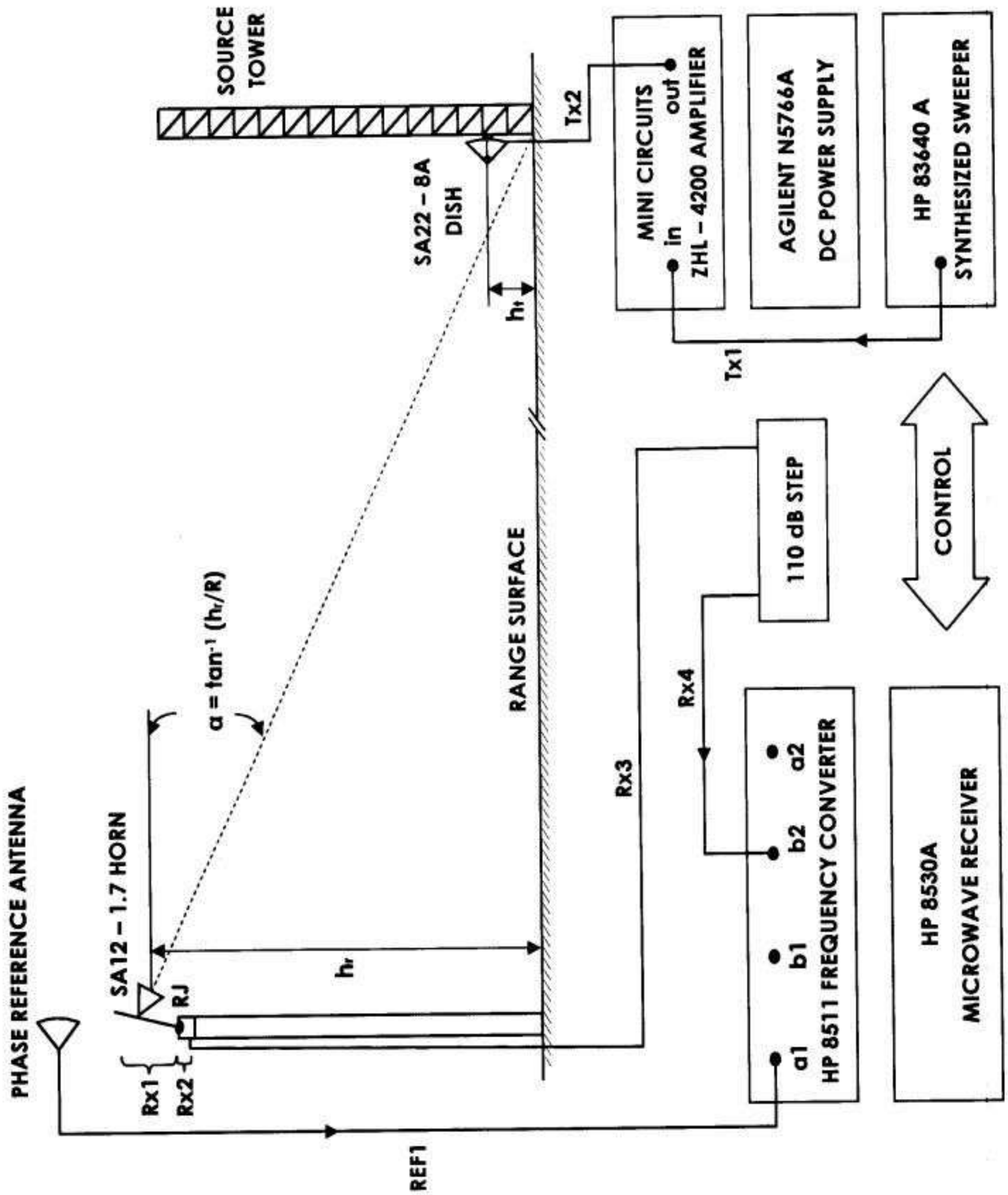


Figure 2.9 SA22 – 8A dish with 0.5 to 3 GHz LPDA feed set at 1.46 m

Figure 2.10 Synthesized sweeper with DC power supply on top, fibre optic control link at left and Mini Circuits amplifier at right

Figure 2.11 Mini Circuits ZHL – 4240 medium power amplifier (0.7 to 4 GHz, 40 dB gain, + 28 dBm output power)

Figure 2.12 SA12 – 1.7 receive horn (AUT) mounted on test positioner with the white 0.5 to 3GHz LPDA antenna at the base of the positioner

Figure 2.13 HP8530A microwave receiver, HP8511A frequency converter and 110 dB step attenuator. Reference signal is on the left and test signal on the right

Figure 2.14 Dynamic range for SA 12 – 1.7 horn with SA 22 – 8A source dish and Mini Circuits ZHL – 4240 amplifier

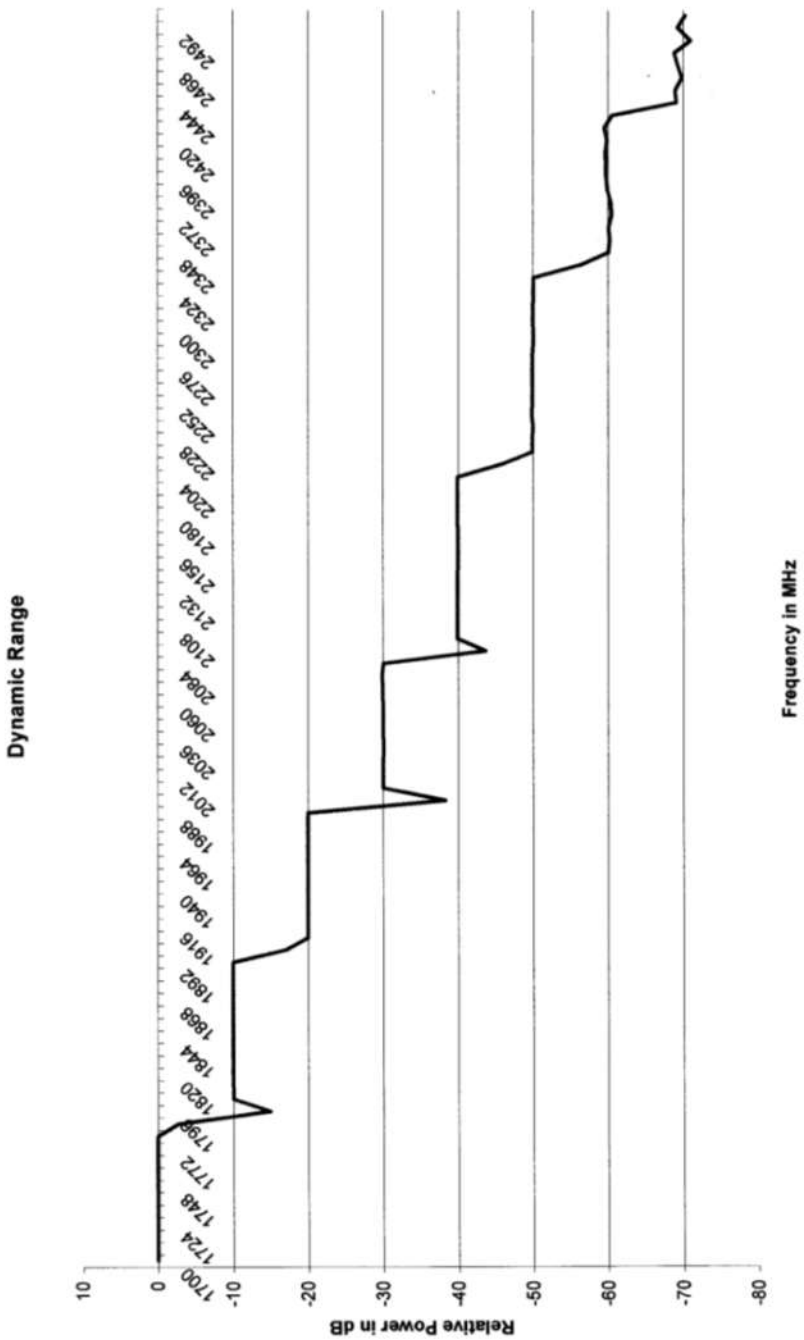
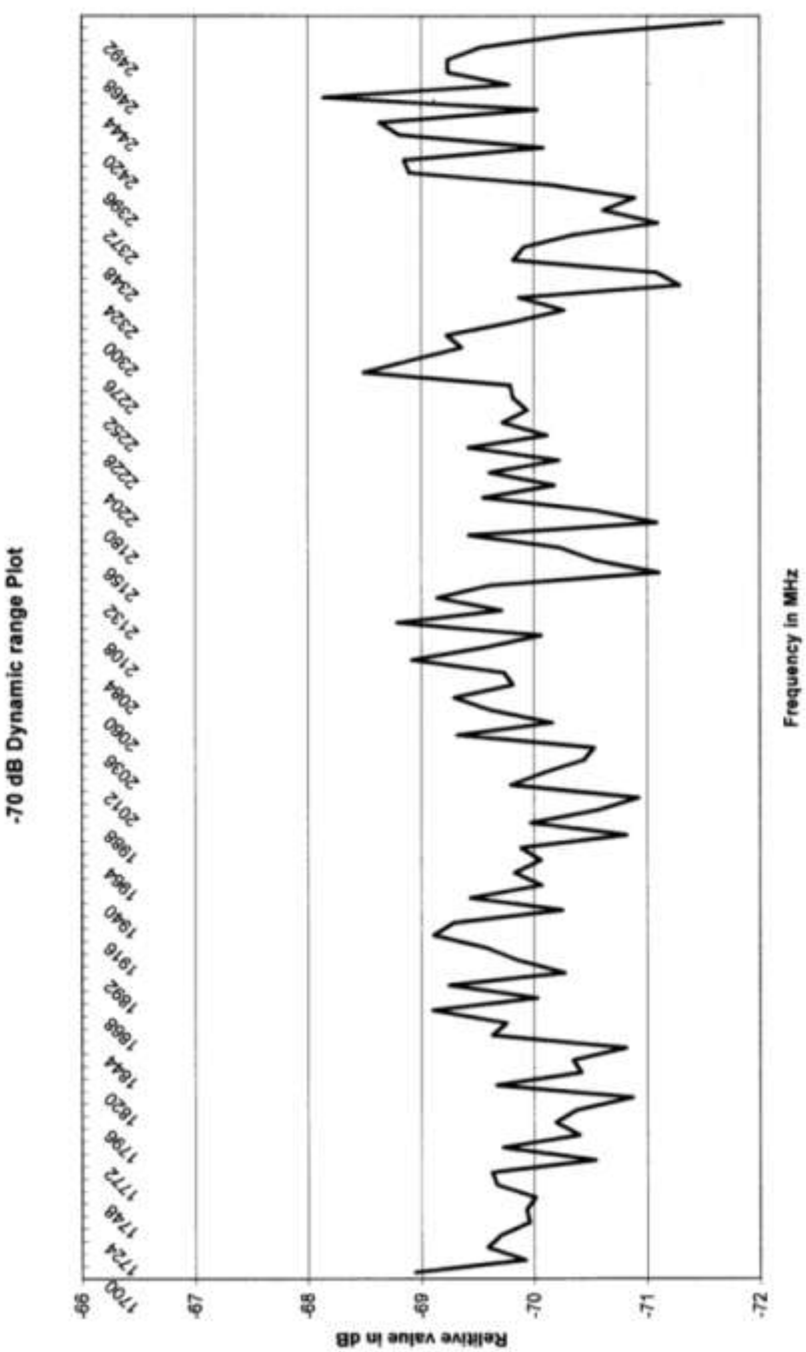


Figure 2.15 Amplitude stability (± 2 dB) with 70 dB attenuation for SA 12 – 1.7 horn with SA 22 – 8A source dish and Mini Circuits ZHL – 4240 amplifier



3. SIZE OF ANTENNAS AS A FUNCTION OF FREQUENCY

3.1 Positioner capability

The receive tower at the operations building is a freestanding reinforced concrete cylinder. The base of the cylinder is integrally attached to the reinforcing of a cube of reinforced concrete buried below the base of the tower. The cube of concrete acts as a counterpoise to prevent the tower toppling over in high wind load conditions (e.g. a large aperture antenna and a relatively high wind speed).

The receive tower has a Scientific Atlanta model 55240B azimuth/ elevation/ azimuth (az/el/az) positioner mounted at its top. The horizontal surface of the upper azimuth turntable is 11.2 m above ground level. The turntable has a mechanical specification of a vertical load of 10 000 lb (4 545 kg) and a bending moment of 10 000 lb. ft. (13 553 N m). For general radar antenna testing, these mechanical specifications are more than adequate. For larger antennas (5 m diameter or more) the mass, gravitational bending moment and the wind-induced bending moment must be calculated to see that the positioner mechanical specifications are not exceeded. The wind-induced bending moment increases with the square of the wind speed and the area of the reflector antenna.

Figure 3.1 is a useful graph for obtaining an estimate of the pressure for various wind speeds. Imperial units are on the left and bottom of the figure and SI units are at the top and on the right. To convert a wind speed in km/h to m/s divide km/h by 3.60. The inserted tables on the plot show that the pressure is a function of the temperature, the shape and the structure of the dish. A wire mesh construction has a lower wind load than a solid reflector. Even though some mesh constructions appear quite open to the observer, the individual wires contribute significant drag. The net force is obtained by multiplying the pressure by the projected surface area of the reflector. If a large reflector antenna is to be measured, a wind loading study should be done. The NATR has acquired a weather station to record wind speeds, both gusts and average speeds. These speeds can be used in the wind loading study.

3.2 Range distance requirements

3.2.1 Phase taper

Antennas are normally tested at some finite range where the incident field approximates a plane wave [5]. The commonly employed far-field criterion gives the range length as

$$R \geq \frac{2D^2}{\lambda} \tag{3.1}$$

where

R = test range = 500 m

D = largest dimension of AUT (m)

λ = wavelength (m).

The microwave range was designed to meet equation 3.1 at 3 GHz for a 5 m aperture. This was envisaged as the largest aperture to be measured and is typical of S-band air traffic control radars.

3.2.2 *Inductive coupling*

In addition to the far-field criterion we must also limit the inductive coupling. Here the commonly accepted criterion is that the test range should be greater than 10 wavelengths [3]. Inductive coupling is more of a factor at the lower frequencies. The ratio of the amplitude of the inductive field E_i to the radiated field E_r from the source antenna at range R is given by:

$$E_i / E_r = \lambda / (2\pi R) \qquad (3.2)$$

If $R > 10\,\lambda$ then $20 \log (E_i / E_r) < -36$ dB.

At 500 m test range this means the largest wavelength is 50 m which is a frequency of 6 MHz. Thus at the normal microwave frequencies above 500 MHz, inductive coupling is not a factor. However, the microwave range is being used more and more for HF/VHF/UHF measurements. At 1 MHz the inductive coupling is about -20 dB.

3.2.3 *Longitudinal amplitude taper*

For reflector antennas the aperture (diameter) of the dish is the main parameter. However, there are end-fire antennas (Yagis, log periodic dipole arrays, end-fire arrays) which have significant length L. In this case the taper along the length of the antenna must be limited to about 1 dB. If the centre of the antenna is at R the front of the antenna will be L/2 closer to the source and the rear of the antenna L/2 further away. The amplitude taper A over the length of the antenna is:

$$A = 20 \log \frac{R + L/2}{R - L/2} \qquad (3.3)$$

To achieve the 1 dB limit we need $R \geq 10$ L. At 500 m range this gives $L \leq 50$ m which clearly is not a problem. At shorter ranges (e.g. anechoic chambers) this condition may be difficult to achieve.

3.2.4 *Antenna diameters and gain at 500 m range*

From the foregoing it is clear that the far-field phase criterion places the most stringent requirement for testing reflector antennas. We can deduce the largest linear dimension D of the reflector from Equation 3.1 as

$$D \leq \sqrt{\frac{1}{2} R \lambda} \tag{3.4}$$

These diameters can be used to calculate the typical gains and beam widths of antennas to be tested at 500 m range. The gain for a circular dish of diameter D is given by

$$G = \frac{4\pi}{\lambda^2} A_e \tag{3.5}$$

where A_e is the effective aperture (area) of the dish. This is related to the physical aperture A_p by

$$A_e = \in_{ap} A_p \tag{3.6}$$

for a well-designed reflector antenna $\in_{ap}$ = 0.65. Thus

$$G = 0.65 \left(\frac{\pi D}{\lambda} \right)^2 \tag{3.7}$$

The 3 dB (half power beamwidth) of a circular aperture can be approximated by

$$\theta_{hp} = 70° / (D / \lambda) \tag{3.8}$$

Equations 3.4, 3.7 and 3.8 are used to set up Table 3.1 which gives the maximum diameters, gain and beamwidth of reflector antennas which can be measured with small errors at a test distance of 500 m on the microwave range.

Table 3.1 Maximum reflector diameter (based on Equation 3.4) and associated gain and half power beamwidth at various frequencies for 500 m test range

Frequency (GHz)	Diameter D (m)	Gain (dBi)	3 dB BW (deg)
0.5	12.2	34.2	3.4
1	8.7	37.3	2.4
2	6.1	40.3	1.7
3	5.0	42.1	1.4
5	3.9	44.3	1.1
10	2.7	47.2	0.78
18	2.0	49.7	0.58
26	1.7	51.4	0.48
40	1.4	53.5	0.37

Notes:

1. The reflector diameters at 0.5 and 2 GHz will probably exceed the mass and torque specifications of the test positioner.
2. For large apertures in wavelengths (100 or more) the beamwidths become very narrow (above 10 GHz) and place stringent requirements on the positioner accuracy.
3. The table shows the largest antennas which can be measured at the far-field range of Equation 3.1.

3.3 Source heights and dish diameters

Section 2.1 has set the criteria for the source heights (Equation 2.1) as a function of frequency and maximum dish diameters (Equation 2.2) to achieve best overall field distribution in the test zone of diameter D. The source height h_t at each frequency sets the source dish diameter d to a maximum value of $2h_t$. At this diameter of the source dish its bottom rim will just touch the range surface at the source tower.

The source tower has an overall height of 12.8 m and it has an automated carriage with a free travel of 11 m. The carriage is equipped with a polarization positioner which can continuously rotate about its axis or be set for V or H polarizations. The polarization positioner has manually adjustable fixtures which allow for adjustment of the polarization positioner in azimuth and elevation. This arrangement gives complete freedom to align the source antenna properly.

The NATR has a selection of source dish antennas available to be mounted on the source tower. Their dishes typically have tripod mounts with adjustment mechanisms to position the dish feed antenna on the axis of the dish and at the focal point of the dish. To ensure best test field illumination at the receive tower, careful boresight alignment of the feed is required to ensure that the mechanical reference face and the electrical axis of the dish are at precisely right angles to each other. This will mean that as the polarization axis of the polarization positioner rotates, the main beam of the source dish keeps a fixed pointing direction and does not rotate on a cone. Such conical scan movements of the main beam from the dish will mean that the symmetry of the field in the test zone is a function of the source polarization. This is extremely undesirable.

The available source dishes and their feeds are given in Table 3.2. The dishes are listed in decreasing size using the Scientific Atlanta part number designation where the -10A in 22-10A means a 10 ft diameter dish.

Table 3.2 Source dish diameters and feeds

Reflector	Diameter m (ft)	Feed type	Frequency (GHz)
22-10A	3 (10)	Horn/ LPDA	Horns are standard waveguide bands from 1.7 to 40 GHz. LPDA feed is custom 0.5 - 3 GHz
22-8A	2.4 (8)	Horn	
22-6A	1.8 (6)	Horn	
22-4A	1.2 (4)	Horn	
22-2A	0.6 (2)	Horn	
22-1A	0.3 (1)	Horn	

Notes:

1. The NATR has a series of horn feeds from Scientific Atlanta which are compatible with the various reflectors. These feeds cover the standard waveguide bands 1.7 - 2.6 GHz, 2.60 - 3.95 GHz and so on up to 26 - 40 GHz.
2. There is a wideband 0.5 - 3 GHz log periodic dipole array (LPDA) feed which can be used on the 22-10A and 22-8A reflectors.

3. Not all feeds can be used on all of the dishes. Refer to the Scientific Atlanta catalogue for a cross reference table which gives the feed compatibility.

Table 3.3 presents the ideal heights for the source antennas at various frequencies using Equation 2.1 and also the maximum source diameter d_{max} based on a test zone diameter derived from the maximum AUT dish diameters D_{max} at the $R = 2D^2/\lambda$ range in Table 3.1. The diameter d_{max} is derived from Equation 2.2.

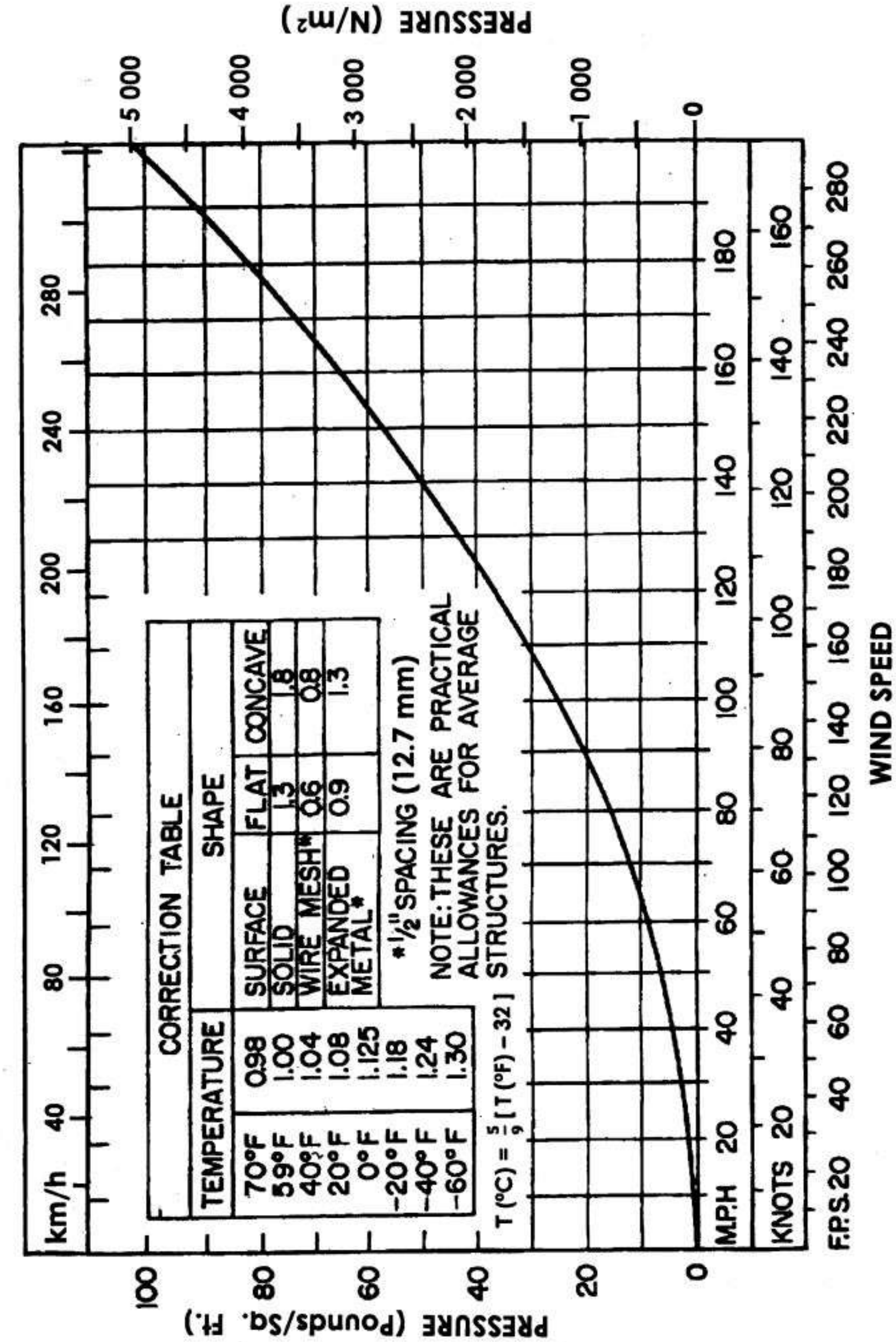
Table 3.3 Ideal transmit height and source dish diameters at 500 m test range

Frequency (GHz)	Ideal height (m)	D_{max} AUT diameter (m)	d_{max} source diameter (m)	Dish type
0.5	5.91	12.2	9.03	22-10A
1	2.95	8.7	6.44	22-10A
2	1.48	6.1	4.51	22-8A
3	0.98	5.0	3.70	22-6A
5	0.49	3.9	2.89	22-4A
10	0.30	2.7	2.00	22-4A/22-2A
18	0.16	2.0	1.48	22-1A
26	0.11	1.7	1.26	Horn
40	0.07	1.4	1.04	Horn

Notes:

1. The source heights decrease with increasing frequency.
2. Double the source height effectively sets the source dish diameter.
3. At 10 GHz the source height is 0.3 m which gives a dish diameter of 0.6 m. The bottom rim of the dish is then on the range surface. The 22-2A dish was used as the source antenna for the field probes.
4. Above 18 GHz custom horns must be used since the 22-1A dish is too large. The 22-1A will still be useable to 26 GHz since the bottom rim will be only about 40 mm below the range surface. A small depression can be made in the range surface below the dish to accommodate this.

Figure 3.1 **Wind-loading data for various wind speeds, reflector shapes and surfaces**



4. MEASUREMENT OF GAIN

4.1 Linearly polarized gain

The most commonly used procedure for gain measurement on the microwave range is the gain transfer method [3]. Here the gain of the antenna under test (AUT) is measured by comparing the received power on the AUT with that received on a reference antenna of known gain. This reference antenna is often called a "standard gain" antenna (SGA). At the microwave range these gain standards are normally horn antennas.

The SGA is first connected to the receiver via a stable coaxial cable and the SGA is peaked in azimuth and elevation. The polarization of the source must also be aligned with that of the SGA. The antenna under test (AUT) is then connected via the same test cable (See schematic in Figure 4.1(a)). If the same test cable cannot be used, the change in losses between the cables must be measured and taken into account. For large antennas under test, the gain reference antenna is often mounted on the back of the AUT. The AUT and the SG horn are then connected via a switch which can be controlled from the operations room. All interconnecting cables must have the same losses or the relative losses must be taken into account. Absorber is often used to reduce reflections off the AUT into the SGA (See Figure 4.1(b)).

The difference in received power (P_T) on the test antenna and that on the standard antenna (P_S) represents the difference in gain.

$$\left(G_T\right)_{dBi} = \left(G_S\right)_{dBi} + 10\log\left(\frac{P_T}{P_S}\right) \quad (4.1)$$

which can be written as

$$G(AUT, dBi) = G(Ref, dBi) + D(dB) \quad (4.2)$$

where D = dB difference between the reference and the AUT. If the reference lies below the AUT, the gain of the AUT is higher than the reference antenna.

The gain of the reference antenna is normally provided by the supplier as a table or graph of gain versus frequency. The gain values are quoted to a typical accuracy of ± 0.25 dB.

Ideally, both the AUT and the reference antenna must be well matched. If not, there will be VSWR mismatch errors which should be taken into account [3]. The reference antennas usually have good VSWR and the AUT is measured without VSWR correction since this gives the operational or realised gain in the 50 ohm coaxial system. For many applications a 6 dB or a 10 dB attenuator is placed on the test cable directly behind the AUT or SGA output connector. This improves the match to the receiver and reduces small amplitude ripples in the gain measurement caused by VSWR interactions in the AUT/SGA - test cable - receiver chain.

4.2 Circularly polarized gain

Gain measurements for circularly or elliptically polarized antennas are complicated because the gain reference antennas are usually linearly polarized. That is, no simple equivalent procedure to the gain transfer method described in Paragraph 4.1 is available because of the lack of commercially available circularly polarized reference antennas. For custom applications (normally narrow band, less than 30% bandwidth) some users manufacture and calibrate their own standard gain circularly polarized antennas. This is impractical on a general purpose test range.

Circularly polarized (CP) gain is usually measured using linearly polarized gain reference antennas. These are the same reference antennas used for the linear gain. The use of linearly polarized reference antennas is valid because the total power in the elliptically polarized wave can be separated into two orthogonal linear polarizations.

The gain calibration is done as before for vertical polarization between the source antenna and the SGA. The elliptically polarized AUT is then installed and the "partial gain" for V polarization measured. The source antenna is then rotated to horizontal polarization and the partial gain for H polarization is measured. V and H gains are calculated using Equation 4.2.

The circularly polarized gain is then computed from the partial gains for the orthogonal linear polarizations (normally V and H polarizations) as follows [3]:

$$G(dBci) = 10 \log (g_v + g_h) \qquad (4.3)$$

where

g_v = V polarized gain in numbers (not dBi)

g_h = H polarized gain in numbers (not dBi)

The linear gains are normally given as dBi relative to a linearly polarized isotropic antenna (often written as dBli). The circularly polarized gain is then expressed as dBci, dB relative to a circularly polarized isotropic antenna. If the partial gains for the V and H polarizations have nearly the same dBi values (say within 2 dB), then CP gain can be approximated by adding 3 dB to the dB average of the V and H linear gains.

Figure 4.1(a) Schematic showing successive connection of AUT and gain reference to receiver via a switch

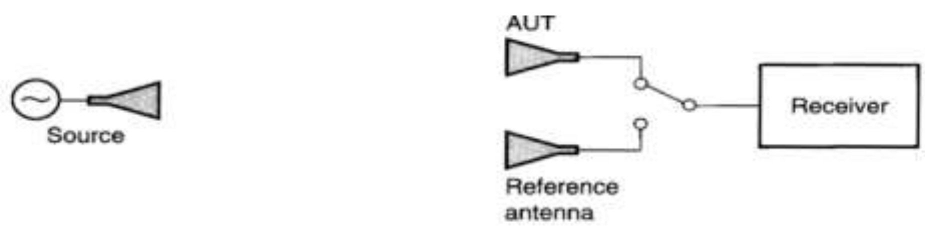
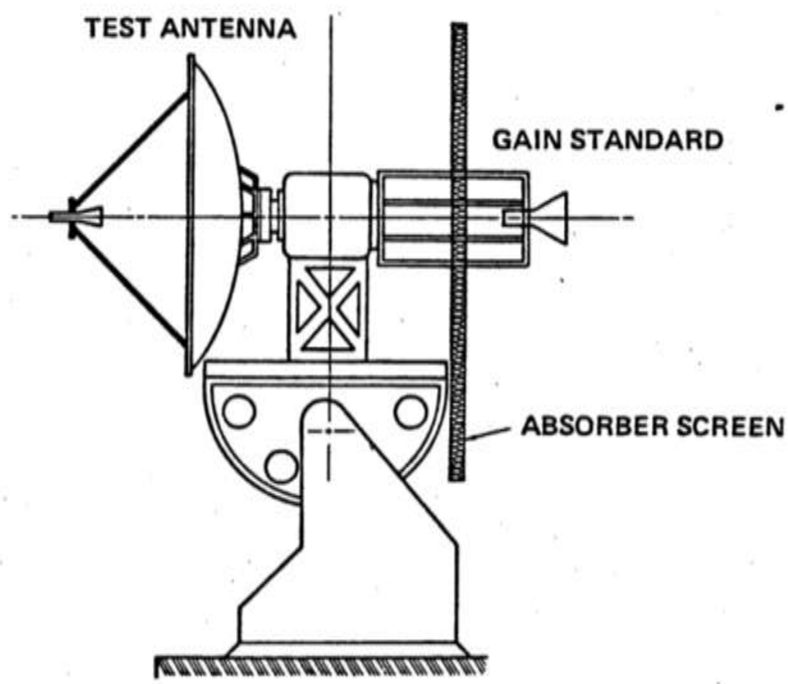


Figure 4.1(b) Test set-up showing AUT and gain reference mounted back-to-back (note absorber to eliminate reflections into gain standard)



5. POLARIZATION

Many antennas are linearly polarized (V, H or slant 45°). Occasionally radar antennas have a special circularly polarized mode which is switched in to reduce rain clutter and improve the signal to clutter ratio. This section discusses the capability of the microwave range to measure various patterns for linear and circular polarizations.

5.1 Linear polarization

The ground reflection range can make excellent linear cross polarization measurements provided sufficient care is taken. Extensive evaluation of the cross polarization characteristics of the microwave range was done in S - and X - bands [1]. A 1.83 m (6 ft) reflector with a 2.6 to 3.95 GHz horn feed at its focus was used as the source antenna for S - band. The axis of the source antenna was aligned to be parallel to the range surface. This maintains symmetry between the source main beam and its image in the range surface as the source antenna is rotated about its axis. At other orientations of the source antenna axis the direct beam and its image beam squint relative to each other and degrade the symmetry with rotation.

A standard gain horn antenna was used on a field probe (a linear H-beam which moves the horn antenna across the test aperture) to measure the co- and cross - polarized field over the full aperture of the field probe (5.5m). The field probe is shown schematically in Figure 5.1.

Over the full aperture of 5.5 m (18 ft) cross polarization of better than 39 dB was measured in S-band. This is at the limit of the cross polarization of the horn which was used as the polarization probe in the test area. The SGAs are typically specified at 35 to 40 dB cross polarization.

At X-band (9.6 GHz) the cross polarization was measured at better than 35 dB for H polarization transmit and for V polarization receive. For V polarization transmit and H polarization receive the cross polarization was better than 40 dB. There appears to be a slight depolarization from the range surface. More careful alignment of the source antenna should improve this to 40 dB.

The foregoing discussion illustrates that excellent co- and - cross polarization measurements can be made for linear polarizations on the microwave range. This is a pre-requisite for making circularly polarized measurements.

5.2 Circular polarization and axial ratio

Most radar antennas are linearly polarized. Linear polarization means that the tip of the E-field vector moves in a plane containing the E-field and the direction of propagation. There are some radars which have a circular polarization mode to reduce backscatter from rain. If a right hand circular polarization (RHCP) radar antenna illuminates rain the reflected wave is predominantly left hand circular polarization (LHCP). A complex target such as an aircraft may have multiple reflections which will return a RHCP signal for an even number of reflections. This improves the signal-to-rain-clutter ratio. For a circularly polarized wave the tip of the electric field vector moves on a circle. This is illustrated in Figure 5.2 which shows linear, circular and elliptical polarizations. The special case of circular polarization requires two linear E-field components at right angles to each other and a 90° phase difference between them.

If the two fields at right angles do not have equal amplitudes and 90° phase shift we get elliptical polarization as shown in Figure 5.3. The dB ratio of the major to minor axes of the polarization ellipse is the axial ratio given by

$$AR = 20\log \frac{OA}{OB} \tag{5.1}$$

An axial ratio of 0 dB represents a perfectly CP wave, while 20 dB or more gives a nearly linearly polarized wave. Circularly polarized radars often have a requirement for the AR to be less than 1 dB. Figure 5.3 shows how good the amplitude tracking and the 90° phase shift must be to achieve various axial ratios.

Because the microwave range has excellent linearly polarized characteristics it can make very good axial ratio pattern measurements. One way to make such pattern measurements is to spin the linearly polarized source rapidly about its axis while the AUT rotates slowly in azimuth or elevation. This results in a series of amplitude ripples in the patterns. The peak-to-peak ripples in dB give the AR. Figure 5.4 illustrates this procedure.

5.3 Integrated cancellation ratio

A figure of merit for circularly polarized radar antennas to assess the reduction of rain clutter is the integrated cancellation ratio (ICR). When circular polarization is used as a means of rejecting the reflection from rain in the radar response, the circular polarization over a wide volume of space is of interest. The ICR places emphasis on the necessity for having the radar beam circularly polarized all over and not just at its peak. Pattern levels below 10 dB down from the peak have very small impact on the ICR. It can be shown [4] and [6] that the ICR is

$$ICR = 10\log \frac{\sum_{\phi} \left(P_{max} - P_{min}\right)^2 \sin\theta}{\sum_{\phi} \left(P_{max} + P_{min}\right)^2 \sin\theta} \tag{5.2}$$

where P_{max} is the top envelope of the ripples, P_{min} is the bottom and θ is the elevation angle at the azimuth cut being measured (θ is 0° straight up and 90° on the horizon). Reference 6 gives a detailed table for computing the ICR. Figure 5.5 shows a typical azimuth cut of a radar antenna taken with a spinning linear source [4]. The cancellation ratio can be deduced at a particular angle from the axial ratio at that angle using the curve of Figure 5.6.

ICR is typically specified over a particular angular volume but for measurement convenience it applies to each pattern cut. From Figure 5.6 it is clear that the lower the axial ratio the better the cancellation ratio. An axial ratio of 1 dB gives a cancellation ratio of about 20 dB. It is critically important that the test range is capable of measuring very small axial ratios (less than 0.5dB). The microwave range at the NATR can be used to make these measurements but careful setting up of the range is essential. ICR is a specialised measurement and is not often made at the NATR anymore but this measurement is important for air traffic control radars.

Figure 5.1 Schematic diagram of field probe which has automated movement of horn antenna used as the field probe

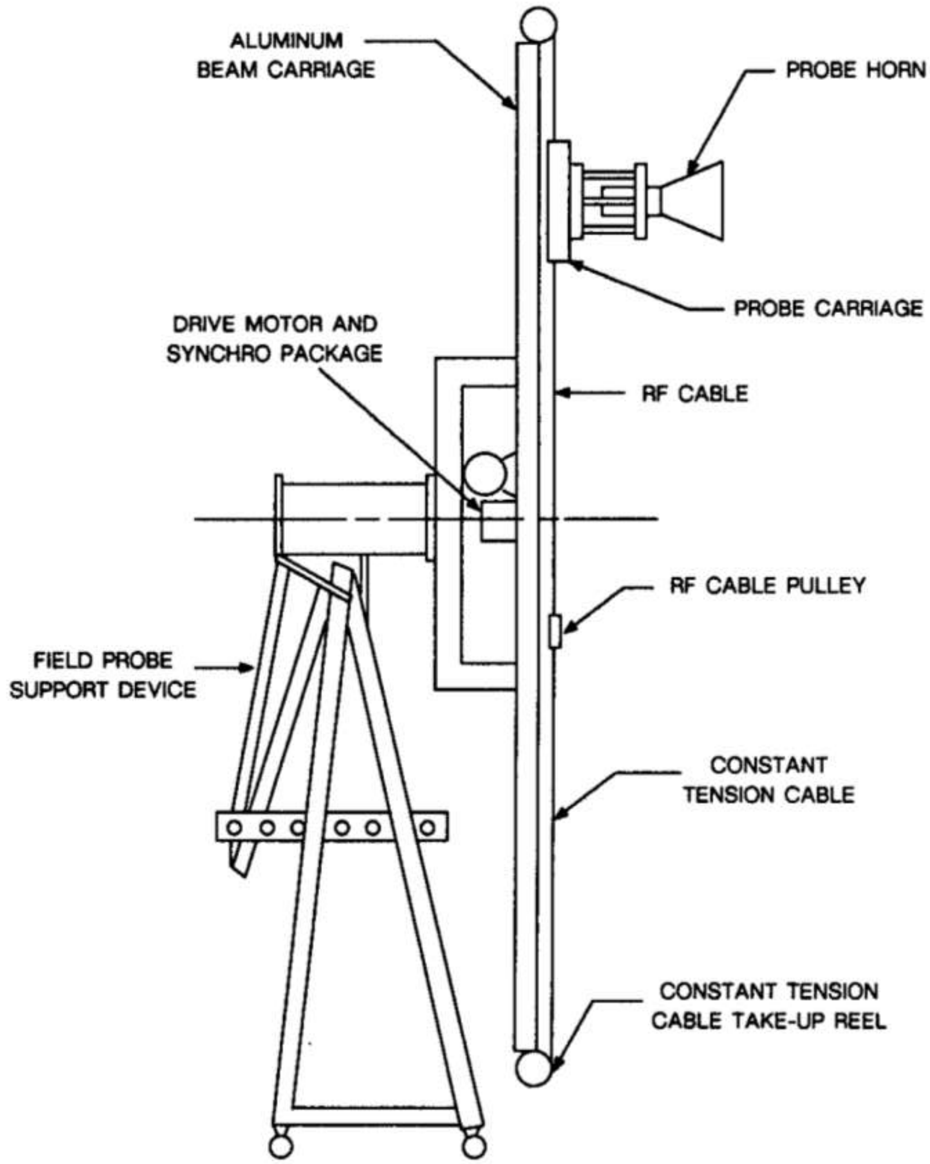
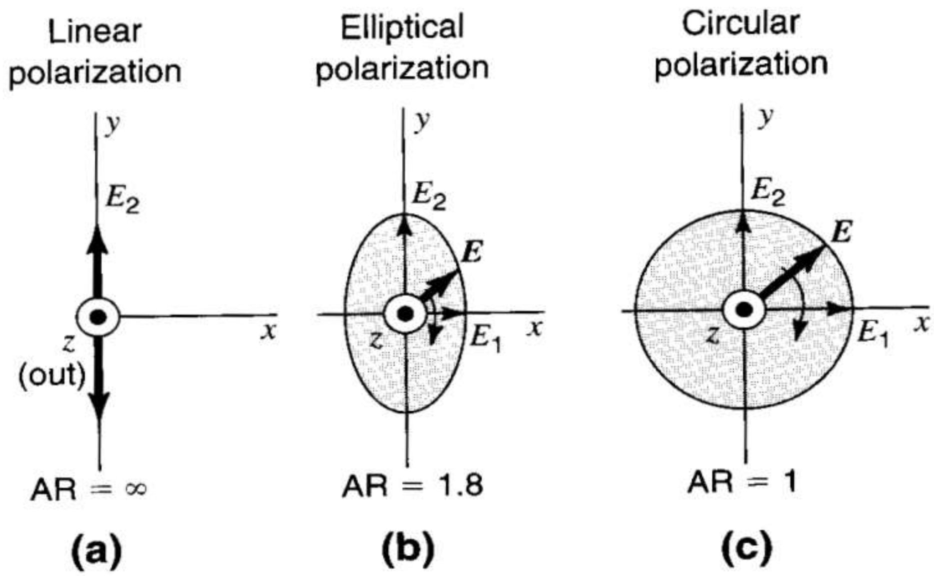


Figure 5.2 Schematic diagrams for linear, circular and elliptical polarizations



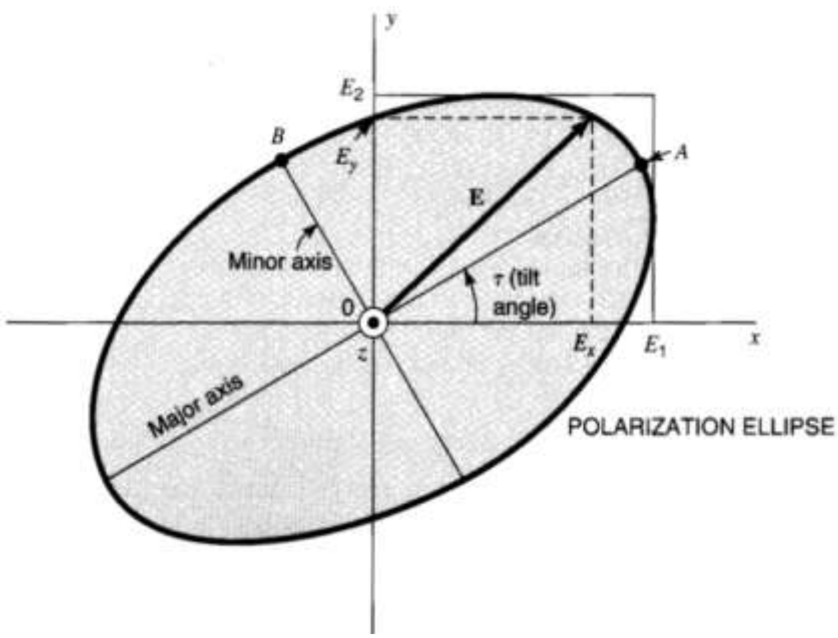


Figure 5.3 Amplitude and phase errors to achieve a specified axial ratio

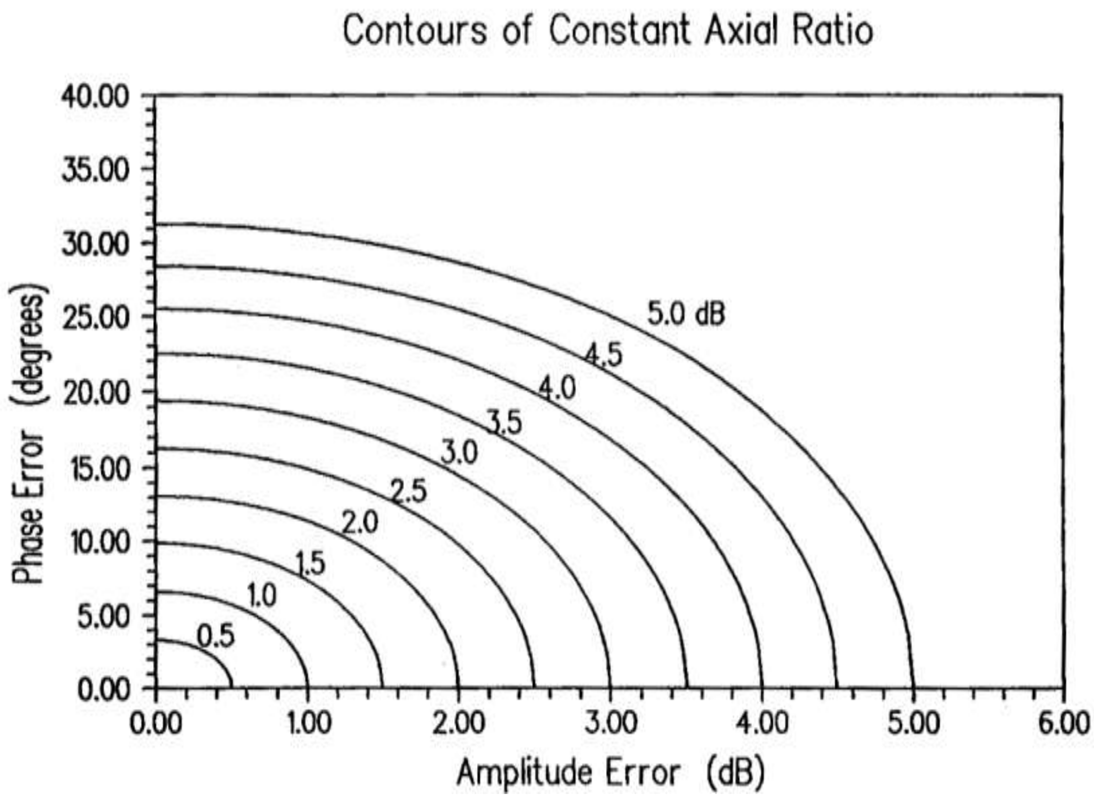
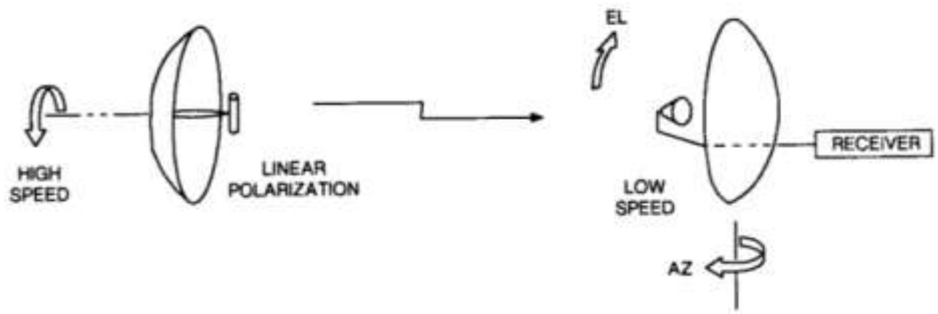
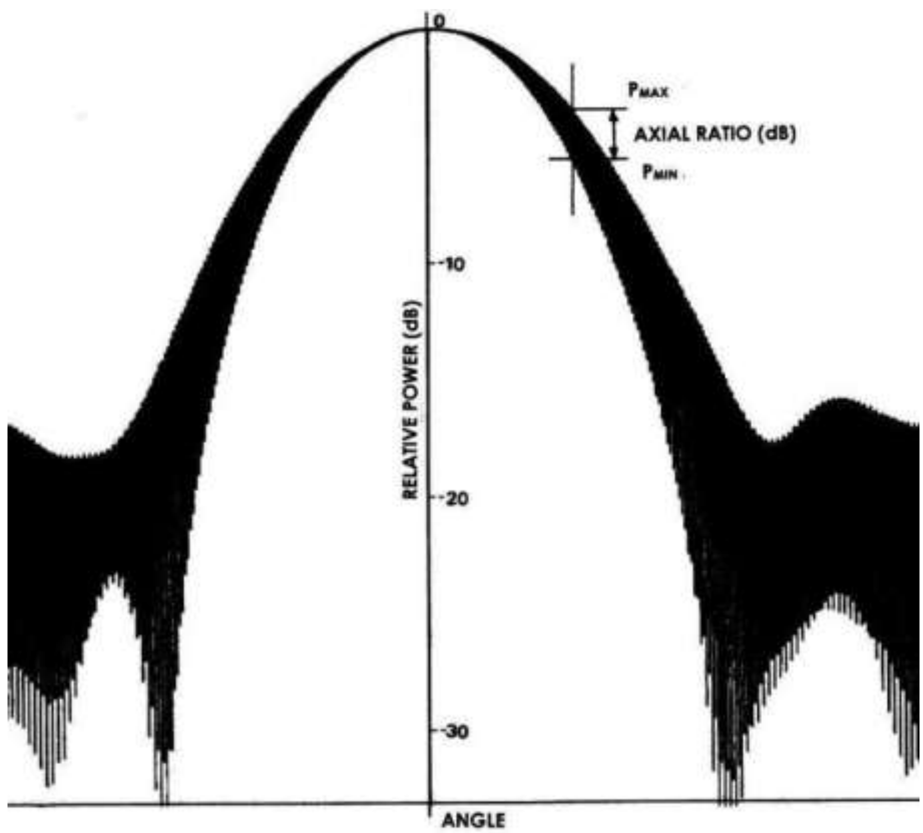


Figure 5.4 (a) Procedure for measuring CP with a linearly polarized spinning dish,
(b) typical response ($P_{max} - P_{min}$ in dB = AR at specified angle)



(a)



(b)

Figure 5.5 Typical axial ratio pattern using spinning linear source used to compute ICR

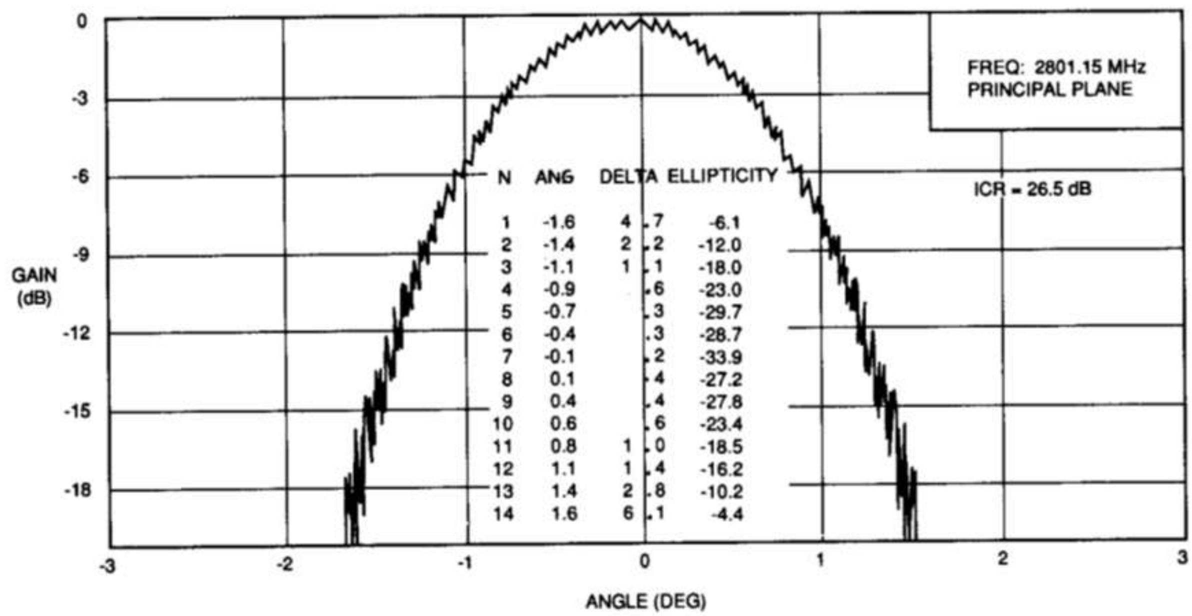
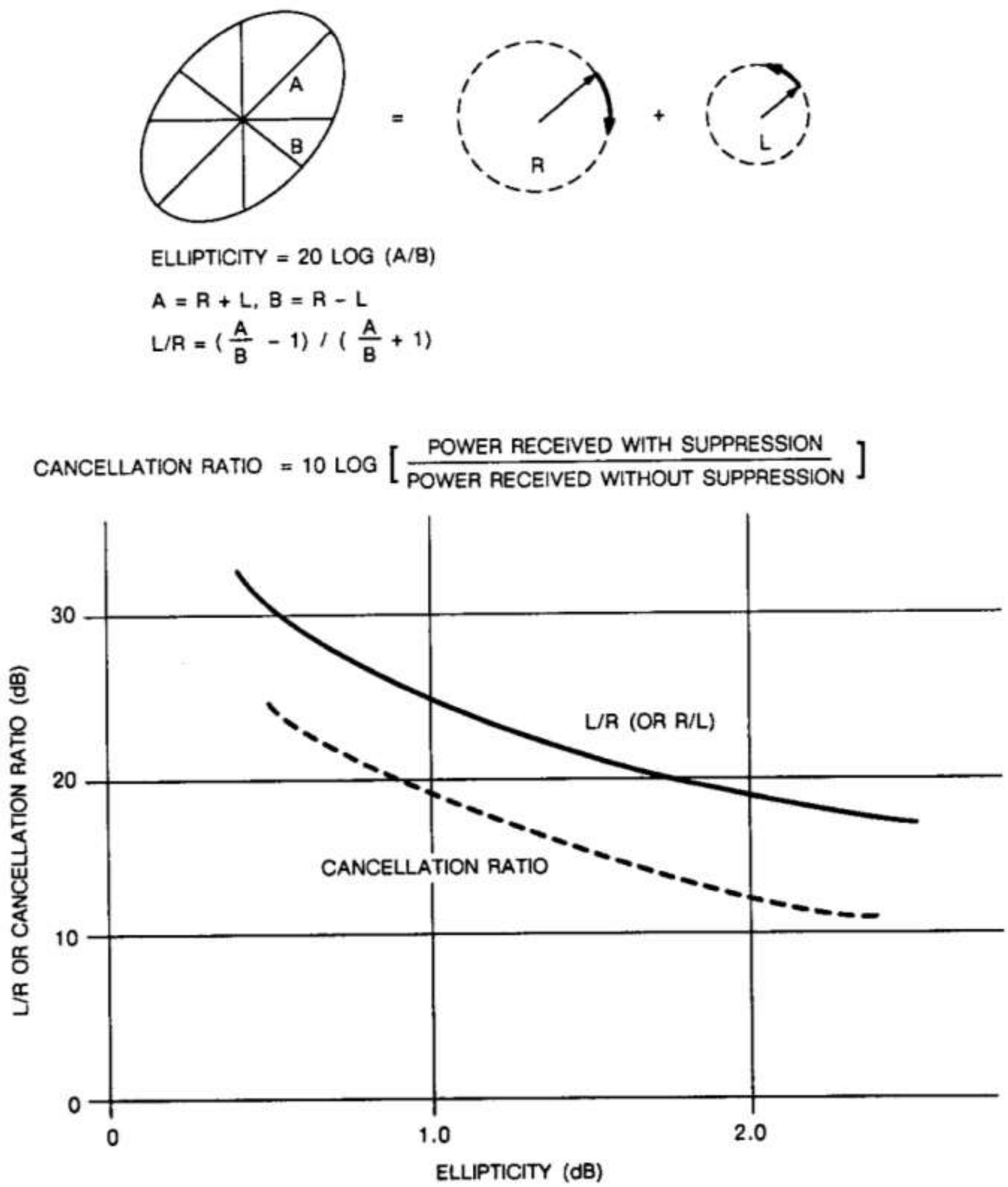


Figure 5.6 Relationships between axial ratio (ellipticity) and cancellation ratio



6. MEASUREMENT ERRORS

The accuracy of gain and pattern measurements depends on how closely the antenna range conforms to establishment of a uniform plane wave over the test antenna. The ground reflection range uses the ground reflection to establish the test field and as such this is not an "undesired" reflection. Undesired reflections can come from trees, fences, buildings, etc. and they cause ripple in the test field. The polarization purity of the source antennas and the reference antennas can also impact the final gain measurements.

Test equipment stability, linearity and dynamic range used to be major potential sources of error. However, with the advent of modern vector network analysers the contributions of the test equipment to measurement errors are now very small. A potential source of error which cannot be calibrated out in the measurement set-up relates to unstable or poor quality coaxial cables or other means to connect the AUT and SGA to the test equipment. The operator must ensure that the test cables make stable and repeatable connections. Poor connector attachment can be a major source of error.

Most low-loss coaxial cables are stable under bending (flexing) provided the minimum bend radius requirements are not exceeded. A major problem with coaxial cables both in amplitude and phase transmission occurs when the cable is twisted along its length (torsion). When interchanging the test cable between the AUT and SGA there should be a minimum of twisting and bending when moving the cable from one antenna to the other. The VSWR (reflection coefficient) and insertion loss of all test cables should be checked before any measurement programme. Any defective (or "suspicious") cables should be replaced at the start of the measurements.

This section will address test range induced errors in gain and pattern measurements. Instrumentation errors are covered in the equipment manuals provided by their suppliers. Extraneous interference caused by transmitters which are not part of the NATR must be treated on a per case basis.

6.1 Pattern measurements

In Section 2.2, the amplitude taper over the test zone was shown to be in the region of 0.25 to 0.5 dB. The far-field phase criterion of Equation 3.1 provides a phase taper over the aperture of 22.5°. The illuminating field on the test aperture is thus not uniform in phase and amplitude as in the true far-field. The non-ideal test field introduces measurement errors in the sidelobes and the gain. While the amplitude and phase tapers act together to give the net results, it is easier to discuss them separately.

6.1.1 Effect of phase taper on sidelobes

Extensive work has been done to quantify the effects of finite test ranges on the measured sidelobes [5] and [7] to [9]. Figure 6.1 shows calculated antenna patterns for a circular parabolic dish of diameter D wavelengths. The horizontal axis is in terms of $(D/\lambda)\sin\theta$ where θ is the azimuth angle. In this sense the patterns are "universal" antenna patterns which can be applied to any circular aperture. The patterns are shown at ∞ range, $2D^2/\lambda$ and $4D2/\lambda$. The nominal infinite range first sidelobe is -28 dB down from the peak. At range $4D^2/\lambda$ this first sidelobe has increased by a few tenths of a dB and the first null is slightly filled in. Note that the second and third sidelobes and second and third nulls are unaffected. This is an extremely important result which is discussed in detail in [7].

The results in Figure 6.1 and those presented in [7] show that by measuring at finite test range the first sidelobe is affected first. The other sidelobes remain unaffected until the first sidelobe makes a shoulder and starts to blend into the main beam. Then the second sidelobe starts to increase leaving the third unaffected, and so on. Patterns similar to Figure 6.1 have been computed for various sidelobe levels and test ranges, however these are not easy to use since many patterns must be inspected to decide on the sidelobe increase. Reference [8] has developed a useful graph which allows the increase in the first sidelobe to be read off directly for various first sidelobe levels and test ranges. This is shown in Figure 6.2. Another representation of the sidelobe error is presented in [9] which has a set of curves showing the error in the first sidelobe (increase) for various first sidelobe design levels (see Figure 6.3). This figure is useful because it shows the normalised range at which the first sidelobe forms a shoulder just before it merges with the main lobe. Observe that the vertical axis in Figure 6.2 is in multiples of D^2/λ while the horizontal axis of Figure 6.3 is in multiples of $2D^2/\lambda$.

The use of Figure 6.2 is illustrated by reference to a 5 m wide antenna operating at 3 GHz. At R = 500 m the test range is $2D^2/\lambda$. The AUT has an actual first sidelobe specification of -20 dB. To achieve 1 dB accuracy project the -20 dB up to the 1 dB line and then across to the range distance which is $2D^2/\lambda$. An actual first sidelobe at -20 dB will thus be measured as -19 dB. This was verified for a 5 m wide S-band air traffic control radar measured on the microwave range. Figure 6.3 gives the same result.

What is also clear from Figures 6.2 and 6.3 is that the higher the accuracy required for a particular first sidelobe level the greater the test range must be. In addition, the lower the design first sidelobe is, the greater the range required to measure this sidelobe to a desired accuracy. In order to measure a 40 dB first sidelobe to an accuracy of 1 dB the test range must be $6D^2/\lambda$ or 1 500 m. Such ultra low sidelobes are difficult to achieve with front-fed reflector antennas. Antenna arrays and carefully designed offset reflectors can achieve these ultra low sidelobes. If the AUT was 3 m in diameter (rather than 5 m); then at 3 GHz the $6D^2/\lambda$ criterion is met at 500 m range.

The microwave range at the NATR can be used to make excellent far-field measurements. The user must be aware of the effects of the test distance on sidelobe measurements (particularly the first sidelobe which increases first) as discussed in this section. The phase taper also has an impact on the gain which is discussed later.

6.1.2 Effect of amplitude taper on sidelobes

A symmetrically tapered amplitude distribution over the AUT as is achieved with the properly aligned microwave range has effects on the measured sidelobes and gain. The taper in the incident field effectively reduces the edge taper on the AUT and thus reduces the gain and sidelobes by a predictable amount. Figure 6.4(a) shows the tapered field distribution over the aperture and the change in gain and sidelobe level [4]. In this case the gain and all the sidelobes are reduced by a small amount. For a typical parabolic reflector with a 10 dB edge taper Figure 6.4(b) shows the gain and sidelobe reductions for various edge tapers. At the microwave range the taper for the largest antennas can be 0.5 dB but for smaller antennas the taper is typically less than 0.25 dB. The gain and sidelobe measurement errors are small.

6.1.3 Effect of extraneous reflections

For the ground reflection range, extraneous reflections refer to reflections from nearby trees, buildings, fences, test fixtures, etc. This is shown schematically in Figure 6.5 where the main beam of the AUT points at some trees while the sidelobes point at the source. The direct signal is picked up on the sidelobe but the interfering reflection from the trees is received on the main beam which may have 40 dB or more gain than the sidelobe being measured. The graph of Figure 6.6 can be used to estimate the impact on the measured sidelobe. The interfering signal could add in phase with the direct signal thereby increasing the measured sidelobe or out of phase which decreases the measured sidelobe.

For example, say the sidelobe in Figure 6.5 is at -40 dB relative to the main beam, if the reflected signal from the trees is 65 dB down from the direct wave from the source, the sidelobe will be measured to an accuracy of 1 dB. However, if the reflection is 55 dB down the accuracy becomes 6 dB. The lower the sidelobes, the lower the extraneous reflections should be.

The microwave range is designed in such a way that there are no reflecting surfaces in the azimuth plane close to the range axis. The operations building is set at an angle to the range axis and reflections generally scatter away from the test zone. It may be possible for undesired reflections to be received by the AUT via the antenna's own mounting fixtures on the receive tower. The user must take precautions to reduce such reflections from his installation.

6.2 Gain measurements

The measured gain is affected by the phase and amplitude tapers. A theoretical study could be done for any particular antenna but for a general purpose test range this is not practical. Extensive work was done by the authors of reference [5] to generate a useful look-up table which can be applied to typical parabolic reflector antennas. Table 6.1 shows a gain correction table. For amplitude and phase tapers the measured gain is lower than the actual gain by the dB values entered in the table. The dB values in the table can be added to the measured gain to get the far-field gain.

Table 6.1 Gain reduction factors relative to true far-field gain as a function of range length (phase taper) and amplitude taper of illuminating field computed for a 10 dB parabolic taper on the dish

Range Length (D^2/λ)	Decibel amplitude taper of illuminating field (sin (x)/(x))						
	0	0.05	0.1	0.15	0.2	0.25	0.5
1	0.204	0.224	0.246	0.265	.286	0.307	0.404
2	0.051	0.071	0.093	0.112	.134	0.154	0.252
3	0.023	0.043	0.065	0.084	.105	0.126	0.224
4	0.013	0.033	0.055	0.074	.096	0.116	0.214
5	0.008	0.029	0.051	0.070	.091	0.111	0.210
6	0.006	0.026	0.048	0.067	.089	0.109	0.207
7	0.004	0.025	0.047	0.066	.087	0.107	0.206
8	0.003	0.024	0.046	0.065	.086	0.106	0.205
9	0.003	0.023	0.045	0.064	.085	0.106	0.204
10	0.002	0.023	0.045	0.064	.085	0.105	0.204
∞	0	0.020	0.043	0.062	.083	0.103	0.202

For 0 dB amplitude taper and ∞ test range, the correction factor is 0 dB. At a range of $2D^2/\lambda$ and 0.5 dB amplitude taper the correction factor is 0.25 dB. At the same range with 0.25 dB amplitude taper the correction is 0.15 dB. It is clear that the gain measurement errors are small when the microwave range is properly configured.

6.3 Polarization errors

It is generally assumed that the source and gain reference antennas are purely linearly polarized (i.e. an infinite axial ratio). The reference horns at the NATR have good linear polarization (35 to 40 dB axial ratio). It has been shown by measurement [1] that the source dishes at the microwave range also have good linear polarization (35 to 40 dB axial ratios).

6.3.1 Linearly polarized AUT

Polarization errors for a linearly polarized source and AUT are discussed in [5]. If the test antenna is linearly polarized with an axial ratio of 25 dB, the error associated with using a range source antenna with finite axial ratio is given in Table 6.2.

Table 6.2 Errors in measured gain for a 25 dB axial ratio AUT and source antennas with various axial ratios.

Source antenna AR (dB)	Errors in measured gain	
	Same sense	Opposite sense
20	-0.035	0.063
25	-0.014	0.041
30	-0.002	0.003
35	0.005	0.022
40	0.009	0.019
45	0.011	0.016
50	0.012	0.015

Note: The linearly polarized antennas are effectively elliptically polarized with large axial ratio. The antennas can be either left or right elliptically polarized.

It is clear that the linearly polarized gain errors are extremely small even for 20 dB axial ratio source antennas. For practical purposes these errors can be neglected on the microwave range.

6.3.2 Circularly polarized AUT

The discussion in [5] is for the measurement of circularly polarized gain using the partial gain method for linearly polarized source antennas (see Section 4.2). Equation 4.3 gives the circular gain for ideal linearly polarized source antenna (infinite axial ratio). The axial ratio and sense of polarization for the linearly polarized source antenna and the sense of polarization for the AUT must be known for compensation of the measured data. The errors in the measured CP gains are given in Table 6.3.

Table 6.3 Errors in measured circularly polarized gain of the AUT using source antennas with various axial ratios.

Source antenna AR (dB)	Errors in measured CP gain	
	Same sense	Opposite sense
20	0.828	-0.915
25	0.475	0.505
30	0.270	0.279
35	0.153	0.156
40	0.086	0.109
45	0.049	0.049
50	0.027	0.028

Comparison of Tables 6.2 and 6.3 shows that the CP gain measurement is far more sensitive to the axial ratio of the source antenna. If the source is only linearly polarized to 20 dB axial ratio, the measurement errors are nearly ±1 dB for CP and only 0.06 dB for linear polarization. At 35 dB axial ratio the gain error is 0.15 dB. This emphasises the fact that excellent linear polarization is required in the test field when measuring circularly polarized antennas. The microwave range achieves 35 to 40 dB.

The primary requirement for testing circularly polarized antennas at the microwave range has been for radar antennas which have a circularly polarized mode for rain clutter reduction. Clearly good axial ratio measurements are required to calculate the integrated cancellation ratio discussed in Section 5.3.

Many electronic warfare and satellite communications systems use circularly polarized antennas. These antennas can be measured to good accuracy on the microwave range. In these cases custom designed test set-ups may be required depending on the specified measurement accuracy.

Figure 6.1 Calculated antenna patterns illustrating the effects of quadratic phase errors encountered when measuring at the test ranges shown

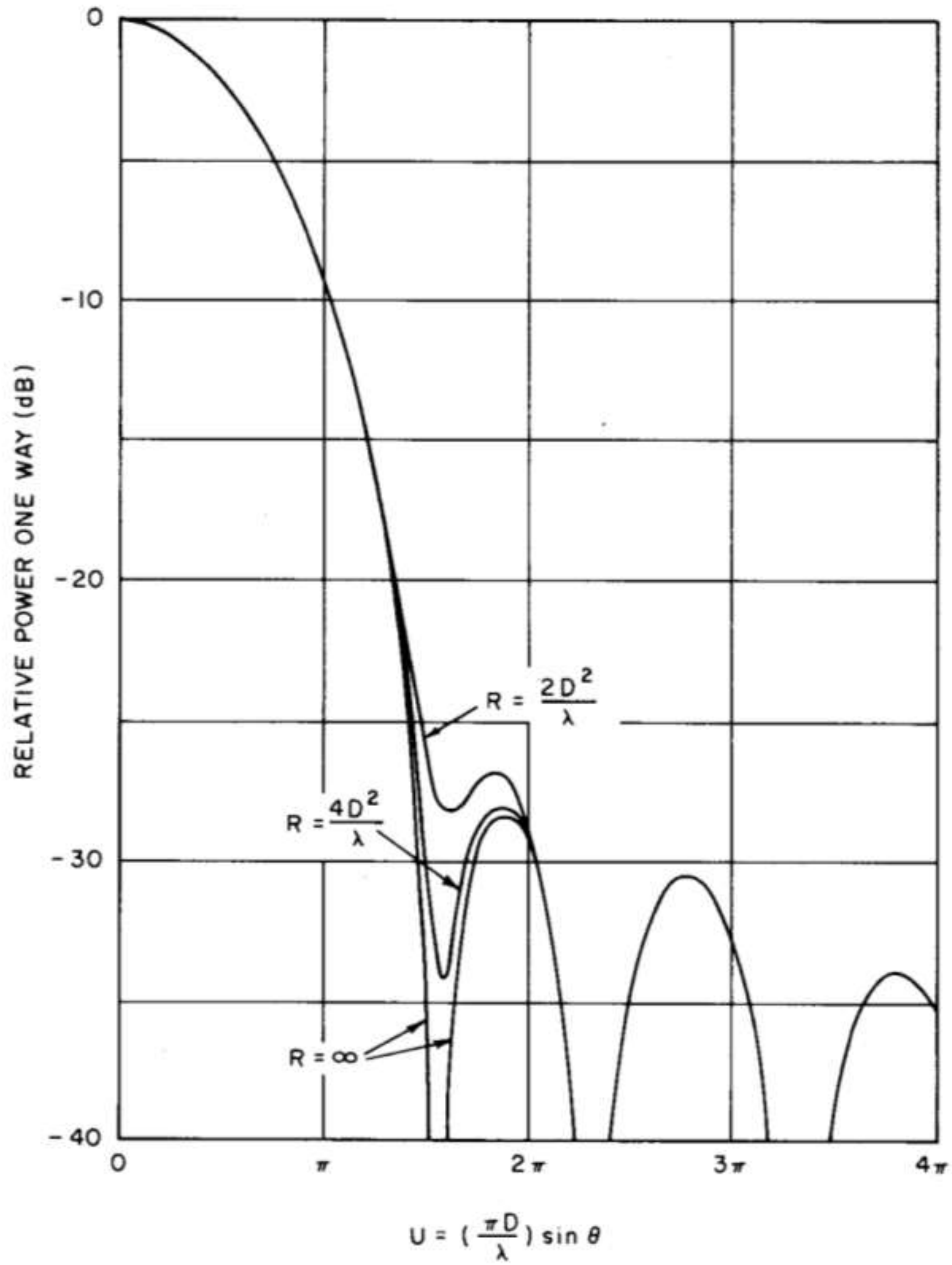


Figure 6.2 Range distance required for measuring the correct first sidelobe to the accuracy (error) stated on the straight lines

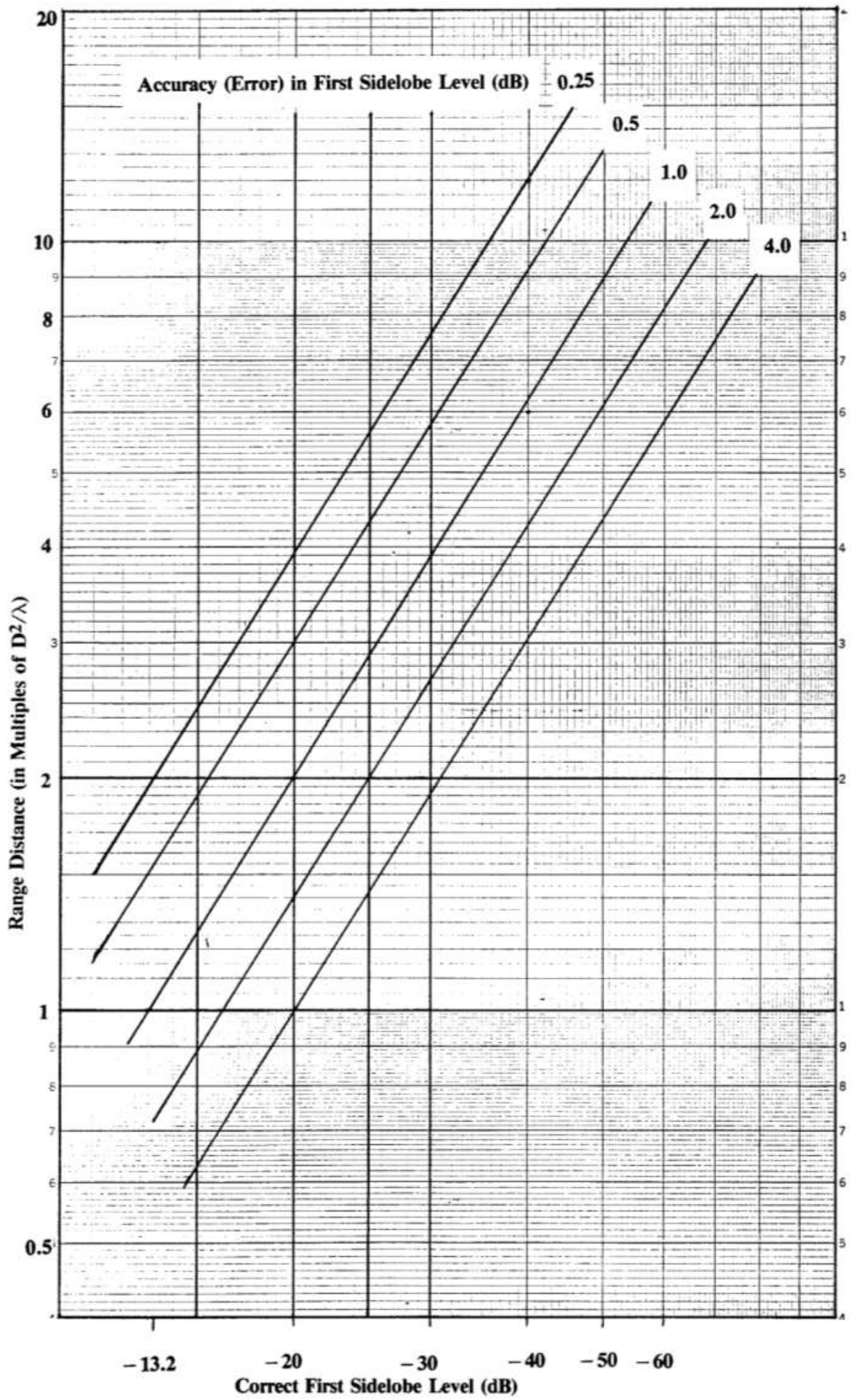


Figure 6.3 First sidelobe change versus normalised far-field distance for various first sidelobe design levels (note onset of shoulders)

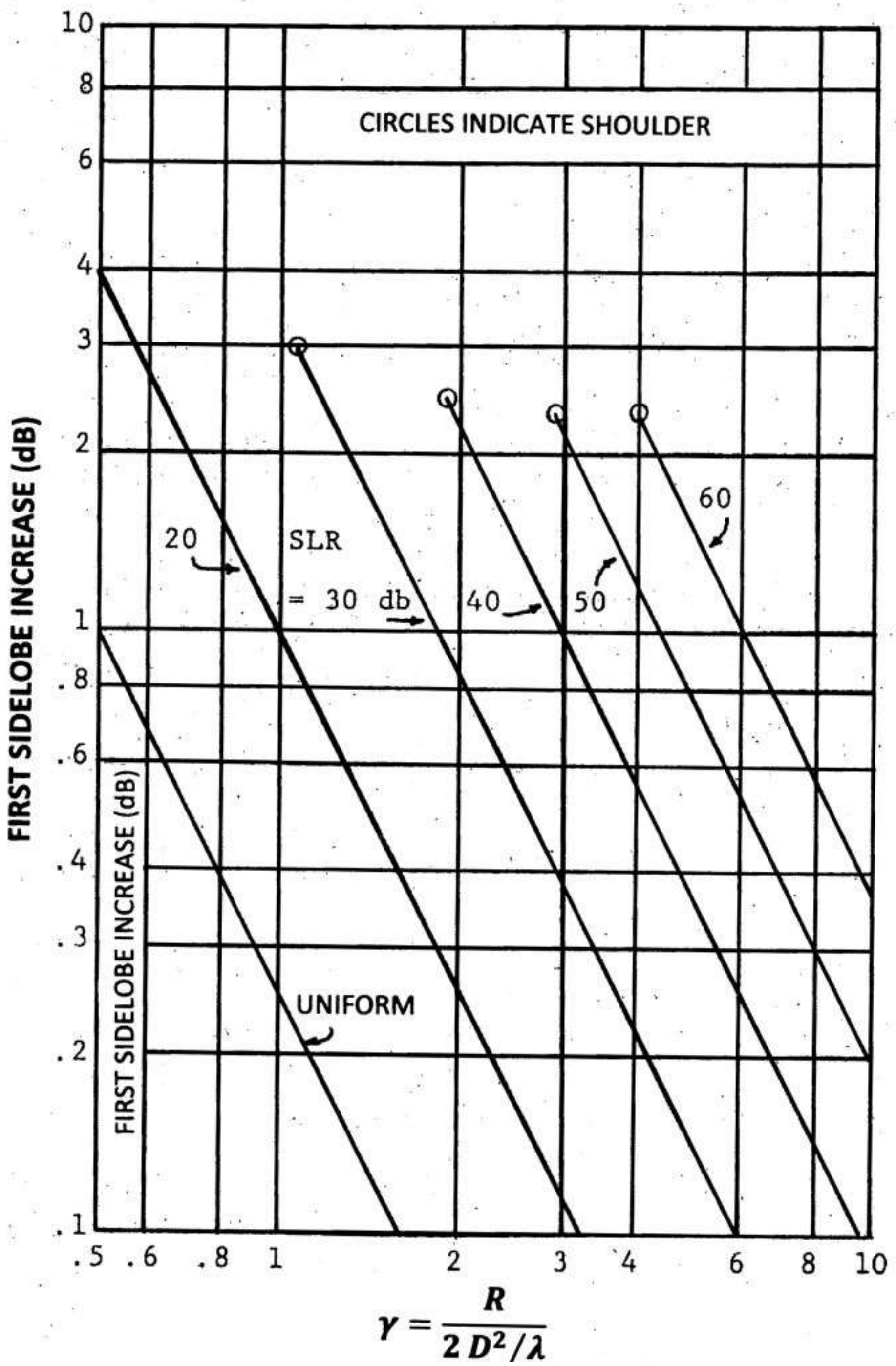


Figure 6.4 Effect of amplitude taper in the test field
(a) explanation of terms
(b) impact on gain and first and second sidelobes

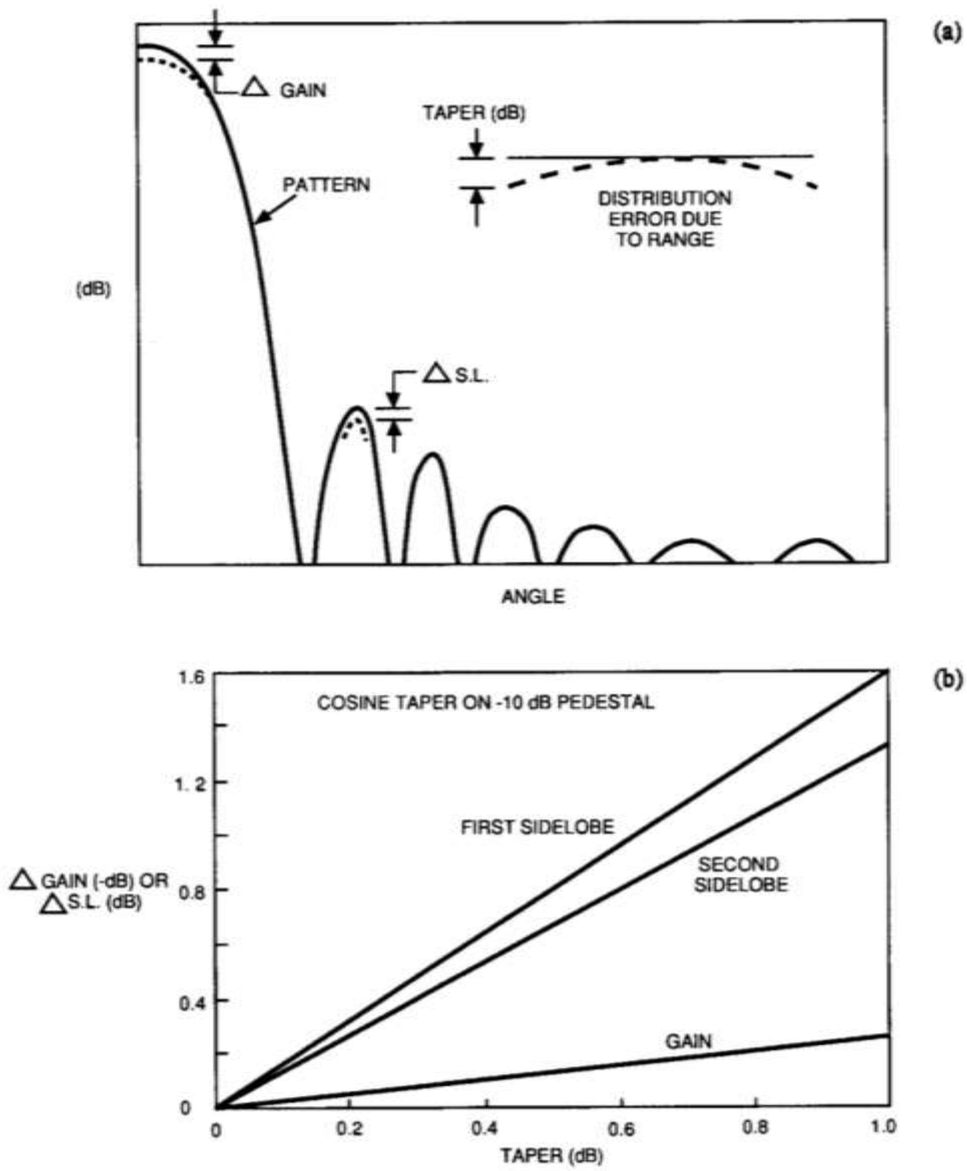


Figure 6.5 Diagram showing extraneous reflection entering main beam when sidelobe points at source

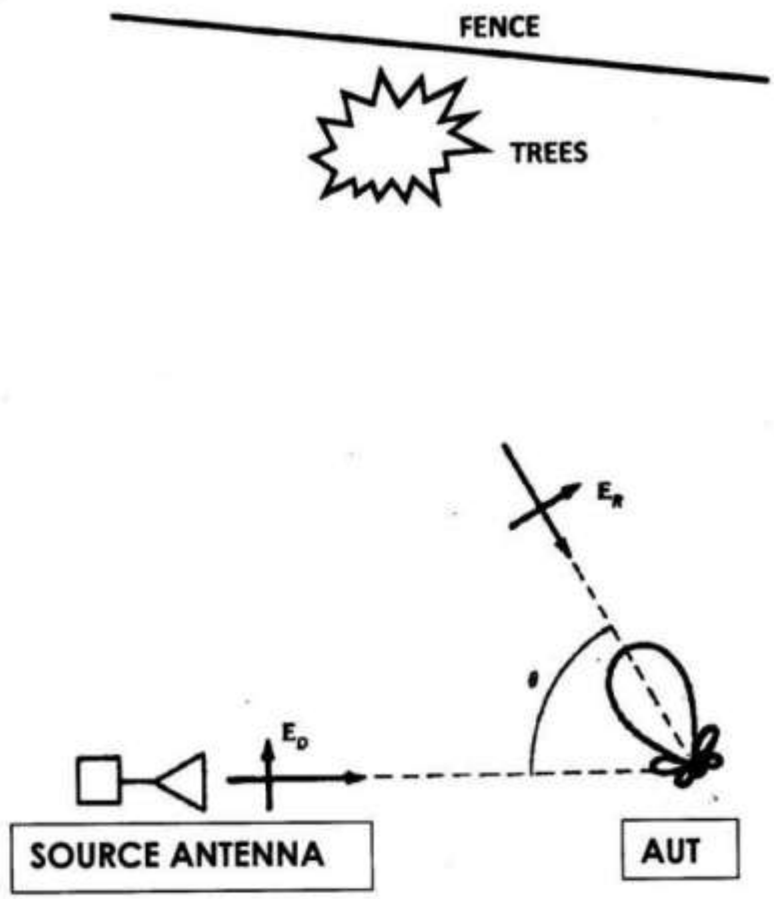
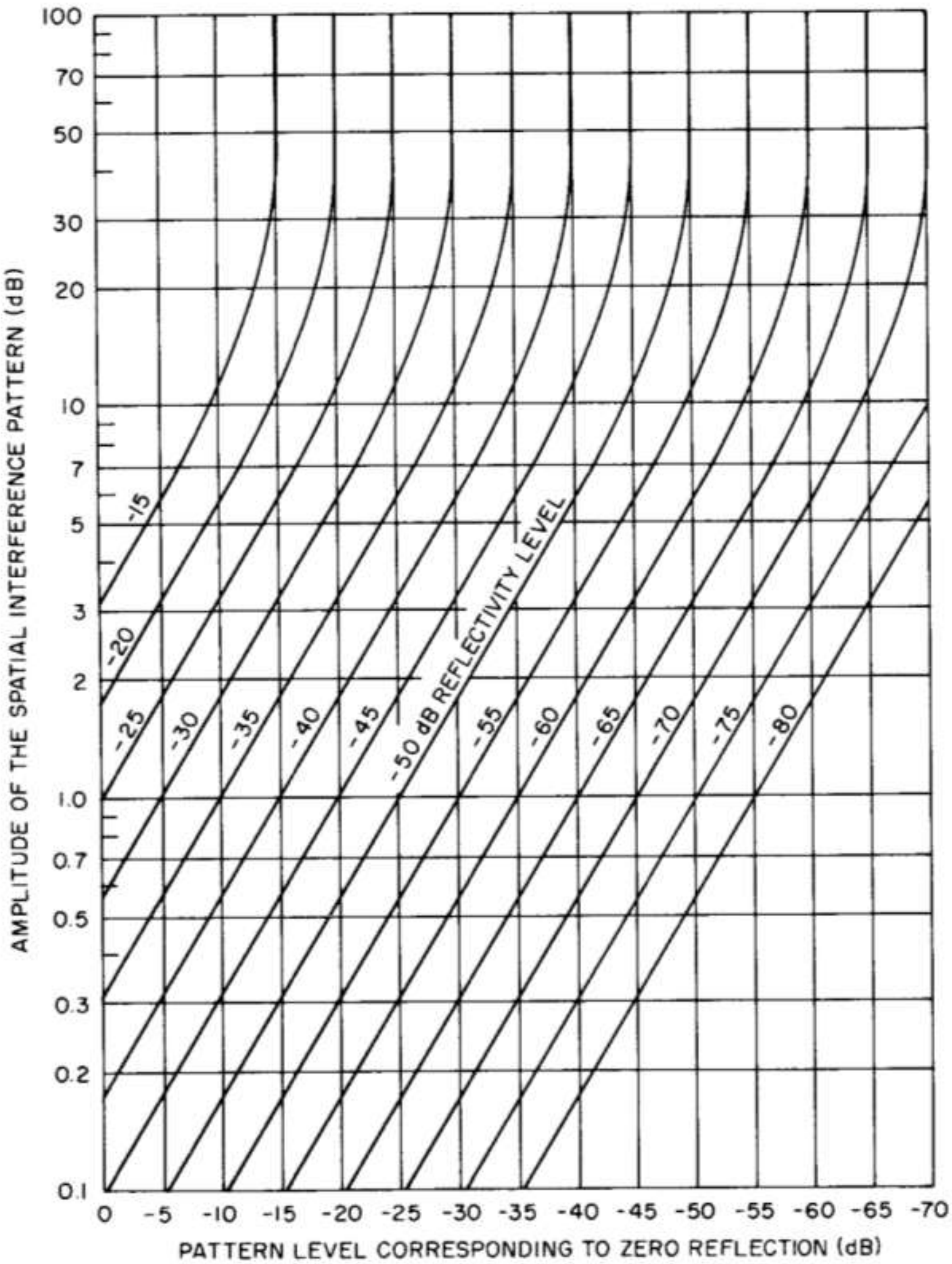


Figure 6.6 Amplitude of the spatial interference level for a given reflectivity level and antenna pattern level



7. SQUINT AND BORESIGHTING

7.1 Squint and boresight calibration

Squint is a condition where a specified axis of an antenna (e.g. the peak of the main beam) deviates from a specified reference axis. This is illustrated schematically in Figure 8.1. The x-axis shows the angles from the test positioner readout and the measured pattern has an offset relative to 0˚. Boresighting or boresight calibration is the process which locates the angular position of the electrical axis relative to an identifiable mechanical axis. Boresight calibration is most often required on antennas with narrow antenna patterns (main beams). These narrow beam antenna types (beamwidth typically less than 10˚) fall into two main groups:

(a) Antennas which produce a null in the sensing function at boresight (these include monopulse antennas, con-scan radar antennas, SSR antennas, beam-switching radar antennas, etc.), and

(b) Antennas which produce a maximum in the sensing function at boresight (these include search radars with mechanical or electronic scan, height finders, mortar locators, radio astronomy antennas, etc.).

For narrow beam antennas (e.g. typical high gain radar antennas) the beam peak position is often taken as the electrical axis. Sometimes it is better to use the angular positions at nominally 1 dB below the beam peak to the left and the right of the beam peak to calculate the position of the peak. The position of the beam peak is then calculated as

$$\theta\,(\text{peak}) = \tfrac{1}{2}\,[\,\theta\,(1\text{ dB left}) + \theta\,(1\text{ dB right})\,] \qquad (8.1)$$

where the angles maintain the sign on the pattern, i.e. pattern angles to the left of 0˚ in Figure 8.1 are negative.

For antennas with wide antenna patterns, e.g. CrSM base station antennas with beamwidths of around 65˚, the squint is often calculated at the 3 or 6 dB pattern value. Thus

$$\text{SQUINT} = \tfrac{1}{2}\,[\,\theta\,(3\text{ dB left}) + \theta\,(3\text{ dB right})\,] \qquad (8.2)$$

Boresight calibrations are often done to align and make the electrical and mechanical axes coincident (or sometimes simply to make them parallel). Alternately, the boresight calibration must achieve a fixed off-set angle between the electrical and mechanical axes (e.g. the off-set angle in a con-scan radar antenna or the slight upward squint in the cosecant squared pattern of a search radar antenna. The mechanical reference is often called the "optical axis" and it is fixed with reference to the antenna and its mechanical positioning structure. The term optical axis results because this axis is often established using optical means such as a boresighting telescope, mirrors, laser beams, etc.

The mechanical axis is often defined by a reference face (e.g. the rear mounting flange of a parabolic reflector), dowel pins and holes, a telescope bracket axis, etc. Dowel pins and holes are vital in cases where the AUT must be removed from the test positioner to make adjustments to the antenna (AUT). After such adjustments the AUT must be installed in a mechanically identical way to before adjustments.

It is important that the test range does not introduce errors into the measurements because of improper characteristics of the illuminating field in terms of phase curvature and amplitude taper. The Microwave Range of the NATR was designed with these factors taken into account. When the range is properly configured the main criterion that the illuminating field is *symmetrical* in phase and amplitude over the AUT will be met.

Monopulse antennas present particular challenges if the null depths are deep (- 30 dB) and narrow (less than 0.1˚) and if the antennas are physically large. There is some residual backlash in the gear trains of the positioners (around 0.05˚). The effects of backlash can be reduced if all boresight alignment measurements are made when rotating the antenna in the same direction for all measurements. Large antennas with high wind wading should be measured when conditions are still and there are no wind gusts.

In the azimuth plane the centre of the source antenna at the 500 m source tower sets the range axis. This means that the source position is known and all measurements of the azimuth squint can be made relative to this position. For some antennas it may be possible to turn the antenna upside down (e.g. if the antenna is mounted on the upper azimuth turntable which can then act as a roll axis). The squint can then be checked in the normal and inverted conditions to ensure that the electrical axis stays at the same angle in both conditions.

7.2 Boresight alignment of source dish antennas

All the parabolic reflectors (dishes) at the NATR have removable feeds (see Table 3.2). When a dish is mounted on the polarization positioner of the source tower the rear mounting flange of the dish is the mechanical reference face. The line at right angles to this face is the mechanical axis of the dish. The mechanical axis of the dish is parallel to and coincident with the rotation axis of the positioner. If the electrical axis of the dish antenna pattern (the line containing the peak of the main beam) is not parallel to the mechanical axis (and ideally coincident with it) the peak of the pattern will not point at the centre of the test zone as the polarization is rotated. The antenna beam will move on a cone almost as for a con-scan radar antenna. This means that the symmetrical point in the centre of the test zone will shift as a function of polarization. This will introduce measurement errors.

This section describes how the source dishes can be aligned by installing them on the receive tower. A source antenna which does not have too narrow a beamwidth (horn or small dish) is installed at the source tower on the polarization positioner. This source antenna is set at the desired height determined from Equation 2.1 at the frequency of interest. The source antenna will be rotated in polarization but because it has a relatively wide pattern, small off-sets in its mechanical axes are not critical. For a particular set of E- or H-plane cut the source remains fixed.

The discussion below is for the 1.22 m diameter dish with the S-band horn feed at 3.3 GHz. The process is similar for other dishes and feeds. The source dish is mounted directly on the front face of the upper azimuth turntable. The elevation axis is tilted forward in elevation until the received signal reaches a peak (elevation at about 91.4˚). With the dish mounted as above, the upper elevation axis acts as a polarization or roll axis for the dish.

The source antenna is set for horizontal polarization and the upper azimuth axis is used as a roll positioner to set the 1.22 m for horizontal polarization (0˚). An azimuth cut is then made with the lower azimuth turn table. The measurement is then repeated with the 1.22 m dish rolled through 180˚. These two E-plane cuts are shown in Figure 8.2. If the mechanical and electrical axes of the dish were coincident the two patterns would track each other over most of the main beam. There is an off-set of about 0.2˚ between the two patterns. The horn feed of the 1.22 m dish has an adjusting mechanism to move the feed left and right. By moving the feed, the peak of the main beam is moved. If the adjustment is in the wrong direction then the off-set will increase. After final adjustment of the feed in the E-plane the patterns in Figure 8.3 are achieved. The off-set is now very small.

The source antenna at the source tower is now set for vertical polarization and the 1.22 m dish on the roll axis of the test positioner is rotated to the 90˚ position. An H-plane pattern is measured using the lower azimuth position. The H-plane pattern is then repeated for the roll axis set for 270˚. These two patterns are shown in Figure 8.4. The two patterns are off-set by about 0.8˚. The feed is now adjusted in the H-plane until the 90˚ and 270˚ roll axis plots overlap. This is shown in Figure 8.5 where the two patterns now lie on top of each other. As a final check return to the E-plane (0˚ and 180˚) and check that the H-plane adjustments have not disturbed the E-plane patterns.

With the E- and H-plane patterns in Figures 2.2 and 2.4, an axial ratio error of at least 0.5 dB will be made when measuring circularly polarized antennas. When the feed electrical axis is aligned to the mechanical axis of the dish, the dish does not squint as it is rotated about its axis. The above adjustments are time consuming. For waveguide band horn feeds, it is often good enough to make the E- and H-plane adjustments at the highest frequency in the band of interest. For LPDA feeds where the feed phase centre moves along the length of the LPDA, the adjustments should be checked at the low, mid and high frequencies. The dish beam is the narrowest at the highest frequency and the adjustments are not critical for the highest frequencies. Note that at a distance of 500 m a test zone diameter of 5.5 m only uses the centre ± 0.3˚ of the pattern of the properly aligned source dish. In this angular sector the peak of the beam (Figures 2.3 and 2.5) is nearly constant.

Figure 8.1 Typical antenna pattern showing squint between mechanical axis at 0° and beam peak which defines the electrical axis

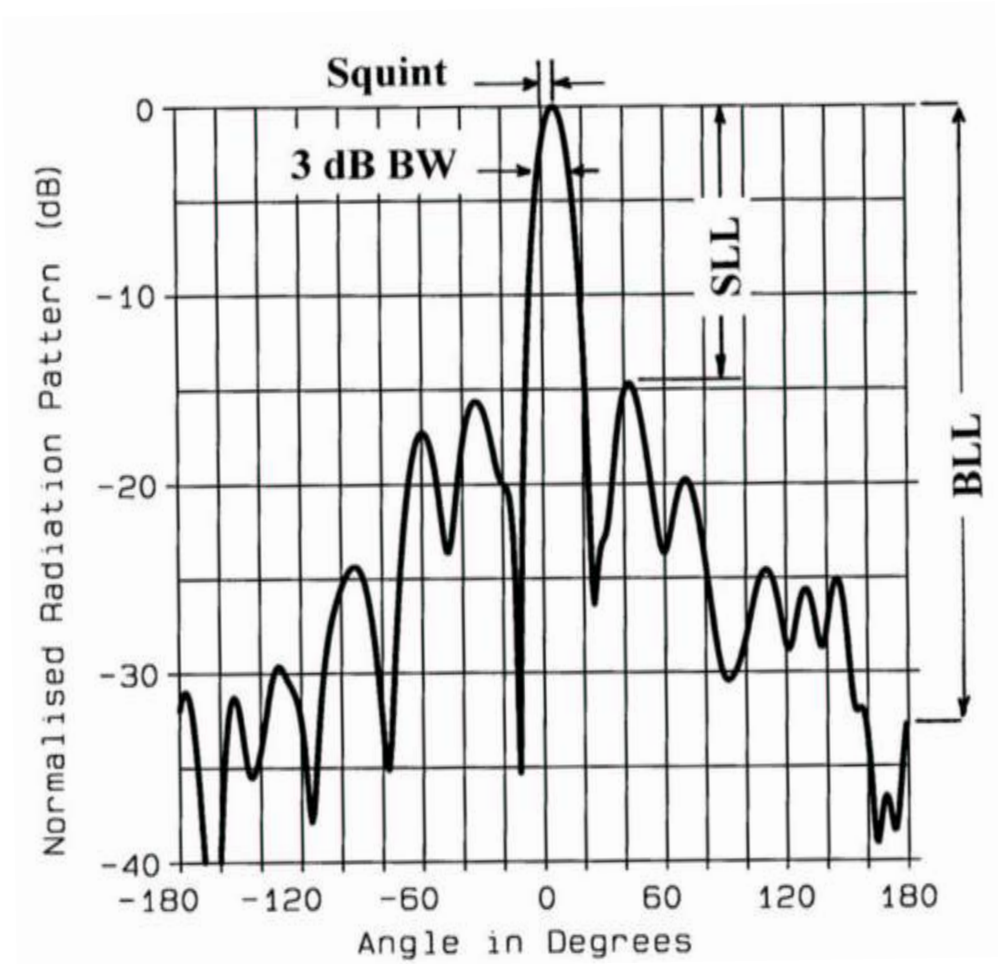


Figure 8.2 E-plane patterns of 1.22 m diameter source dish for S band with 0° and 180° receive polarization before alignment

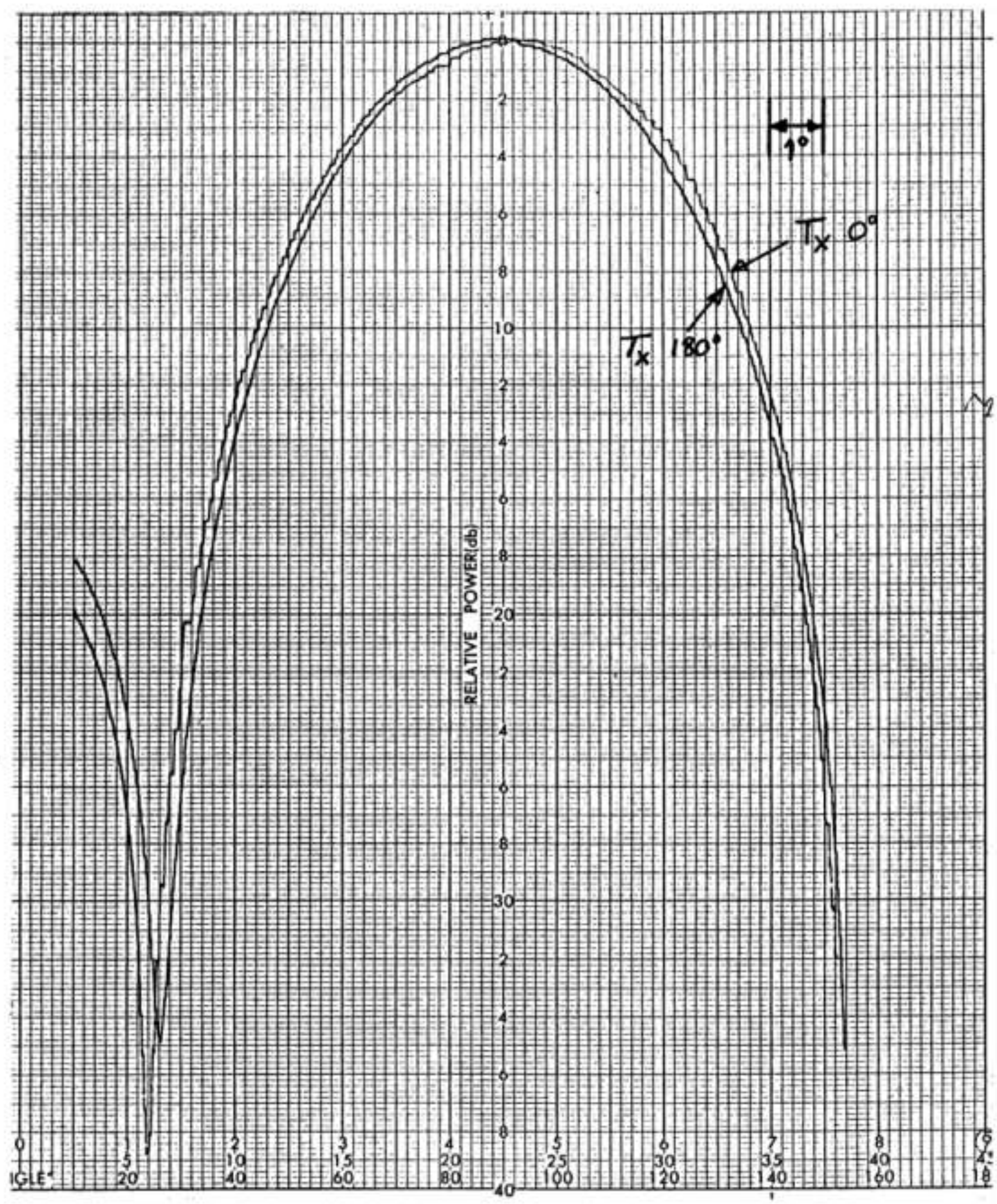


Figure 8.3 E-plane patterns of 1.22 m diameter source dish for S band with 0° and 180° receive polarization after alignment

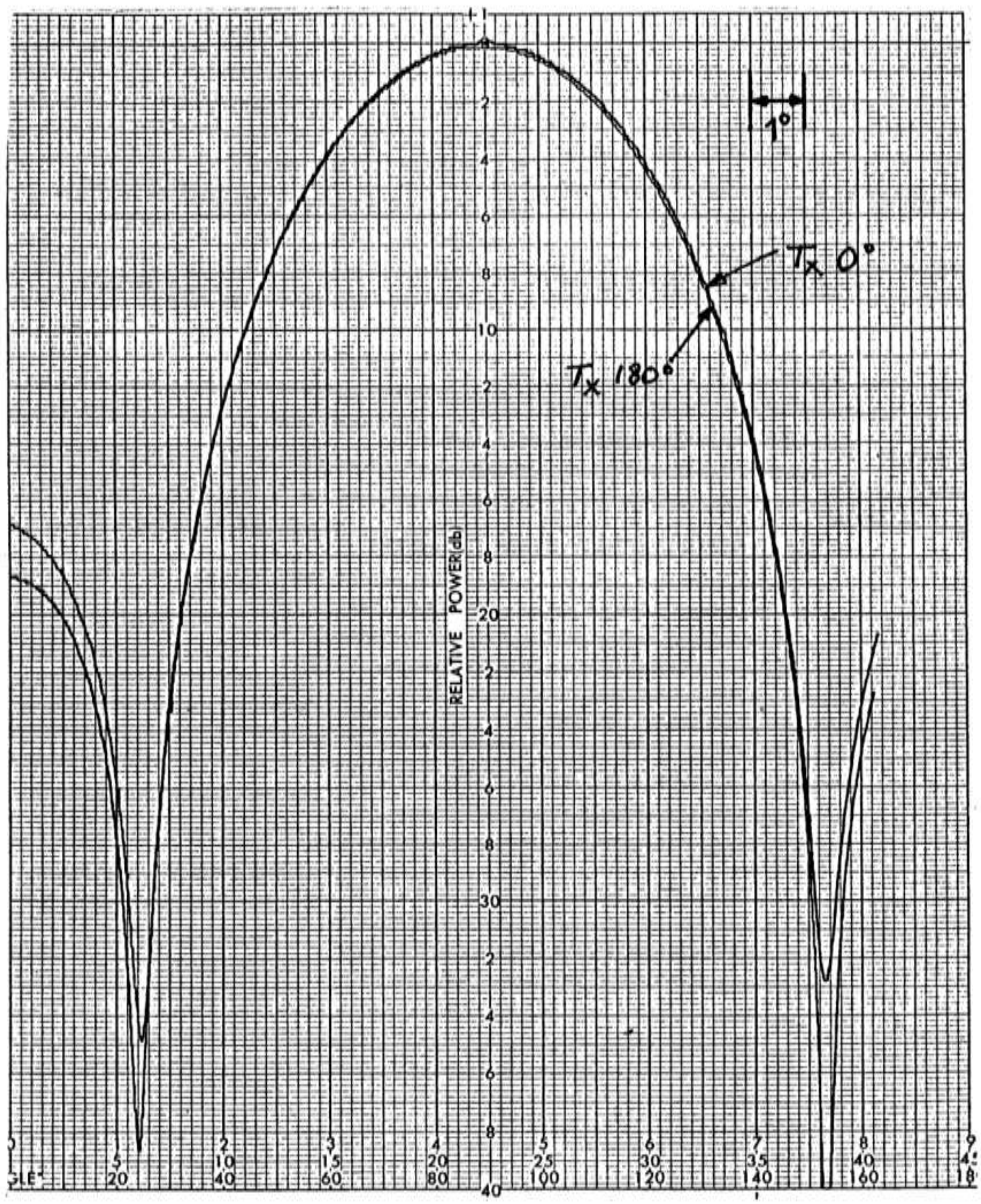


Figure 8.4 H-plane patterns of 1.22 m diameter source dish for S band with 90° and 270° receive polarization before alignment

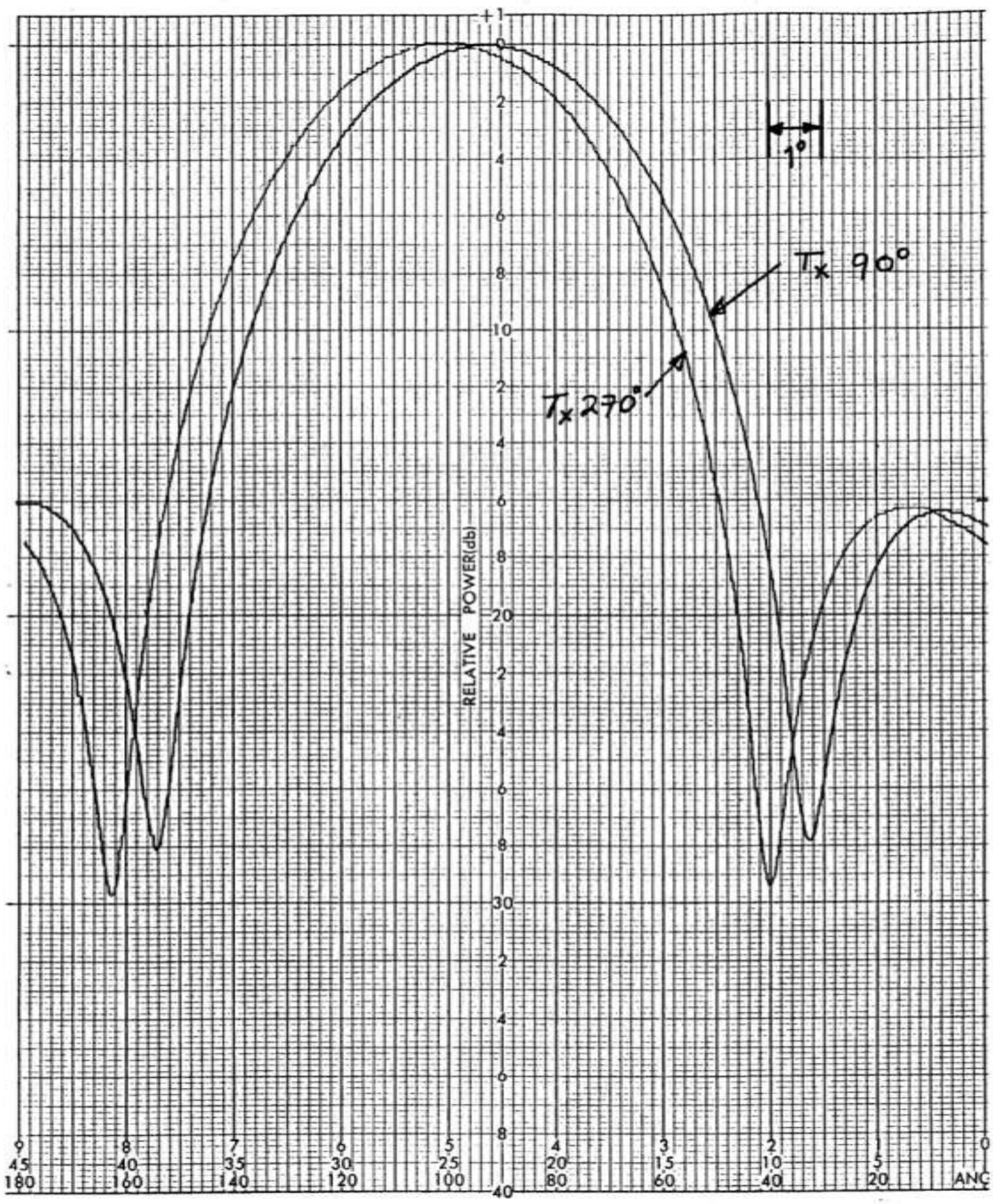
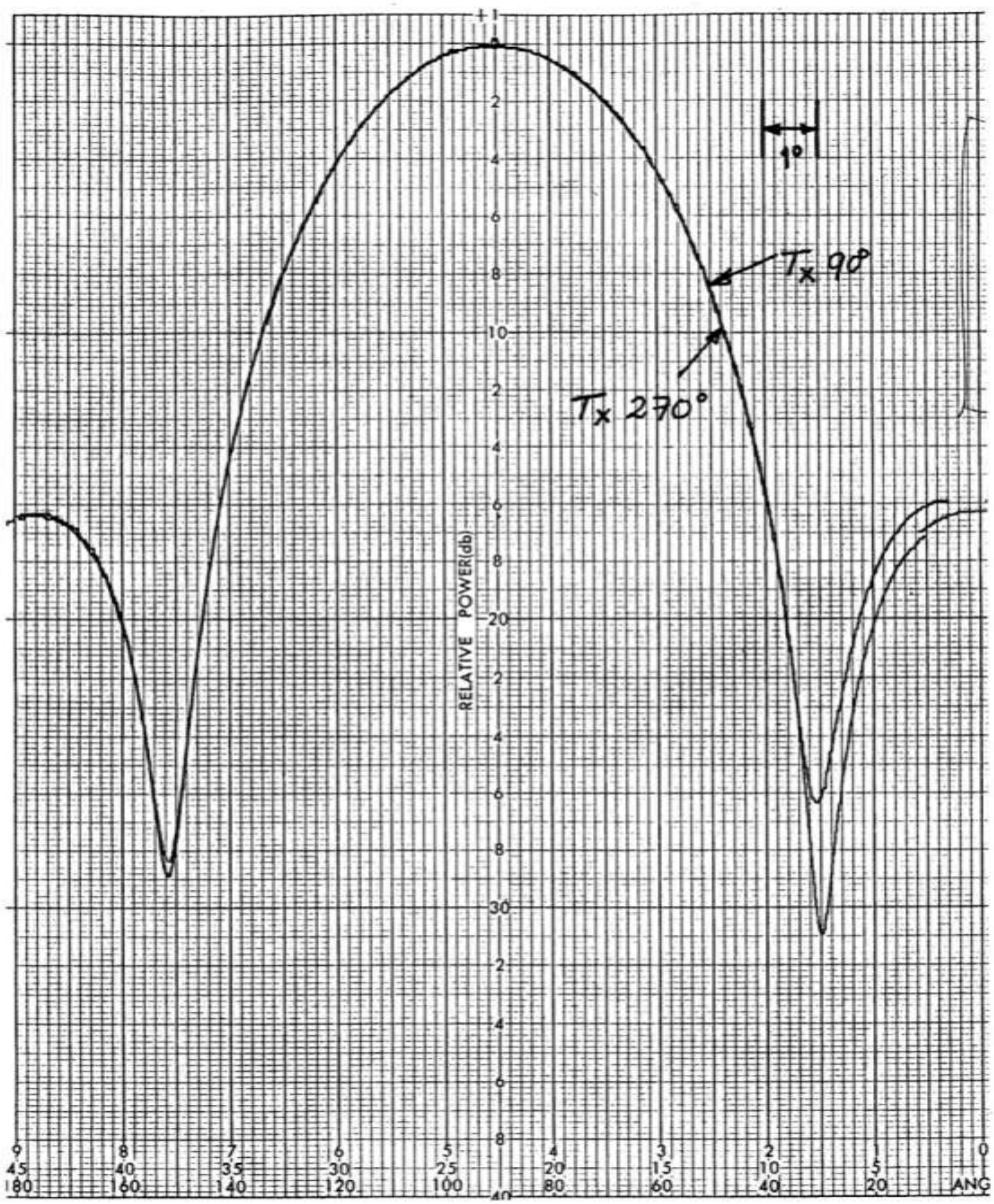


Figure 8.5 H-plane patterns of 1.22 m diameter source dish for S band with 90° and 270° receive polarization after alignment



8. CONCLUSION

This user guide has described the requirements for the operation of the 500 m ground reflection microwave range under controlled and well understood conditions. The largest diameter antennas for the 0.5 to 40 GHz frequency range have been derived based on the $2D^2/\lambda$ far-field criterion. Linear and circular polarization measurements can be made and the range capabilities have been verified in [1]. Pattern and gain measurement errors were discussed. The range has excellent cross polarization (typically beter than 35 dB). Careful alignment of the mechanical and electrical axes of the source antennas will improve the cross polarization. In many applications cross polarization is becoming an important parameter. Modern vector network analyzers and receivers with sophisticated self-calibration capabilities have greatly reduced instrumentation errors.

The 500 m ground reflection microwave range can make accurate measurements on antennas which meet the $2D^2/\lambda$ criterion. Measurements can be made on antennas at test distances less than this requirement. At shorter test distances only the first (or possibly also the second) sidelobe will be in error. The impact on the gain measurement can be deduced from Table 6.1 when the test distance and amplitude taper are known.

The Microwave Range is suitable for measuring radar antennas, IFF antennas, array antennas, reflector antennas, etc. in its ground-reflection mode. This range can also be configured to make excellent measurements at HF/VHF/UHF frequencies (300 kHz to 3 GHz and above). This versatility and proven performance confirm the status of the Microwave Range as a world-class facility.

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APPENDIX: GLOSSARY OF STANDARD ANTENNA TERMS

This glossary is an extract taken from the IEEE definitions and defines some of the most commonly used antenna terms (see IEEE Std. 149-1979, *Definition of terms for Antennas* for a complete listing). Consistent application of standard definitions will eliminate confusion relating to antenna specifications and interpretation of measured results.

Antenna efficiency of an aperture-type antenna. For an antenna with a specified planar aperture, the ratio of the maximum effective area of the antenna to the aperture area.

Aperture of an antenna. A surface, near or on an antenna, on which it is convenient to make assumptions regarding the field values for the purpose of computing fields at external points.
Note: The aperture is often taken as that portion of a plane surface near the antenna, perpendicular to the direction of maximum radiation, through which the major part of the radiation passes.

Aperture illumination. The field over the aperture as described by amplitude, phase and polarization distributions.

Axial ratio (of a polarization ellipse). The ratio of the major to minor axes of a polarization ellipse. The ratio varies from infinity to 1 as the polarization changes from linear to circular.

Beam. The major lobe of the radiation pattern.

Bistatic cross section. The scattering cross section in any specified direction other than back toward the source.

Cardinal plane. For an infinite planar array whose elements are arranged in a regular lattice, any plane of symmetry normal to the planar array and parallel to an edge of a lattice cell.

Circularly polarized field vector. At a point in space, a field vector whose extremity describes a circle as a function of time.
Note: Circular polarization may be viewed as a special case of elliptical polarization where the axial ratio has become equal to one.

Copolarization. The polarization that the antenna is intended to radiate or receive.

Cross polarization. In a specified plane containing the reference polarization ellipse, the polarization orthogonal to a specified reference polarization.
Note: The reference polarization usually is the copolarization. Two fields have orthogonal polarizations if their polarization ellipses have the same axial ratio, major axes at right angles, and opposite senses of rotation. If the reference polarization is right-handed circular, the cross polarization is left-handed circular, and vice versa.

Directivity (of an antenna) (in a given direction). The ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions.
Notes: (1) The average radiation intensity is equal to the total power radiated by the antenna divided by 4π. (2) If the direction is not specified, the direction of maximum radiation intensity is implied.

Directivity, partial (of an antenna for a given polarization). In a given direction, that part of the radiation intensity corresponding to a given polarization divided by the total radiation intensity averaged over all directions.

Note: The (total) directivity of an antenna, in a specified direction, is the sum of the partial directivities for any two orthogonal polarizations.

Effective area of an antenna (in a given direction). In a given direction, the ratio of the available power at the terminals of a receiving antenna to the power flux density of a plane wave incident on the antenna from that direction, the wave being polarization-matched to the antenna.

Notes: (1) If the direction is not specified, the direction of maximum radiation intensity is implied. (2) The effective area of an antenna in a given direction is equal to the square of the operating wavelength times its gain in that direction divided by 4π.

Elliptically polarized field vector. At a point in space, a field vector whose extremity describes an ellipse as a function of time.

Note: Any single-frequency field vector is elliptically polarized if *elliptical* is understood in the wide sense as including circular and linear. Often, however, the expression is used in the strict sense meaning noncircular and nonlinear.

E plane, principal. For a linearly polarized antenna, the plane containing the electric field vector and the direction of maximum radiation.

Equivalent flat plate area of a scattering object. For a given scattering object, an area equal to the wavelength times the square root of the ratio of the monostatic cross section to 4π.

Note: A perfectly reflecting plate parallel to the incident wavefront and having this area, if it is large compared to the wavelength, will have approximately the same monostatic cross section as the object.

Far-field region. That region of the field of an antenna where the angular field distribution essentially is independent of the distance from a specified point in the antenna region.

Note: In free space, if the antenna has a maximum overall dimension, D, which is large compared to the wavelength, the far-field region commonly is taken to exist at distances greater than $2D^2/\lambda$ from the antenna, λ being the wavelength. The far-field patterns of certain antennas, such as multibeam reflector antennas, are sensitive to variations in phase over their apertures. For these antennas $2D^2/\lambda$ may be inadequate.

Fraunhofer region. The region in which the field of an antenna is focused.

Note: In the Fraunhofer region of an antenna focused at infinity, the values of the fields, when calculated from knowledge of the source distribution of an antenna, are sufficiently accurate when the quadratic phase terms (and higher order terms) are neglected.

Fresnel region. The region (or regions) adjacent to the region in which the field of antenna is focused (that is, just inside the Fraunhofer region).

Note: In the Fresnel region in space, the values of the fields, when calculated from knowledge of the source distribution of an antenna, are insufficiently accurate unless the quadratic phase terms are taken into account, but are sufficiently accurate if the quadratic phase terms are included.

Front-to-back ratio. For directional antennas, the ratio of the antenna's effectiveness toward the front to its effectiveness toward the back (often expressed as a dB value).

Gain, absolute (of an antenna) (in a given direction). The ratio of the radiation intensity, in a given direction, to the radiation intensity that would be obtained if the power accepted by the antenna were radiated isotropically.

Notes: (I) Gain does not include losses arising from impedance and polarization mismatches. (2) The radiation intensity corresponding to the isotropically radiated power is equal to the power accepted by the antenna divided by 4π. (3) If an antenna is without dissipative loss, then in any given direction, its gain is equal to its directivity. (4) If the direction is not specified, the direction of maximum radiation intensity is implied. (5) The term absolute gain is used in those instances where added emphasis is required to distinguish gain from relative gain; for example, absolute gain measurements.

Gain, partial (of an antenna for a given polarization). In a given direction, that part of the radiation intensity corresponding to a given polarization divided by the radiation intensity that would be obtained if the power accepted by the antenna were radiated isotropically.
Note: The (total) gain of an antenna, in a specified direction, is the sum of the partial gains for any two orthogonal polarizations.

H plane, principal. For a linearly polarized antenna, the plane containing the magnetic field vector and the direction of maximum radiation.

Intercardinal plane. Any plane that contains the intersection of two successive cardinal planes and is at an intermediate angular position.
Note: In practice, the intercardinal planes are located by dividing the angle between successive cardinal planes into equal parts. Often, it is sufficient to bisect the angle so that there is only one intercardinal plane between successive cardinal planes.

Isotropic radiator. A hypothetical, lossless antenna having equal radiation intensity in all directions.
Note: An isotropic radiator represents a convenient reference for expressing the directive properties of actual antennas.

Linearly polarized field vector. At a point in space, a field vector whose extremity describes a straight line segment as a function of time.
Note: Linear polarization may be viewed as a special case of elliptical polarization, where the axial ratio has become infinite.

Major lobe; main lobe. The radiation lobe containing the direction of maximum radiation.
Note: In certain antennas, such as multi-lobed or split-beam antennas, there may be more than one major lobe.

Mean sidelobe level. The average value of the relative power pattern of an antenna taken over a specified angular region, which excludes the main beam, the power pattern being relative to the peak of the main beam.

Minor lobe. Any radiation lobe except a major lobe. *See sidelobe*.

Monostatic cross section; backscattering cross section. The scattering cross section in the direction toward the source.
Note: Compare this term with the term *bistatic cross section*.

Near-field region. That part of space between the antenna and far-field region.
Note: In lossless media, the near-field may be further subdivided into reactive and radiating near-field regions.

Orthogonal polarization (with respect to a specified polarization). In a common plane of polarization, the polarization for which the inner product of the corresponding polarization vector and that of the specified polarization is equal to zero.
Notes: (I) The two orthogonal polarizations can be represented as two diametrical points on the Poincare sphere. (2) Two elliptically polarized fields having the same plane of polarization have

orthogonal polarizations if their polarization ellipses have the same axial ratio, major axes at right angles, and opposite senses of polarization.

Omnidirectional antenna. An antenna having an essentially non-directional pattern in azimuth and a directional pattern in elevation. The deviation from omni in the azimuth pattern is the difference between the maximum gain in the pattern (wherever it occurs in the pattern) and the minimum gain (wherever it occurs in the pattern) divided by two. This difference is expressed as a ± dB value.

Phase center. The location of a point associated with an antenna such that, if it is taken as the center of a sphere whose radius extends into the far-field, the phase of a given field component over the surface of the radiation sphere is essentially constant, at least over that portion of the surface where the radiation is significant.

Note: Some antennas do not have a unique phase center.

Plane of polarization. A plane containing the polarization ellipse.

Note: For a plane wave in an isotropic medium, the plane of polarization is taken to be normal to the direction of propagation.

Polarization efficiency; polarization mismatch factor. The ratio of the power received by an antenna from a given plane wave of arbitrary polarization to the power that would be received by the same antenna from a plane wave of the same power flux density and direction of propagation, whose state of polarization has been adjusted for a maximum received power.

Note: The polarization efficiency is equal to the magnitude of the inner product of the polarization vector describing the receiving polarization of the antenna and the polarization vector of the plane wave incident at the antenna.

Polarization pattern (of an antenna). (1) The spatial distribution of the polarizations of a field vector excited by an antenna taken over its radiation sphere. (2) The response of a given antenna to a linearly polarized plane wave incident from a given direction and whose direction of a polarization is rotating about an axis parallel to its propagation vector; the response being plotted as a function of the angle that the direction of polarization makes with a given reference direction.

Polarization, receiving (of an antenna). That polarization of a plane wave, incident from a given direction and having a given power flux density, which results in maximum available power at the antenna terminals.

Polarization of a wave (radiated by an antenna in a specified direction). In a specified direction from an antenna and at a point in its far field, the polarization of the (locally) plane wave which is used to represent the radiated wave at that point.

Principal half-power beamwidths. For a pattern whose major lobe has a half-power contour which is essentially elliptical, the half-power beamwidths in the two pattern cuts that contain the major and minor axes of the ellipse respectively.

Radar cross section. For a given scattering object, upon which a plane wave is incident, that portion of the scattering cross section corresponding to a specified polarization component of the scattered wave. *See scattering cross section*.

Radiation efficiency. The ratio of the total power radiated by an antenna to the net power accepted by the antenna from the connected transmitter.

Radiation intensity. In a given direction, the power radiated from an antenna per unit solid angle.

Radiation pattern; antenna pattern. The spatial distribution of a quantity which characterizes the electromagnetic field generated by an antenna.

Note: When the amplitude or relative amplitude of a specified component of the electric field vector is plotted graphically, it is called an *amplitude pattern*, *field pattern*, or *voltage pattern*. When the square of the amplitude or relative amplitude is plotted, it is called a *power pattern*.

Radiation pattern cut. Any path on a surface over which a radiation pattern is obtained.

Note: For far-field patterns the surface is that of the radiation sphere.

For this case the path formed by the locus of points for which θ is a specified constant and Φ is a variable is called a *conical cut*. The path formed by the locus of points for which Φ is a specified constant and θ is a variable is called a *great circle cut*. The conical cut with θ equal to 90° is also a great circle cut. A spiral path which begins at the north pole (θ = 0°) and ends at the south pole (θ = 180°) is called a *spiral cut*.

Radiation resistance. The ratio of the power radiated by an antenna to the square of the rms antenna current referred to a specified point.

Note: The total power radiated is equal to the power accepted by the antenna minus the power dissipated in the antenna.

Realized gain. The gain of an antenna reduced by the losses due to the mismatch of the antenna input impedance to a specified impedance.

Note: The realized gain does not include losses due to polarization mismatch between two antennas in a complete system.

Realized gain, partial (of an antenna for a given polarization). The partial gain of an antenna for a given polarization reduced by the loss due to the mismatch of the antenna input impedance to a specified impedance.

Scattering cross section. For a scattering object and an incident plane wave of a given frequency, polarization, and direction, an area which, when multiplied by the power flux density of the incident wave, would yield sufficient power that could produce by isotropic radiation the same radiation intensity as that in a given direction from the scattering object. *See monostatic cross section*, *bistatic cross section* and *radar cross section*.

Note: The scattering cross section is equal to 4π times the ratio of the radiation intensity of the scattered wave in a specified direction to the power flux density of the incident plane wave.

Sense of polarization. For an elliptical or circularly polarized field vector, the sense of rotation of the extremity of the field vector when its origin is fixed.

Note: When the plane of polarization is viewed from a specified side, if the extremity of the field vector rotates clockwise [counterclockwise], the sense is right-handed [left-handed]. For a plane wave, the plane of polarization shall be viewed looking in the direction of propagation.

Shoulder lobe; vestigial lobe. A radiation lobe which has merged with the major lobe, thus causing the major lobe to have a distortion which is shoulder-like in appearance when displayed graphically.

Sidelobe. A radiation lobe in any direction other than that of the major lobe. *See mean side lobe level* and *minor lobe*.

Squint. A condition in which a specified axis of an antenna, such as the direction of maximum directivity or of a directional null, departs slightly from a specified reference axis.

Notes: (1) Squint is often the undesired result of a defect in the antenna; but, in certain cases, squint is intentionally designed in order to satisfy an operational requirement. (2) The reference axis is often taken to be the mechanically defined axis of the antenna; for example, the axis of a paraboloidal reflector.

Standard [reference] directivity. The maximum directivity from a planar aperture of area *A*, or from a line source of length *L*, when excited with a uniform amplitude, equiphase distribution.

State of polarization; polarization state (of a plane wave [field vector]). At a given point in space, the condition of the polarization of a plane wave [field vector] as described by the axial ratio, tilt angle, and sense of polarization.

Tilt angle (of a polarization ellipse). When the plane of polarization is viewed from a specified side, the angle measured clockwise from a reference line to the major axis of the ellipse.
Note: For a plane wave, the plane of polarization shall be viewed looking in the direction of propagation.

Vestigial lobe. *See shoulder lobe*.

Figure A.1 Pattern properties for antennas