Microwave Trainer
56-200

Feedback

- e-learning
- Electricity & Electronics
- Control & Instrumentation
- Process Control
- Mechatronics
- Telecommunications
- Electrical Power & Machines
- Test & Measurement

Technology Training for tomorrow's world

Microwave Trainer

56-200

Feedback Instruments Limited

Feedback Instruments Ltd, Park Road, Crowborough, E. Sussex, TN6 2QR, UK.
Telephone: +44 (0) 1892 653322, Fax: +44 (0) 1892 663719.
email: feedback@fdbk.co.uk website: http://www.fbk.com

Manual: 56-200 Ed05 042005 *Printed in England by FI Ltd, Crowborough*

Feedback Part No. 1160–56200

Notes

THE HEALTH AND SAFETY AT WORK ACT 1974

We are required under the Health and Safety at Work Act 1974, to make available to users of this equipment certain information regarding its safe use.+

The equipment, when used in normal or prescribed applications within the parameters set for its mechanical and electrical performance, should not cause any danger or hazard to health or safety if normal engineering practices are observed and they are used in accordance with the instructions supplied.

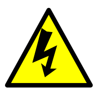
If, in specific cases, circumstances exist in which a potential hazard may be brought about by careless or improper use, these will be pointed out and the necessary precautions emphasised.

While we provide the fullest possible user information relating to the proper use of this equipment, if there is any doubt whatsoever about any aspect, the user should contact the Product Safety Officer at Feedback Instruments Limited, Crowborough.

This equipment should not be used by inexperienced users unless they are under supervision.

We are required by European Directives to indicate on our equipment panels certain areas and warnings that require attention by the user. These have been indicated in the specified way by yellow labels with black printing, the meaning of any labels that may be fixed to the instrument are shown below:

CAUTION -
RISK OF
DANGER



CAUTION -
RISK OF
ELECTRIC SHOCK

CAUTION -
ELECTROSTATIC
SENSITIVE DEVICE

Refer to accompanying documents

PRODUCT IMPROVEMENTS

We maintain a policy of continuous product improvement by incorporating the latest developments and components into our equipment, even up to the time of dispatch.

All major changes are incorporated into up-dated editions of our manuals and this manual was believed to be correct at the time of printing. However, some product changes which do not affect the instructional capability of the equipment, may not be included until it is necessary to incorporate other significant changes.

COMPONENT REPLACEMENT

Where components are of a 'Safety Critical' nature, i.e. all components involved with the supply or carrying of voltages at supply potential or higher, these must be replaced with components of equal international safety approval in order to maintain full equipment safety.

In order to maintain compliance with international directives, all replacement components should be identical to those originally supplied.

Any component may be ordered direct from Feedback or its agents by quoting the following information:

1. Equipment type	2. Component value
3. Component reference	4. Equipment serial number

Components can often be replaced by alternatives available locally, however we cannot therefore guarantee continued performance either to published specification or compliance with international standards.

OPERATING CONDITIONS

WARNING:

This equipment must not be used in conditions of condensing humidity.

This equipment is designed to operate under the following conditions:

Operating Temperature	10°C to 40°C (50°F to 104°F)
Humidity	10% to 90% (non-condensing)

CE DECLARATION CONCERNING ELECTROMAGNETIC COMPATIBILITY

Should this equipment be used outside the classroom, laboratory study area or similar such place for which it is designed and sold then Feedback Instruments Ltd hereby states that conformity with the protection requirements of the European Community Electromagnetic Compatibility Directive (89/336/EEC) may be invalidated and could lead to prosecution.

This equipment, when operated in accordance with the supplied documentation, does not cause electromagnetic disturbance outside its immediate electromagnetic environment.

ACKNOWLEDGEMENTS

Feedback Instruments Ltd acknowledge all trademarks.

IBM, IBM - PC are registered trademarks of International Business Machines.

MICROSOFT, WINDOWS XP, WINDOWS 2000, WINDOWS ME, WINDOWS NT, WINDOWS 98, WINDOWS 95 and Internet Explorer are registered trademarks of Microsoft Corporation.

TABLE OF CONTENTS

Notes

1 The microwave trainer, system and components

1.1 Introduction

The Feedback 56-200 is a microwave trainer which enables students to familiarise themselves with the basics waveguide components making up microwave systems. It is an easy-to-use bench-top system, which is completely self-contained. It allows students to undertake microwave measurements and investigate the working of components making up the radio frequency transmission sections of microwave radio transmitters and receivers and radar systems.

The trainer and manual provide the means for students to gain a working knowledge of microwaves and their applications and carry out basic microwave systems practical studies. The manual is written in a straightforward, clear style, and the trainer is suitable for technician and degree-level training and study.

1.2 Equipment

The trainer comprises a set of waveguide components and a console that contains the microwave source power supply and circuits associated with measuring microwave power and signal strength. The components and source are designed in standard waveguide size WG16 that operates in the X-band range 8.2 to 12.4 GHz, a very important band for microwave radio, satellite communications and radar applications. The source itself operates at a fixed frequency of 10.425 GHz.

The components shown in their packing-case positions are illustrated in Figure 1.1 and the complete list is given in paragraph 1.3.

1.3 System Components

Quantity	Identity No.	Description
		System waveguide: WG16 (WR90) which has the following specification: Internal dimensions: 0.9" x 0.4" or 22.86 x 10.16 mm H_{10} cut-off frequency and wavelength: 6.56 GHz, 45.7 mm Normal operating range: 8.2 to 12.4 GHz
1	P	X-band CW FET Dielectric Resonant Oscillator (DRO) Source Frequency: fixed 10.425 GHz Output power: 10 mW minimum
2	A	Variable attenuators, resistive vane-central slot type; used to set attenuation level and control power transmission in waveguides. Maximum attenuation at vane setting 0° approx. 36 dB; minimum at 90° less than 1 dB
1	B	Waveguide slotted line, for sampling electric field pattern in waveguide; used with diode-probe detector to measure guide wavelength, vswr and impedance
1	C	Slotted line-probe tuner used as an impedance matching device
1	D	Cavity wavemeter. Circular cross-section cavity wavemeter which resonates in the E_{011} mode and designed to measure frequency in the X-band range
1	E	H-plane or shunt Tee waveguide junction; acts as a power divider in the plane containing the incident H (magnetic field)
1	F	Direction coupler: side-wall coupler type with directional coupling property, used to monitor power and measure vswr
1	G	E-plane or series Tee waveguide junction; acts as a power divider in the plane containing the incident E (electric field)

Quantity	Identity No.	Description
1	H	Hybrid Tee, also known as a 'magic Tee', is a superposition of a shunt and a series Tee junction to form a 4-way junction; used to effect common transmitter/receiver antenna operation and in balanced mixer circuits
2	J	Waveguide-to-coaxial transformer, used to interconnect waveguide to coaxial line and vice-versa
1	K	Resistive termination, a waveguide section containing a taper of lossy material to absorb incident microwave signals; ideally should absorb totally incoming signals without any reflection - it then acts as a matched load
1	M	Diode detector in waveguide mount, used to rectify microwave signals for their detection; at low-power levels diode detector output current is directly proportional to the microwave power being detected.
2	N	Horn antenna, an important microwave antenna widely used as a feed to microwave parabolic reflectors in radio, satellite and radar systems, and also as an antenna in its own right
2	R	Short-circuit termination, metal plates used to short-circuit waveguide section; employed in impedance measurements to determine reference planes, also used to measure guide wavelength and crystal detector law in conjunction with slotted line
2	S	Probe detector, diode detector mounted in coaxial section with inner conductor acting as a probe; used in conjunction with slotted line and directional coupler to detect microwave signals [n.b. Letter S not marked on unit]
1		Manual 56-200

Quantity	Identity No.	Description
1		Length of coaxial cable (approx. length 380 mm and external diameter 10 mm) with N-type connectors
		Accessories, contained in plastic bags: 2 coaxial cable (approx 3 mm external diameter) with BNC connectors for detector connection to console 24 flange coupling clips for interconnecting waveguide components 4 support plates

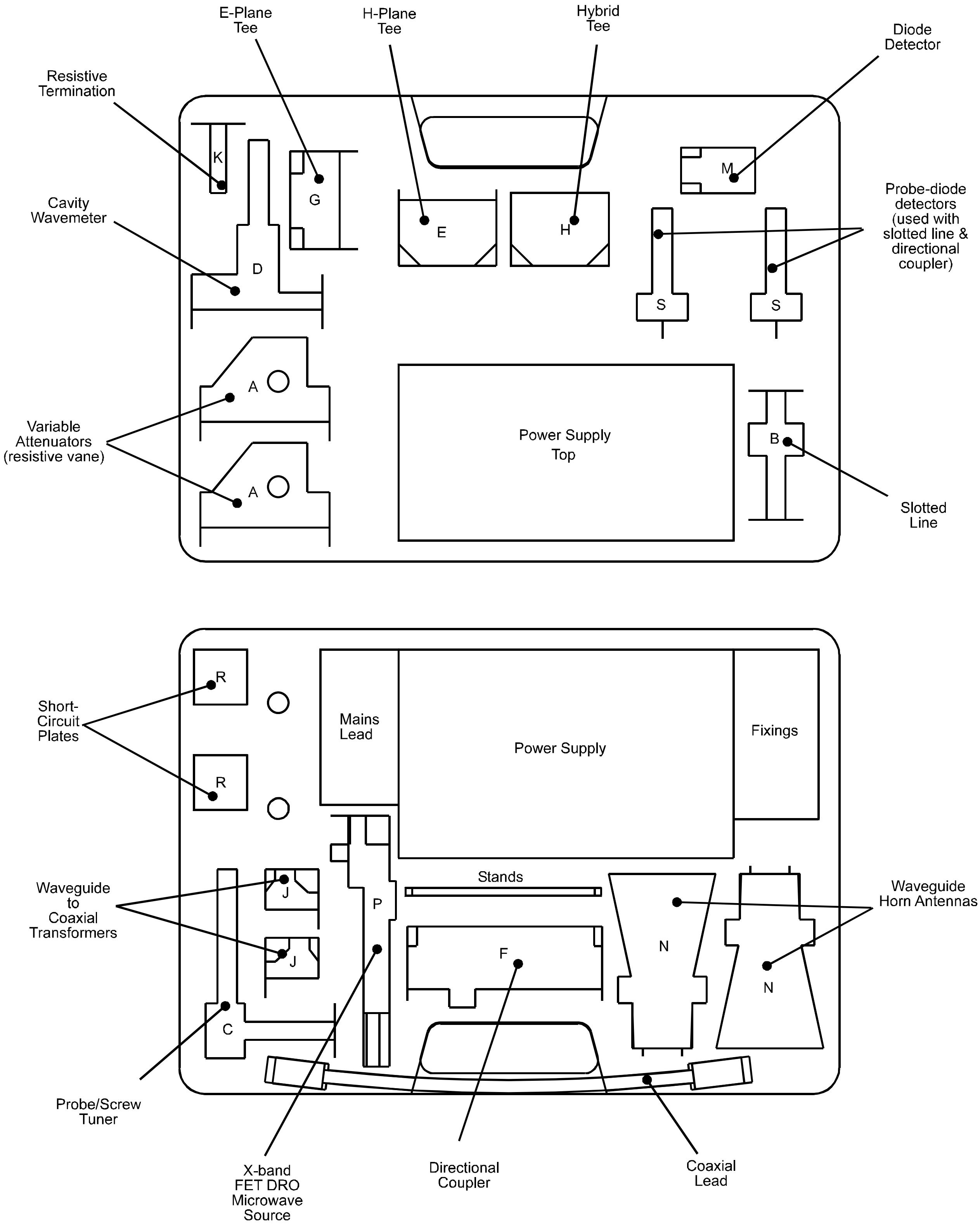


Figure 1-1: Systems Components – Case Content

Notes

1 ASSIGNMENTS

1.1 Introduction

This chapter contains details of the practical assignments which can be undertaken using the microwave waveguide trainer.

All assignments are essentially free-standing. They contain the information necessary for each assignments to be carried out on an individual basis. The numerical order follows a logical development, but assignments can be omitted or their order changed without loss in continuity or understanding.

All assignments are essentially free-standing. They contain the information necessary for each assignments to be carried out on an individual basis. The numerical order follows a logical development, but assignments can be omitted or their order changed without loss in continuity or understanding.

1.2 List of Assignments

1 Introduction of a microwave waveguide bench and measurement of source frequency and wavelength.

2 Measurement of voltage and standing wave ratio (VSWR).

3 Measurement of diode detector law.

4 Measurement of impedance and impedance matching.

5 Horn antenna investigations.

6 Use of a directional coupler in power transmission and reflected measurements.

7 Series, shunt and hybrid T junctions.

8 Waveguide to coaxial transformers.

9 Microwave radio link investigations.

Notes

Assignment 1
Introduction of a Microwave Waveguide Bench and Measurement of Source Frequency and Wavelength

CONTENT

In this first assignment a basic microwave measurement bench is set up to measure frequency and guide wavelength.

The waveguide bench comprises a FET Dielectric Resonant Oscillator (DRO) source, two resistive vane attenuators, a cavity wavemeter, a waveguide slotted line, diode detectors and a resistive load termination.

The frequency of the microwave source is measured using the cavity wavemeter. The guide wavelength l_g is the wavelength of the microwave signal propagating in the waveguide. This is measured using a diode detector-probe unit to sample the standing waves set up in the slotted line when terminated in a short-circuit.

EQUIPMENT REQUIRED

Quantity	Identifying letter	Component description
1	---	Control console
2	A	Variable attenuators
1	B	Slotted line
1	D	Cavity wavemeter
1	K	Resistive termination
1	M	Diode detector
1	P	X-band microwave source
1	S	Probe detector assembly
1	R	Short-circuit plate

OBJECTIVES

When you have completed this assignment you should

- Be familiar with some basic microwave waveguide components and know their use
- Know how to measure frequency using a cavity wavemeter
- Know how guide wavelength λ_g is measured using a slotted line
- Understand the meaning of cut-off wavelength and frequency
- Use the general relationship for waveguides of:

$$\frac{1}{\lambda_g^2} = \frac{1}{\lambda^2} - \frac{1}{\lambda_c^2}$$

to calculate guide wavelength, cut-off wavelength and free space wavelength and frequency.

KNOWLEDGE LEVEL

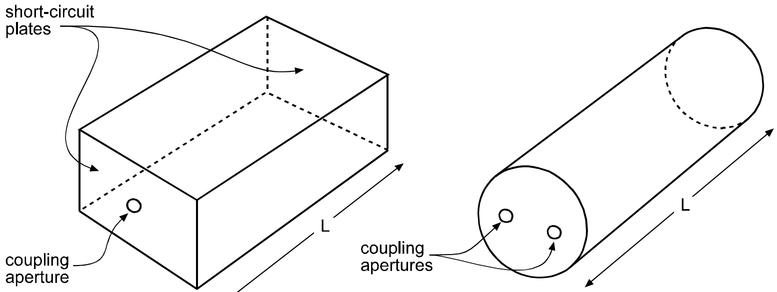
No prior specialist knowledge is required to carry out this assignment. Basic concepts of waves, wavelength and frequency should be known. You should also be familiar with reading a micrometer. To assist in understanding the measurements taken in the assignment it would also be useful to appreciate

- The nature of electromagnetic waves as being composed of oscillating electric and magnetic fields
- The action of a diode in rectifying alternating current and in microwaves in detecting microwave signals for measurement of field strength and power
- That waves in a closed environment can resonate, e.g. in a cavity when the cavity length is a whole number of half-wavelengths
- Certain types of waves, known as modes, can exist in waveguide structures and are characterised by their own particular wave pattern of electric and magnetic field

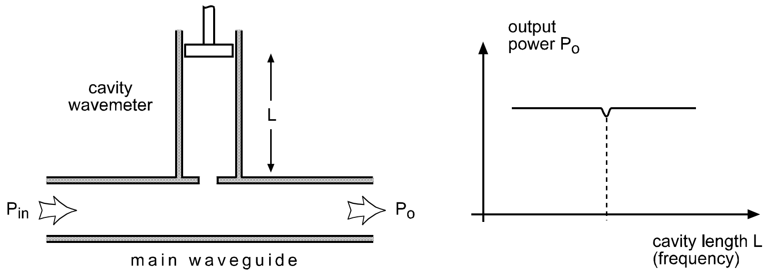
INTRODUCTION

(a) Measurement of source frequency using a cavity wavemeter

Frequency in a microwave system can be measured using electronic counter techniques or by means of a cavity wavemeter.



(a)
Rectangular and circular waveguide cavities: resonance when L = 1/2nλg



(b)
Cavity wavemeter as an absorption wavemeter to measure frequency: output power shows a dip at resonance L = 1/2nλg

Figure 2-1-1: Cavity Wavemeters

The principle of operation of the cavity wavemeter is based on the fact that very high Q-resonances can be obtained in metal waveguide cavities. Such cavities are usually of uniform circular or rectangular cross-section and resonate when their axial length equals an integral number of half guide wavelength, i.e. with reference to figure 2.1.1(a) when:

$$L = \frac{1}{2} n \, \lambda_g$$

where L = axial length of cavity

n = 1,2,3...., the order of resonance

λ_g = guide wavelength of resonating mode

Figure 2-1-1(b) illustrates a practical way of using a cavity as an absorption-type wavemeter. The cavity length L may be varied by altering the position of the short-circuit plunger. Off resonance the cavity absorbs little or no power from the main waveguide transmission system. However, at resonance considerable power is coupled into the cavity and this results in a corresponding dip observed in the main transmitted power. L at resonance can be very accurately determined. Knowing L, the type of resonant mode and the order of resonance enables the exciting frequency, the source frequency f, to be calculated. From theory:

$$f = \frac{c}{l} = 3\,x\,10^8 \sqrt{\left[\frac{1}{l_g^2} + \frac{1}{l_c^2}\right]}$$

$$= 3\,x\,10^8 \sqrt{\left[\left(\frac{n}{2L}\right)^2 + \frac{1}{l_c^2}\right]}$$

where

c = 3×10^8 m/s, the velocity of electromagnetic waves in free space

l_c = cut-off wavelength of mode resonant in the cavity

n = order of resonance: n = 1, 2, 3.....

More usually a calibration curve of frequency f versus L is provided. In high-quality wavemeters a cylindrical spiral scale measuring plunger position is calibrated to read frequency direct as illustrated in Figure 2-1-2.

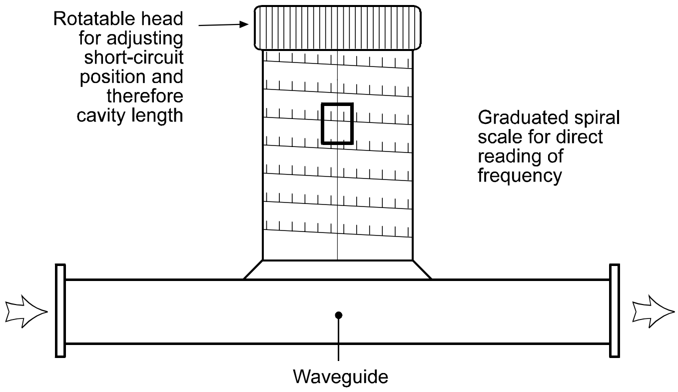


Figure 2-1-2: Cavity Wavemeter Calibrated for Direct Reading Frequency

A diagram of the wavemeter used in the microwave trainer is sketched in Figure 2-1-3. The cavity consists of circular waveguide of diameter D = 28.3 mm. Its length can be adjusted to be a maximum of approximately 22 mm. Over the X-band frequency range the cavity can only support two modes, the H_{11} and the E_{01}, whose cut-off wavelengths are given respectively by:

$$H_{11} : l_c = 1.71\ D\ ;\ E_{01} : l_c = 1.31\ D$$

so for the case of the wavemeter where D = 28.3 the values of cut-off wavelength and frequency are:

$$H_{11} : l_c = 48.4\ \text{mm}\ ,\ f_c = 6.2\ \text{GHz}$$

$$E_{01} : l_c = 37.1\ \text{mm}\ ,\ f_c = 8.1\ \text{GHz}$$

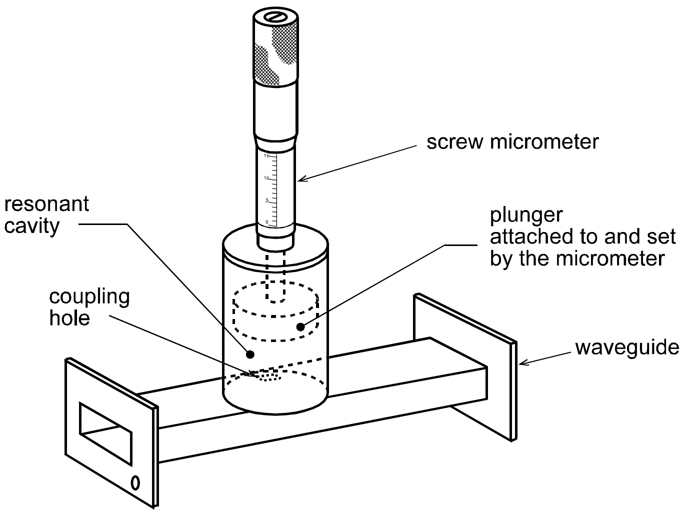


Figure 2-1-3: Cavity Wavemeter used in Microwave Trainer

A graph of resonant frequency versus micrometer reading for the wavemeter is plotted in Figure 2-1-4 for the first order resonant mode E_{011}. The wavemeter used in the trainer is in fact designed to operate in the E_{011} mode. Thus if a dip is found in the transmitted power, see Figure 2-1-1(b), when the wavemeter micrometer is, for example, at 11.2 mm the frequency can be measured using the curve of Figure 2-1-4 and in this case would correspond to f = 10.4 GHz.

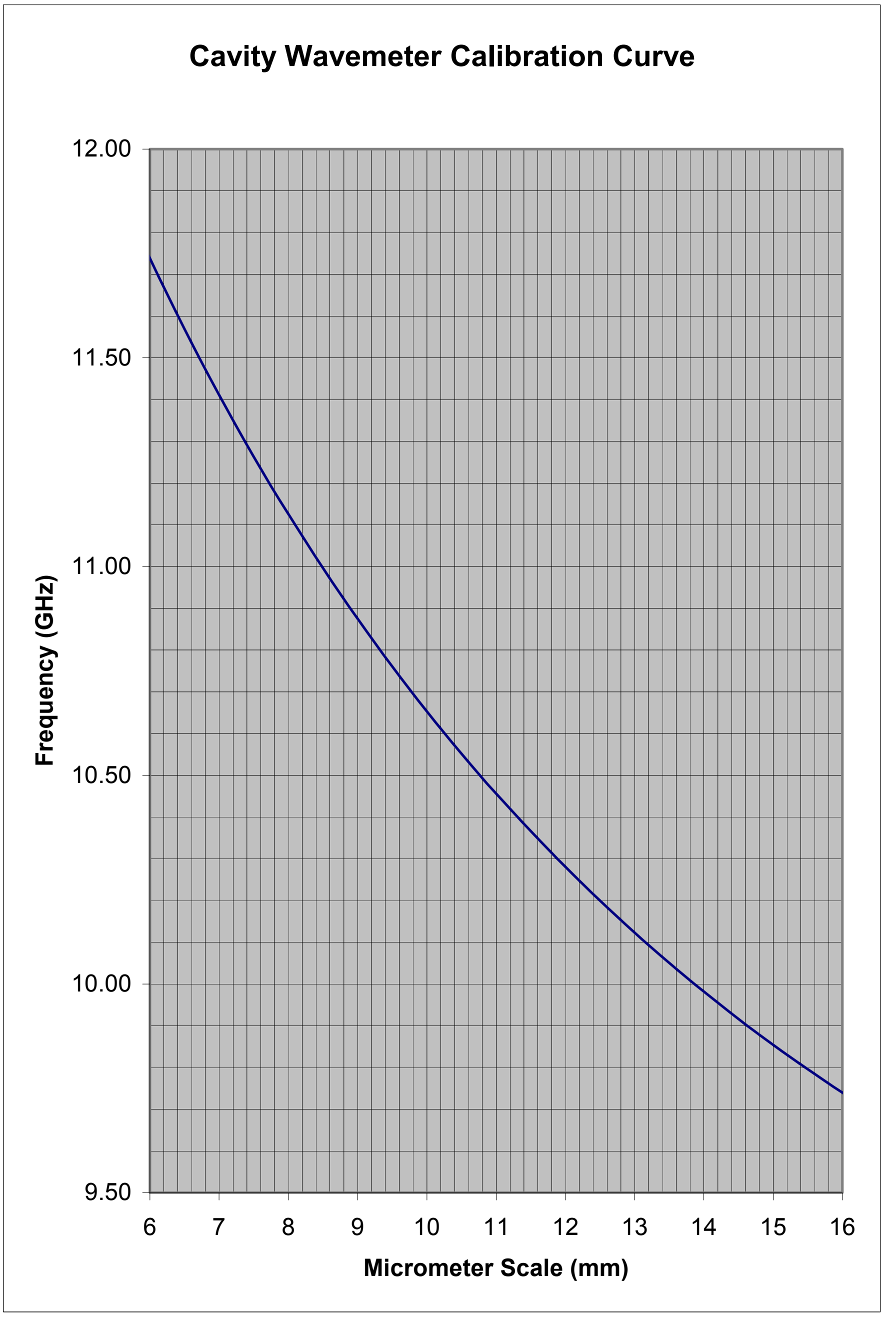


Figure 2-1-4: Calibration Curves for Cavity Wavemeter

Diameter D = 28.3 mm.
Note 0.00 point of micrometer scale
corresponds to cavity length of 11.63 mm and this offset
is taken into account in the above graph.

(b) Guide wavelength and its measurement

Free space wavelength l is the distance travelled by the wavefront of the electromagnetic wave in free space in the duration of one cycle. It is related to frequency f by:

$$f\lambda = c \quad , \quad c = 3 \times 10^{8}\,m/s$$

$$\lambda = \frac{c}{f}$$

When the waves are guided by a waveguide they travel in the form of distinctive wave patterns known as modes and the wavelength of the guided transmission is known as the guide wavelength λ_g. For rectangular and circular waveguides, λ_g is related to l by the formulae:

$$\frac{1}{\lambda_g^2} = \frac{1}{\lambda^2} = \frac{1}{\lambda_c^2} \qquad (2)$$

$$\lambda_g = \frac{\lambda\lambda_c}{\sqrt{\left(\lambda_c^2 - \lambda_0^2\right)}} \qquad (3)$$

where λ_c = the cut-off wavelength of the mode propagating .

For the case of rectangular waveguides, transmission is invariably limited to single-moded operation in its dominant H_{10} mode. The cut-off wavelength for the H_{10} mode is:

$$\lambda_c = 2a$$

where a = internal broadside dimension of the waveguide. Thus if f is known, λ can be determined and λ_g can be calculated for a given size of waveguide using formula 3.

The guide wavelength can be measured experimentally using the slotted waveguide section and probe-detector components shown diagrammatically in Figure 2-1-5. By terminating the slotted section in a short circuit a standing wave pattern can be set up. The incident wave from the source and the reflected wave from the short-circuit combine to give a resultant standing wave in the section whose electric field amplitude varies as shown in Figure 2-1-6 The distance between successive nulls is in the standing wave ½λ_g and can be measured very accurately. Thus guide wavelength can be determined experimentally to a high degree of accuracy.

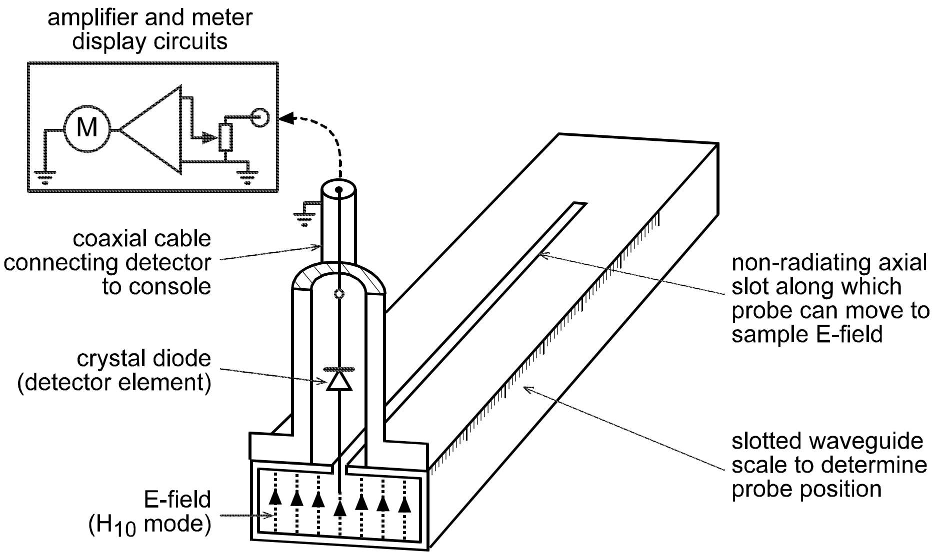


Figure 2-1-5

Diagrammatic sketch indicating main components of a waveguide slotted line and diode-probe detector for investigating standing waves in rectangular waveguide.

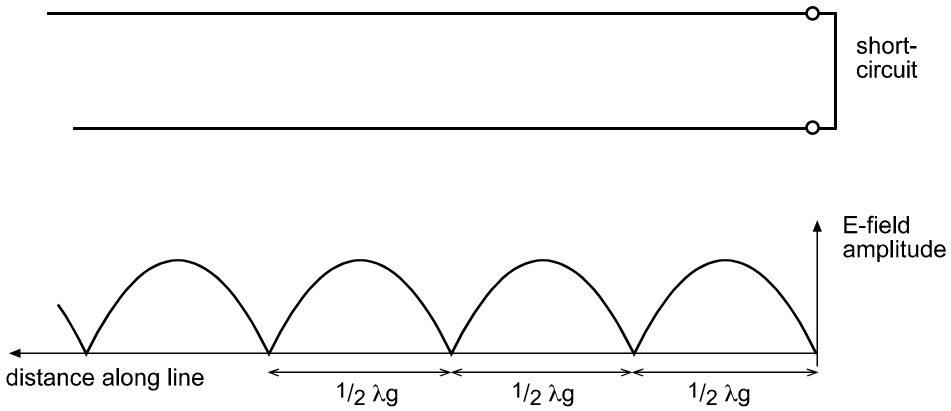


Figure 2-1-6

Standing wave electric field pattern on a short-circuited line; distance between successive null =$^1/_2\ \lambda g$

EXPERIMENT PROCEDURE

WARNING:

Although the microwave power levels generated by the equipment are below 10 mW and not normally dangerous, the human eye can suffer damage exposed to direct microwave radiation. Therefore:

NEVER look directly into an energised waveguide.

(a) Measurement of frequency using a cavity wavemeter

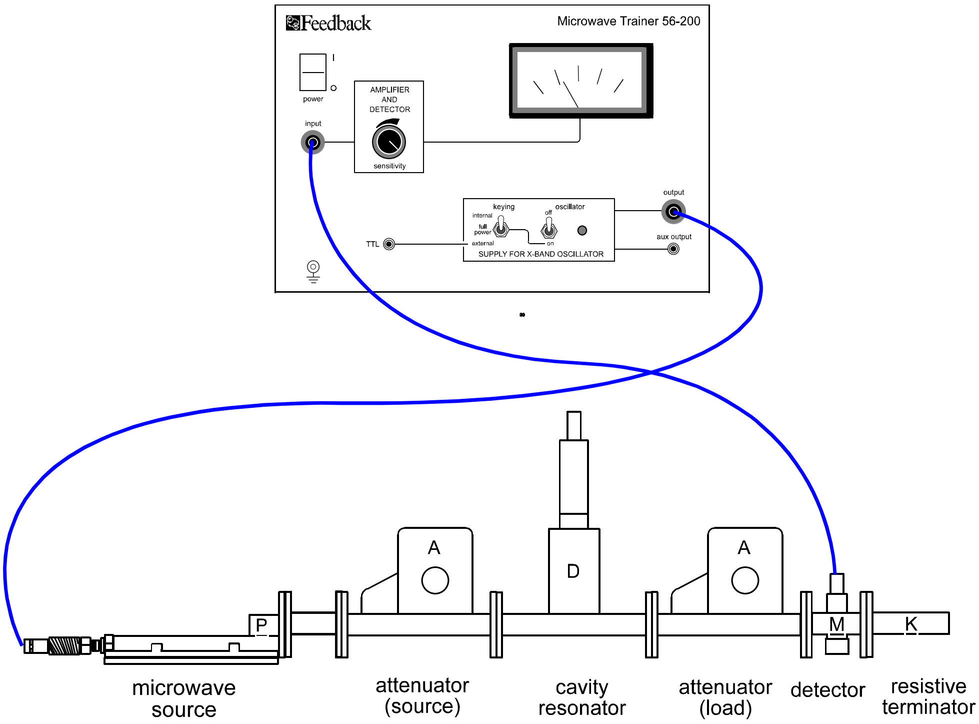


Figure 2-1-7

1. Set up the microwave components as shown in Figure 2.1.7 with the switch positions on the control console initially as follows:

   main green power switch : off;

   amplifier and detector sensitivity control knob : to mid-position;

supply for X-band oscillator :

left-hand switch : switch to internal keying;

right-hand switch : off

2. Make sure the coaxial cables are also in place to connect the power supply for the X-band oscillator and to connect the diode detector output to the amplifier and detector input on the console.

3. Set the micrometer position of the cavity wavemeter fully out to a reading greater than 15 mm. In this position, the short-circuit plunger terminating the far end of the cavity is also fully out and the cavity is at its maximum length.

4. Set the angular position of the resistive vane for both attenuators at about 20°. At these settings, the attenuators reduce the microwave power in the bench by the order of 10 dB and avoid possible diode-detector and display meter overload when we switch on.

5. Now switch on the main green power switch and the right-hand supply switch for the X-band oscillator to energise the bench.

6. Adjust the attenuator adjacent to the diode detector to give a meter reading on the console meter of about 4 mA. Note increasing the vane penetration reduces the microwave power transmitted in the system.

7. Determination of frequency using the wavemeter

Turn the micrometer thimble of the wavemeter very slowly clockwise to move the short-circuit plunger downwards and thus reduce the length of the cavity. Observe the meter deflection whilst this is being done.

Search for a position at which there is a sharp dip in the meter reading. Such a dip corresponds to a resonance at which power is absorbed by the wavemeter and so reduces the transmitted microwave power as detected by the diode-detector and observed on the meter.

Record the micrometer reading at this resonance and determine the frequency of the microwave signals using the E_{011} mode calibration curve of Figure 2-1-4.

NOTE

(i) The linear scale on the micrometer barrel of the wavemeter is graduated in 0.5 mm intervals. The circular scale on the thimble is graduated 0 to 50 with each graduation representing a 0.01 mm movement. Thus the vertical movement of the short-circuit and hence the cavity length and resonance position can be measured extremely accurately.

(ii) The design of the wavemeter is such that it resonates principally in the E_{011} mode and hence the curve for this mode should be used in Figure 2-1-4.

(iii) At micrometer readings of the order of 5 mm and below a number of deep over-lapping resonances may also be observed. These are termed "spurious" and should be ignored. They arise principally to leakage of energy past the plunger.

(b) Measurement of guide wavelength, λ_g

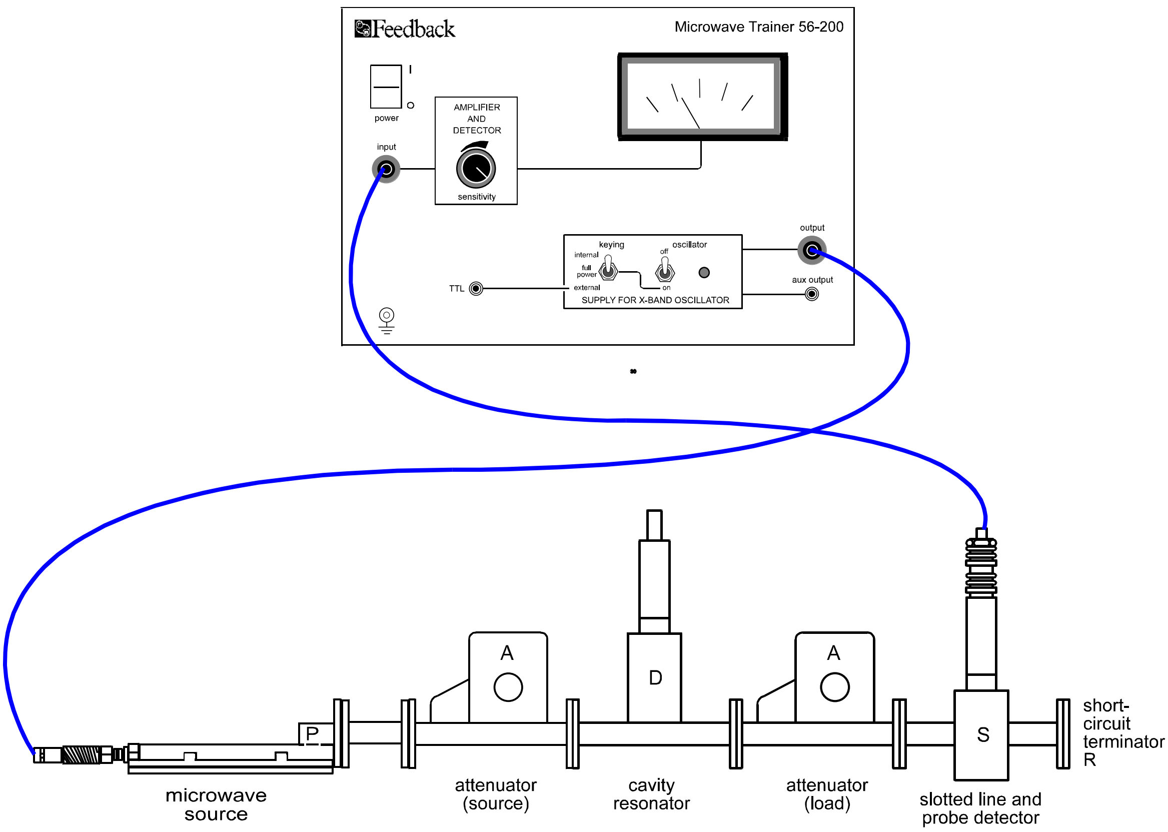


Figure 2-1-8

1. The guide wavelength is measured using the waveguide slotted line component B fitted with one of the diode-detector/probe units S. This unit should be mounted so that its probe penetrates a short distance into the slot thereby allowing the electric field to be sampled. The depth of penetration is a compromise between obtaining reasonable coupling for the probe-detector to provide an observable meter reading without the probe causing undue disturbance of the field in the waveguide and thus invalidate the measurements. In practice a penetration depth of 1 to 2 mm should suffice.

2. Connect the equipment as shown in Figure 2-1-8 with the slotted waveguide section terminated in the metal plate R acting as a short-circuit.

3. Set the switch positions on the control console as follows:

   green power switch : off;

   left-hand switch of supply for X-band oscillator : to internal keying;

   right-hand oscillator switch : off.

   Check also the coaxial cables are correctly connected :

   oscillator output cable on console to microwave source;

   probe-detector unit cable to input of amplifier-detector on the console.

4. Now switch on power and oscillator and adjust attenuators and if necessary the sensitivity control on the console to obtain a detector reading. Move the probe to locate a position of maximum field and re-adjust sensitivity control and/or attenuators to provide a meter reading close to full scale, say 4 mA.

5. Starting from zero on the slotted-line scale move the probe along the slotted waveguide section and locate and record the positions of electric field nulls. It should be possible to locate 3 consecutive nulls:

   x_1 =

   x_2 =

   x_3 =

The guide wavelength:

$\lambda_g = 2\ (x_2 - x_1) =$

$= 2\ (x_3 - x_2)\ =$

$= (x_3 - x_1)\ \ \ =$

6. The waveguide used in the Microwave Trainer is standard WG16 whose internal dimensions are

broad dimension a = 0.9 inch = 22.86 mm

narrow dimension b = 0.4 inch = 10.16 mm

The cut-off wavelength for the dominant mode, the H_{10} mode is

$\lambda_c = 2a$

$= 2 \times 22.86$ mm $= 45.72$ mm for WG 16

Using the result of formula (3) and the above value of λ_c determine the guide wavelength λ_g at the source frequency, f = 10.7 GHz. Compare with the experimentally determined value.

SUMMARY

In this assignment two important parameters have been measured and a number of basic microwave components have been used. Frequency has been measured using a cavity wavemeter and guide wavelength has been measured experimentally using a waveguide slotted line. The relationship between free space wavelength l has also been introduced and used to determine guide wavelength.

Notes

(a) Measurement of frequency using the cavity wavemeter

Starting with the micrometer withdrawn to maximum length a sharp resonance is observed at a micrometer reading of 11.17 mm.

Using calibration curve of Figure 2-1-4 for the E_{011} mode frequency
f = 10.42 GHz.

Alternatively using the formula:

1. $$f = c\left(\frac{1}{4L^2} + \frac{1}{\lambda_c^2}\right)^{\frac{1}{2}}$$

with $c = 3 \times 10^8$ m/s

$L = 11.17 + 11.63$ mm = 22.8 mm

(11.63 mm to correct micrometer scale to read total cavity length)

$\lambda_c = 37.1$ mm for the E_{01} mode

we have

$$f = 3 \times 10^8 \left(1/4 \times 0.023^2 + 1/0.0371^2\right)^{½}$$

$$= 3 \times 10^8 (484 + 727)^{½} = 10.42 \text{ GHz}$$

Note:

At much lower micrometer readings, typically below 5 mm, a number of deep overlapping resonances may be observed. These carry no useful information and should be ignored; they arise due to limitations in the wavemeter design, notable leakage around the short-circuit plunger.

(b) Measurement of guide wavelength, l_g

With the slotted line section terminated in a short-circuit nulls are observed at scale positions

x_1 = 27 mm x_2 = 45.4 mm x_3 = 64 mm

so $X_2 - x_1$ = 18.4 mm , $l_g = 2 \times 18.4 = 36.8$ mm

$x_3 - x_2$ = 18.6 mm , $l_g = 2 \times 18.6 = 37.2$ mm

$x_3 - x_1$ = 37 mm , l_g = 37 mm

Average = 37 mm

Check using formula (3):

2. $$\lambda_g = \frac{\lambda \ \lambda_c}{\sqrt{(\lambda_c^2 - \lambda^2)}}$$

The waveguide size of the guide used in the trainer is WG16 with

a = 0.9 inch = 22.86 mm

so for the H_{10} mode,

λ = 2a = 45.72 mm

and at the measured frequency of f = 10.42 GHz,

3. $$\lambda = \frac{c}{f} = \frac{3 \times 10^8}{10.42 \times 10^9} = 28.76 \text{ mm}$$

we have:

4. $$l_g = \frac{28.76 \times 45.7}{\sqrt{(45.7^2 - 28.76^2)}} = 37 \text{ mm}$$

showing very good agreement with the measured value.

CONTENT

In this assignment, the measurement of voltage standing wave ratio (VSWR) of waveguide components is undertaken using a waveguide slotted-line and probe detector. Voltage standing wave ratio, invariably abbreviated to VSWR, is one of the fundamental parameters used in specifying component performance. It quantifies the degree of mismatch a component presents to the waveguide feed line.

EQUIPMENT REQUIRED

Quantity	Identifying letter	Component description
1	---	Control console
2	A	Variable attenuators
1	B	Slotted line
1	S	Probe-diode detector
1	P	X-band source
1	K	Resistive termination
1	N	Waveguide horn

OBJECTIVES

When you have completed this assignment you should:

- Know the definition of voltage standing wave ratio and its relation to reflection coefficient
- Know how to measure VSWR using a slotted line and probe-detector
- Know the method used to measure high values of VSWR

KNOWLEDGE LEVEL

No prior specialist knowledge is required to carry out this experiment, although completion of Assignment 1 would be a distinct advantage.

INTRODUCTION

When a component is connected into a transmission line system it will cause reflection at the junction between the line and the component unless it is correctly matched or a matching unit is used.

The reflected wave from the component and incident wave from the source set up a standing wave pattern in the feed line as illustrated in Figure 2-2-1

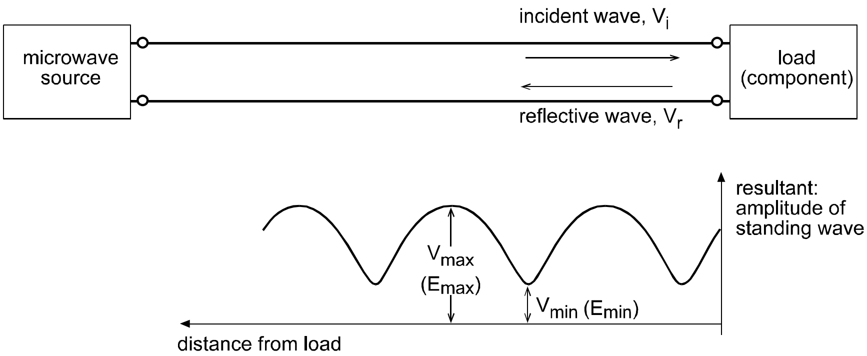


Figure 2-2-1: Standing Wave Pattern: VSWR S = V_{max}/V_{min} or E_{max}/E_{min}

Standing waves cause undesirable effects. They can give rise to very high values of voltage/electric field strength in the waveguide and this can cause breakdown in high-power systems. Reflection caused through the mismatch between the line and component reduces the amount of power that can be transferred below optimum and can adversely affect the efficiency of a communications or radar system. Power can be wasted in a transmitter by being reflected, for example, at the antenna input. Likewise, power can be lost by mismatch reflection at an antenna-receiver station.

The voltage standing ratio, vswr, is universally used to quantify the degree of mismatch. It is denoted by the letter S and defined as:

$$S = \frac{V_{max}}{V_{min}} = \frac{E_{max}}{E_{min}}$$

where V_{max}, E_{max} = voltage or electric field strength at a position of field maximum.

V_{min}, E_{min} = voltage or electric field strength at a position of field minimum.

There is an important relation between vswr, S, and reflection coefficient G:

$G = V_r/V_i$

where V_i = incident wave voltage or electric field amplitude

V_r = reflected wave amplitude

and since $V_{max} = V_i + V_r = V_i\,(1 + G)$

$V_{min} = V_i - V_r = V_i\,(1 - G)$

$$S = \frac{V_{max}}{V_{min}} = \frac{1 + \Gamma}{1 - \Gamma}$$

$$\Gamma = \frac{S - 1}{S + 1}$$

The table below provides a useful guide to the relationships of VSWR, reflection coefficient G and power reflected due to mismatch.

vswr S	Reflection coefficient Γ	% power reflected Γ^2%	Comment
1.0	0	0	System matched
1.05	0.048	0.22	Very good match
1.5	0.2	4.0	Fair, just acceptable
2.0	0.33	11.1	Poor match, not usually acceptable
5.0	0.67	44.4	Reject, faulty, unacceptable

The voltage standing wave ratio produced by a waveguide component may be measured using a slotted waveguide section and a diode detector probe. The slotted section is inserted in the waveguide system to provide a means of sampling the standing wave electric field pattern produced by reflection from the component under consideration. The slotted section consists of waveguide with a narrow axial slot cut in the centre of its broad face as indicated in Figure 2-2-2(a). A narrow slot in this position is non-radiating and causes negligible distortion to the waves within the waveguide. Coupling to the electric field is made by a probe which penetrates a small distance through the slot as illustrated in Figure 2-2-2(b). The probe is connected in series with a crystal diode detector and the unit contained in a carriage that may be moved along the slot so enabling the field at different axial positions to be measured. The diode detector rectifies the sampled microwave signal and the rectified current is measured on a DC milliammeter.

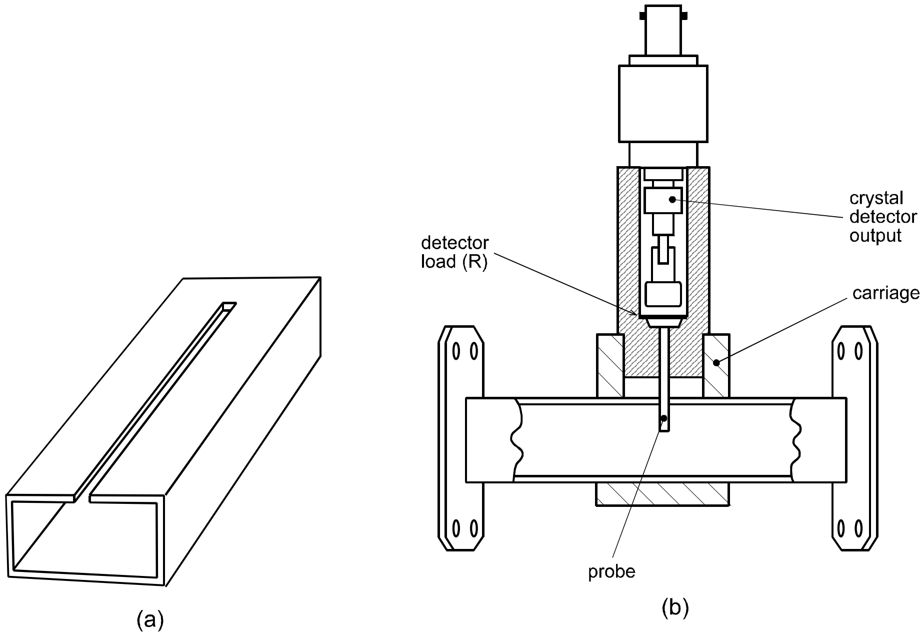


Figure 2-2-2: Slotted Waveguide and Diode Detector Probe Unit for Measuring Voltage Standing Wave Ratio

For small currents the diode-detector obeys a square law such that the detected current is proportional to the square of the electric field induced in the probe. Thus, if the values of maximum and

$$VSWR,\ S = \frac{E_{max}}{E_{min}}$$

minimum current are measured, we have:

$$\text{so } S = \frac{E_{max}}{E_{min}} = \frac{\sqrt{I_{max}}}{\sqrt{I_{min}}} = \sqrt{\frac{I_{max}}{I_{min}}}$$

where I_{max} = current measured at field maximum E^2_{max}

where I_{max} = current measured at field maximum E^2_{min}

for small current conditions

e.g. if I_{max} is measured at 4.6 mA and I_{min} at 3.2 mA

$$S = \frac{E_{max}}{E_{MIN}} = \sqrt{\frac{4.6}{3.2}} = \sqrt{1.44} = 1.2$$

Voltage standing wave ratio can be measured without knowledge of the diode-detector characteristics by using a precision attenuator. The probe is moved to a position of minimum and the value of the attenuator attenuation A_1 dB noted which gives a convenient current reading on the meter for reference. The probe is then moved to locate a position of maximum and the attenuator adjusted to provide the same reading as obtained with the minimum. Suppose the new attenuation value is to be A_2, then:

$$\begin{aligned} \text{attenuation added} &= A_2 - A_1 \\ &= 10 \log (E^2_{max}/E^2_{min}) \\ &= 20 \log (E_{max}/E_{min}) \\ &= 20 \log S \end{aligned}$$

so S =antilog$_{10}$ $(A_2 - A_1)/20$ or $10^{(A2 - A1)/20}$

For large values of vswr, typically S > 3, E_{min} and therefore the detector current will become very small and the ratio E_{max} to E_{min} increasingly difficult to measure accurately. For these cases a method based on locating the minimum and determining the distance between points either side of the minimum at which the field is a constant factor, k say, times the minimum value. With reference to Figure 2-2-3 and assuming a square law for the diode detector, the VSWR may be evaluated from the formula

$$S = \frac{\sqrt{\left[k^2 - \cos^2 (\pi d / \lambda g) \right]}}{\sin \left(\frac{\pi d}{\lambda g} \right)}$$

where d = distance between points where

electric field, $E = k\, E_{min}$

detector current, $I = k^2\, I_{min}$

l_g = guide wavelength

$k^2 = I/I_{min}$

It is common practise to choose $k^2 = 2$ so

d = distance between points where detector current equals 2 I_{min}

and in this case the formula for S reduces to

$$S = \sqrt{1 + \frac{1}{\sin^2\left(\frac{\pi d}{\lambda g}\right)}}$$

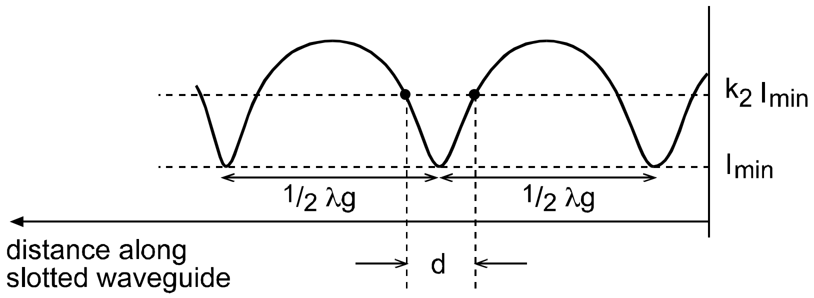


Figure 2-2-3: Plotting around the Field Minimum Method for Measuring High Values of VSWR

e.g. if d = 1.5 mm and l_g = 35.5 mm

pd/lg = 0.133 rad or 7.61°

so:

$$S = \sqrt{\left[1 + \frac{1}{\sin^2 7.61}\right]} = \sqrt{[1 + 57.1]} = 7.6$$

EXPERIMENTAL PROCEDURE

WARNING:

Never look into an open-ended waveguide system.
Microwave radiation can cause harm.

1. Set up the equipment as shown in Figure 2.2.4 with the resistive termination component connected to the slotted line section. The depth of penetration of the probe of the diode detector into the slotted line should be set at approximately 1 to 2 mm.

   Our first task is to measure the VSWR of the resistive termination component.

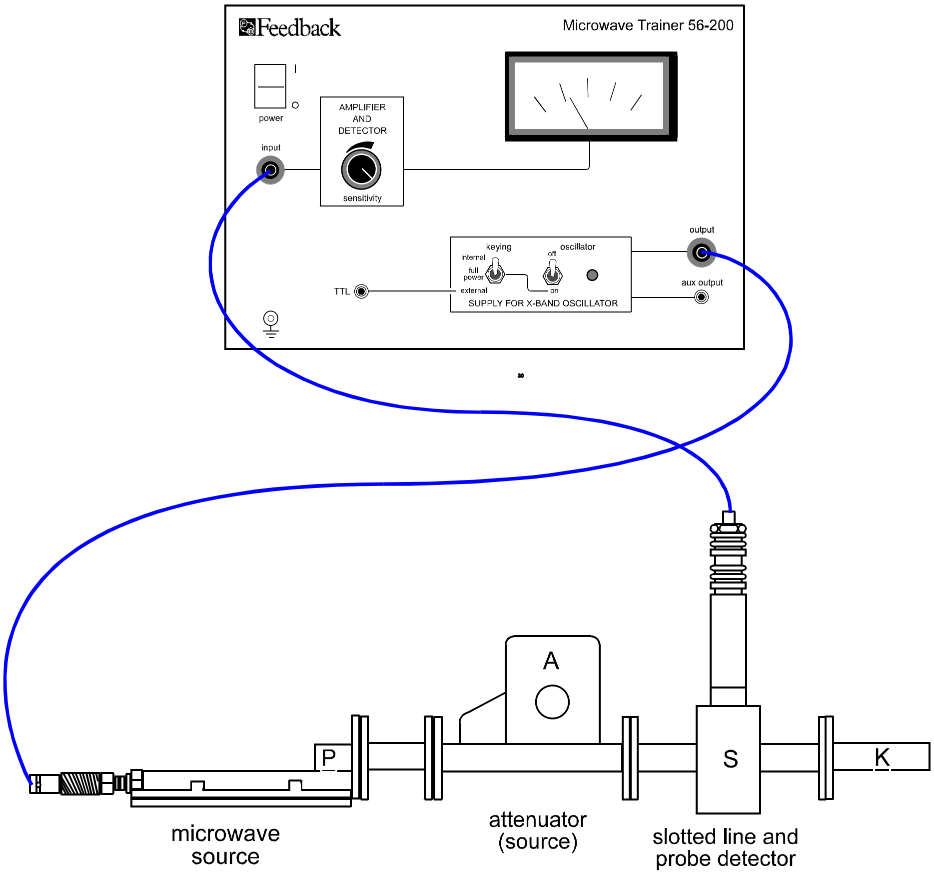


Figure 2-2-4: Experiment set up to Measure VSWR Resistive Load

2. Set the switch conditions on the console as follows:

amplifier and detector: switch to detector output

amplifier and detector sensitivity control: turn to mid-position supply for X-band oscillator:

left-hand keying switch: switch to internal keying

right-hand switch: initially off

Now switch on the console power supply, the main green switch, and energise the microwave bench by switching the right-hand X-band oscillator switch on

3. Set the attenuator at about 25° to provide a reasonable level or attenuation in the system. This is good practice since attenuation between the microwave oscillator and the equipment it drives helps reduce detuning of the oscillator due to reflected signals.

Move the probe-detector unit along the slotted line to locate a position of electric field maximum. Adjust the detector sensitivity of the amplifier and detector on the console and, if necessary, the attenuator setting to obtain a meter reading close to full-scale deflection.

Record the detector current I_{max} corresponding to E_{max}.

4. Now move the probe-detector unit away from the maximum and locate accurately the position of the adjacent minimum.

Record the detector current I_{min} corresponding to E_{min}.

5. Using the results: I_{max} = I_{min} =

Calculate for the resistive load termination:

the VSWR $S = \dfrac{E_{max}}{E_{min}} = \sqrt{\dfrac{I_{max}}{I_{min}}} =$

the reflected coefficient,

$$|\Gamma| = \frac{S - 1}{S + 1} =$$

the % of power reflected,

$$100|\Gamma|^2 = \quad \%$$

the return loss,

$$10 \log \frac{P_r}{P_i} = 20 \log |\Gamma| = \quad \text{dB}$$

where Pr, Pi = reflected and incident powers, respectively

6. **Measurement of VSWR for the horn antennas**

Remove the resistive load termination and connect one of the horn antennas. Measure its VSWR by measuring I_{max} and I_{min} as described in steps 2 and 3. Repeat with the second horn.

Results and calculation of VSWR

Horn 1	Horn 2
I_{max} =	I_{max} =
I_{min} =	I_{min} =
$VSWR = \sqrt{\frac{I_{max}}{I_{min}}} =$	$VSWR = \sqrt{\frac{I_{max}}{I_{min}}} =$

7. **Measurement of larger values of VSWR: plotting around the minimum method**

The horn antenna presents a very good match to the waveguide and hence has a VSWR quite close to unity. A component with a higher value of VSWR can be simulated by using the second attenuator terminated by a short-circuit plate.

(i) Remove the horn and connect the attenuator terminated in a short circuit as shown in Figure 2-2-5.

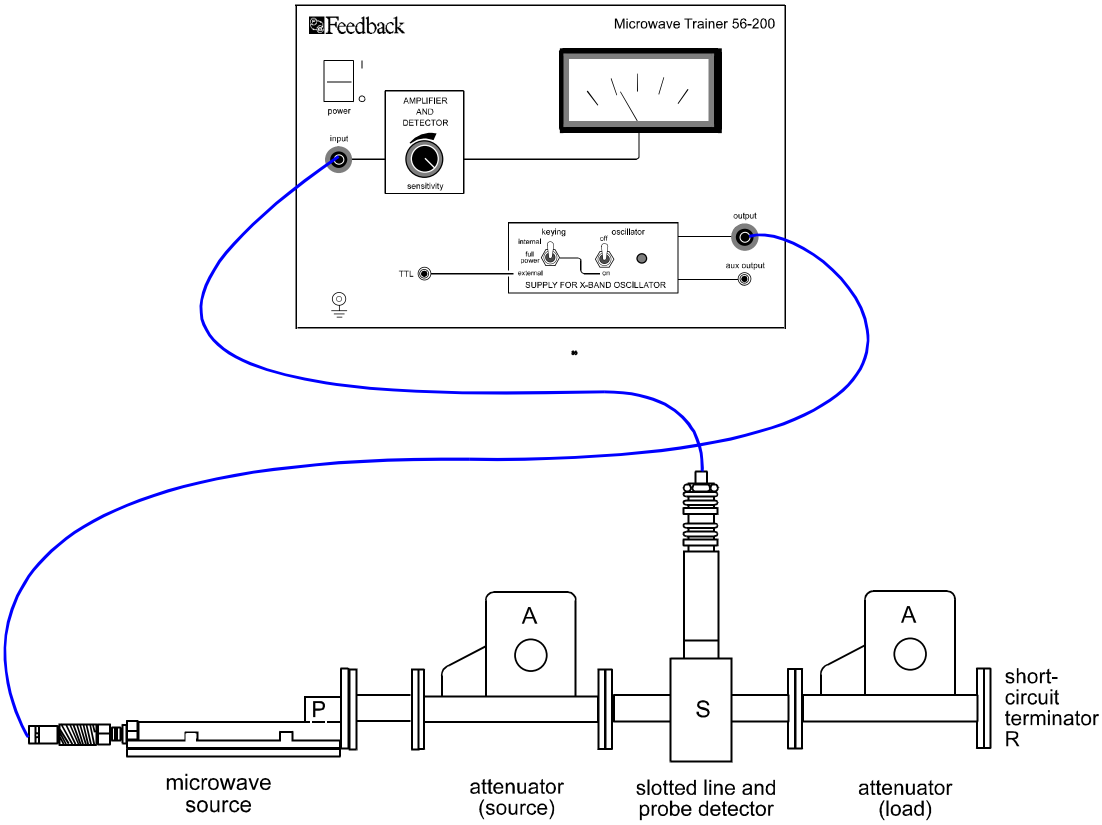


Figure 2-2-5

(ii) Set the attenuator to 40° which provides a fairly low attenuation value (typically 2 to 3 dB).

(iii) Move the probe-detector along the slotted line section to locate an electric field minimum. Note the detector meter reading I_{min}.

(iv) Next move the probe-detector to the right until the detector meter registers twice the minimum current value, i.e. 2 I_{min} (corresponding to $k^2 = 2$). Record the probe-detector position on the slotted line scale, x_1 say.

(v) Now move the detector probe to the left through the minimum and locate the position where the detector meter current is again 2 I_{min}. Record the probe-detector position, x_2.

(vi) Measure the guide wavelength by finding the distance between successive minima and remembering:

distance between successive minima = ½ l_g

(vii) **Results and calculation of VSWR**

position where $I = 2\ I_{min}$: x_1 =

x_2 =

$d = x_1 - x_2$

l_g = 2 x distance between minima =

$$\text{VSWR } S = 1 + \frac{1}{\sin^2\left(\frac{\pi d}{\lambda_g}\right)}$$

SUMMARY

The VSWR of waveguide components has been measured using a waveguide slotted line to measure the ratio of maximum to minimum electric fields produced in the standing wave pattern set up by the components.

Components with a VSWR close to 1 have very low reflection coefficient and present a good match to their waveguide feed.

A method for the measurement of large values of VSWR based on determining the relative width about an electric field minimum has also been undertaken. Both methods assumed a square law for the diode detector. Measurements of VSWR independent of the diode detector characteristics may be made by measuring the ratio E_{max}/E_{min} using a precision attenuator.

Notes

Resistive load

I_{max} = 4 8 mA

I_{min} = 2.4 mA

$$\text{vswr } S = \sqrt{\frac{I_{max}}{I_{min}}} = \sqrt{\frac{4.8}{2.4}} = 1.41$$

Note this result indicates that the resistive load presents a rather poor match with:

$$\text{reflection coefficient}, |\Gamma| = \frac{S-1}{S+1} = \frac{1.41-1}{1.41+1} = 0.17$$

$$\%\text{ power reflected}, |\Gamma|^2 \times 100 = 3\%$$

Horn antenna

Both horns gave very similar results:

I_{max} = 4.2 mA

I_{min} = 3.0 mA

$$\text{so vswr } S = \sqrt{\frac{4.2}{3.0}} = \sqrt{1.4} = 1.18$$

$$|\Gamma| = 0.08$$

$$\%\text{ power reflected} = 0.68\%$$

showing the horns present an excellent match

Attenuator terminated in short circuit

Attenuator setting 40°

$I_{min} = 0.6$ mA

position at $I = 2\, I_{min} = 1.2$ mA

$x_1 = 41.8$ mm

$x_2 = 45.0$ mm

$d = x_1 - x_2 = 45 – 41.8 = 3.2$ mm

$\lambda_g = 36.5$ mm

$$\text{so } \frac{\pi d}{\lambda_g} = \frac{\pi\, 3.2}{36.5} = 0.088\,\pi \text{ rad or } 15.8^\circ$$

$$\text{hence } S = \sqrt{\left[1 + \frac{1}{\sin^2 15.8^\circ}\right]} = \sqrt{[1 + 13.4]} = 3.8$$

CONTENT

The characteristic of rectified current versus microwave power for a diode crystal detector is investigated at low-power levels. The validity of the diode detector "square -law" whereby the rectified current output is directly proportional to incident microwave power (itself proportional to the square of the electric field amplitude) is tested.

EQUIPMENT REQUIRED

Quantity	Identifying letter	Component description
1	---	Control console
1	A	Variable attenuator
1	B	Waveguide slotted line
1	S	Probe diode-detector assembly
1	P	X-band microwave source
1	R	Short circuit plate

OBJECTIVES

When you have completed this assignment you will

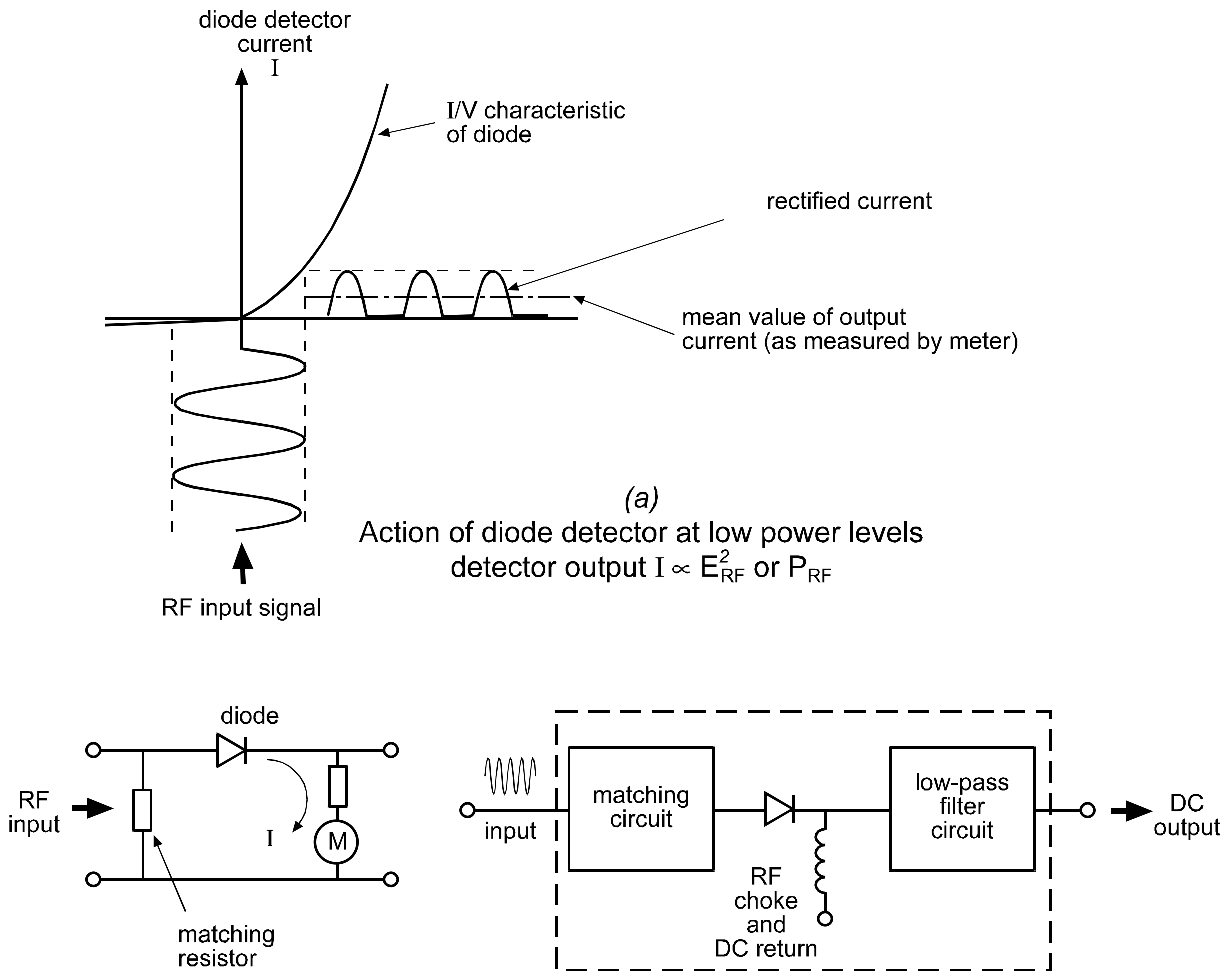
- Appreciate the use of a crystal diode as a sensitive detector of microwave signals
- Know that at low signal levels (typically in the microwatt range) that the rectified output is directly proportional to the square of the electric field and hence the microwave power flowing in the waveguide
- Know how the 'square law' may be tested experimentally
- Gain further experience of standing waves and the use of waveguide components

KNOWLEDGE LEVEL

No prior knowledge is required to carry out this assignment, although it is strongly recommended that assignments 1 and 2 have already been completed.

INTRODUCTION

Diode crystal detectors are widely used as a detector of microwave signals at low power levels. The action of the diode and two basic detector circuits are indicated in Figure 2-3-1.



(a)
Action of diode detector at low power levels
detector output $I \propto E^2_{RF}$ or P_{RF}

(b)
Simple diode detector circuit

(c)
Circuit blocks making up a practical diode detector

Figure 2-3-1: Diode Detector - Action and Basic Circuits

For power level up to about 10 microwatts the crystal detector has a square law characteristic, that is, the detector voltage or current output is directly proportional to the square of the electric field, which in turn implies that the diode detector output is proportional to the power of the microwave signal. The low-level sensitivity of a good quality diode detector is of the order 0.5 millivolts per microwatt. At higher power levels the crystal detector deviates from the square law becoming an approximate linear detector at levels of the order of 0.1 milliwatts. Over the linear part of the characteristic the rectified output is directly proportional to the microwave electric field strength.

In this experiment, the diode-detector characteristic is investigated at low-power levels using a slotted waveguide and a probe diode detector unit (see Assignment 2, Figure 2-2-2). The detector current is measured for the known variation of electric field strength present in the slotted waveguide section when this section is terminated in a short-circuit.

The magnitude of electric field strength in a short-circuited line, see Figure 2-3-2, is given by

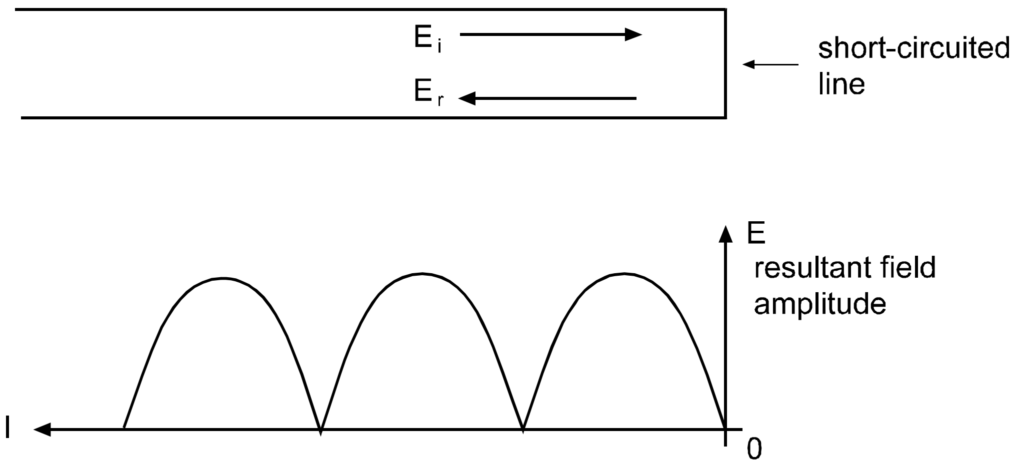
$$E = 2E_i \sin bl$$

where

E_i = incident field amplitude

l = distance from short-circuit

$b = 2p/l_g$



$$E = 2E_i \mid \sin\beta l \mid$$

Figure 2-3-2: Variation of Electric Field Amplitude in the Standing Wave Pattern of a Short-Circuited Line

If we assume the average diode detector rectified current as measured by the meter is related to field strength by the relationship

$$I = k E^n$$

where n = 'crystal law', then we have for the case of a short-circuited line:

$$I = k\,(2E_i \sin bl)^n$$

so if we were to plot log I (y - axis) versus log (sin bl), the graph should be a straight line and the gradient or slope of this line gives n, the crystal law power. For low-power levels we should establish that n = 2 showing the square-law relationship is valid.

EXPERIMENTAL PROCEDURE

WARNING:

Never look into an open-ended waveguide system.
Microwave radiation can cause harm

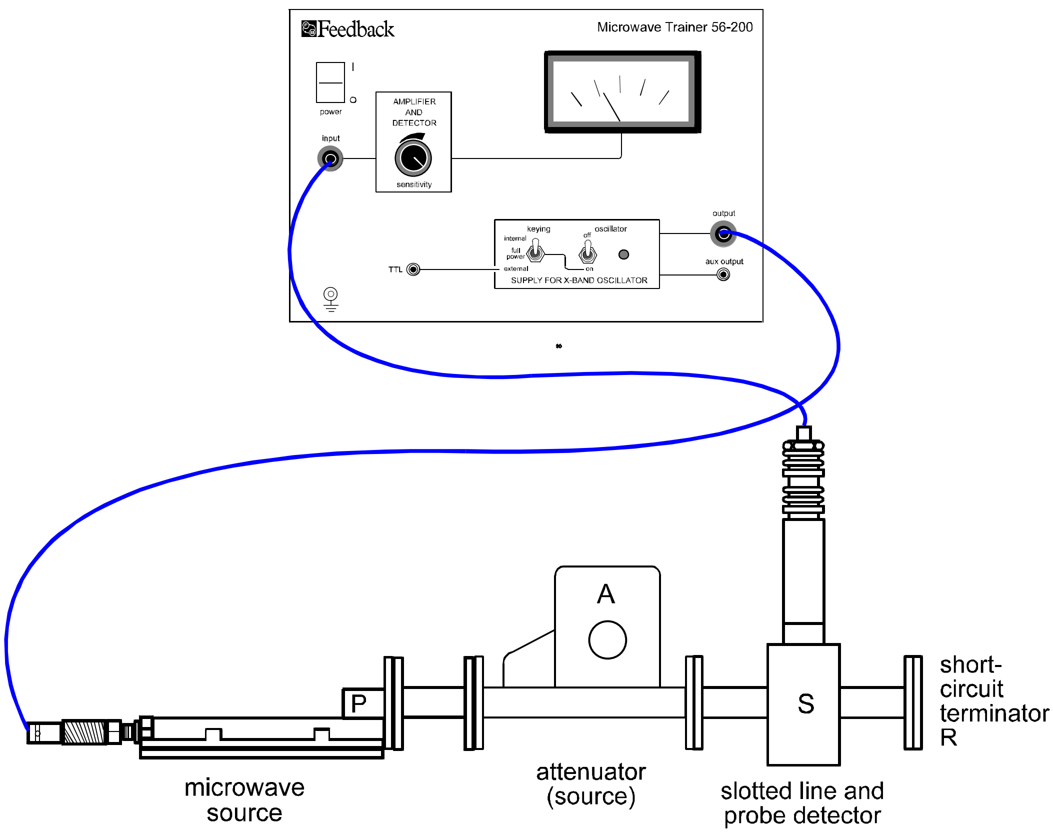


Figure 2-3-3

1. Set up the equipment as shown in Figure 2-3-3 with the slotted waveguide section terminated in a short-circuit plate and the variable attenuator set initially at 0°, its maximum attenuation setting.

The depth of penetration of the probe of the diode detector should be set at approximately 1 to 2 mm. This ensures that the field in the slotted section is not unduly disturbed by the probe but still provides a low-level but measurable input to the diode detector.

2. Set the switches on the console as follows:

left-hand keying switch: switch to internal keying

right-hand switch: initially off

Now switch on the console supply (main green switch on) and energise the microwave oscillator by switching the right-hand switch on.

3. Gradually reduce the attenuation in the attenuator by moving the resistive vane from the 0° position until a deflection is observed on the meter. Move the probe unit along the slotted waveguide section and locate a position of maximum as indicated by maximum detector current reading.

This position corresponds to a maximum of electric field produced by the standing wave pattern set up in the slotted waveguide section by the incident wave and the wave reflected at the short-circuit termination.

Now adjust the attenuator and, if necessary, the detector-amplifier sensitivity control to provide a meter reading of this maximum close to full scale deflection.

4. Next, move the probe along the slotted waveguide and locate accurately the position of an electric field null. At the position of a null the detector current is virtually zero. Working from this point as reference and away from the short-circuit end of the slotted line, take a series of detector current readings at a number of points up to the position of a field maximum. Record the results in a table similar to that given below

Probe position as measured on slotted-section scale - L mm	Probe position relative to electric field null $l = L - L_o$	Diode detector current I mA
At null, E = $L_1 = L_0$ =	$l = 0$	E = $I = 0$

5. Measure the guide wavelength l_g by determining the distance between successive nulls. It should be possible to measure the positions of 3 successive nulls;

first null position, L_1 =

second null position, L_2 =

third null position, L_3 =

then $l_g = 2(L_2 - L_1)$ =

$= 2(L_3 - L_2)$ =

$= L_3 - L_1$ =

6. Calculate the values of log I and log (sin bl) tabulating your results in a table similar to that given below

Distance l	bl = 2pl/l_g = 360 l/l_g°	sin bl	log (sin bl)	Detector current I mA	log I
0	0	0	---	0	---

7. Plot the graph of log I versus log (sin bl) and draw in the "best-fit" straight line.

Measure the gradient of this line and hence deduce the diode-detector law, the value of n. This should be approximately 2.

SUMMARY

Diode crystal detectors are very widely used for low-level microwave signal detection. They are extremely sensitive devices and at very low power levels up to about 10 micro-watts they obey a square law, i.e. detector output current (or voltage) is directly proportional to the square of electric field; equivalently detector output is directly proportional to microwave power.

In this experiment a known field variation in a waveguide is set up by short-circuiting a length of slotted line. This field was sampled using a probe-diode detector unit. The RF field picked up by the probe was rectified by a crystal diode and measured at a number of points between positions of minimum and maximum as the probe unit was moved along the slot of the waveguide section. Results were used to plot a graph and test the validity of the square law condition:

detector output $I \propto E^2$ (electric field squared)

$\propto P_{RF}$ (microwave power)

SPECIMEN RESULTS Guide wavelength measured at, l_g = 37.1 mm

l mm	360 l/l_g°	sin bl	log (sin bl)	I mA	log I
0	0°	0	---	0	---
1.0	9.7	0.169	-0.78	0.2	-0.7
2.0	19.4	0.332	-0.48	0.65	-0.19
3.0	29.1	0.486	-0.31	1.7	+0.23
4.0	38.8	0.628	-0.20	2.65	+0.42
5.0	48.5	0.749	-0.13	3.5	+0.54
6.0	58.2	0.850	-0.07	4.05	+0.60
7.0	67.2	0.927	-0.03	4.45	+0.65
8.0	77.6	0.977	-0.01	4.95	+0.70
9.0	87.3	0.999	-0.005	5.15	+0.71

The graph of log I versus log (sin bl) is plotted below in graph 2-3-4 and a 'best-fit' line drawn. The gradient or slope of this line gives n = 1.86, a fair approximation to the square-law relationship.

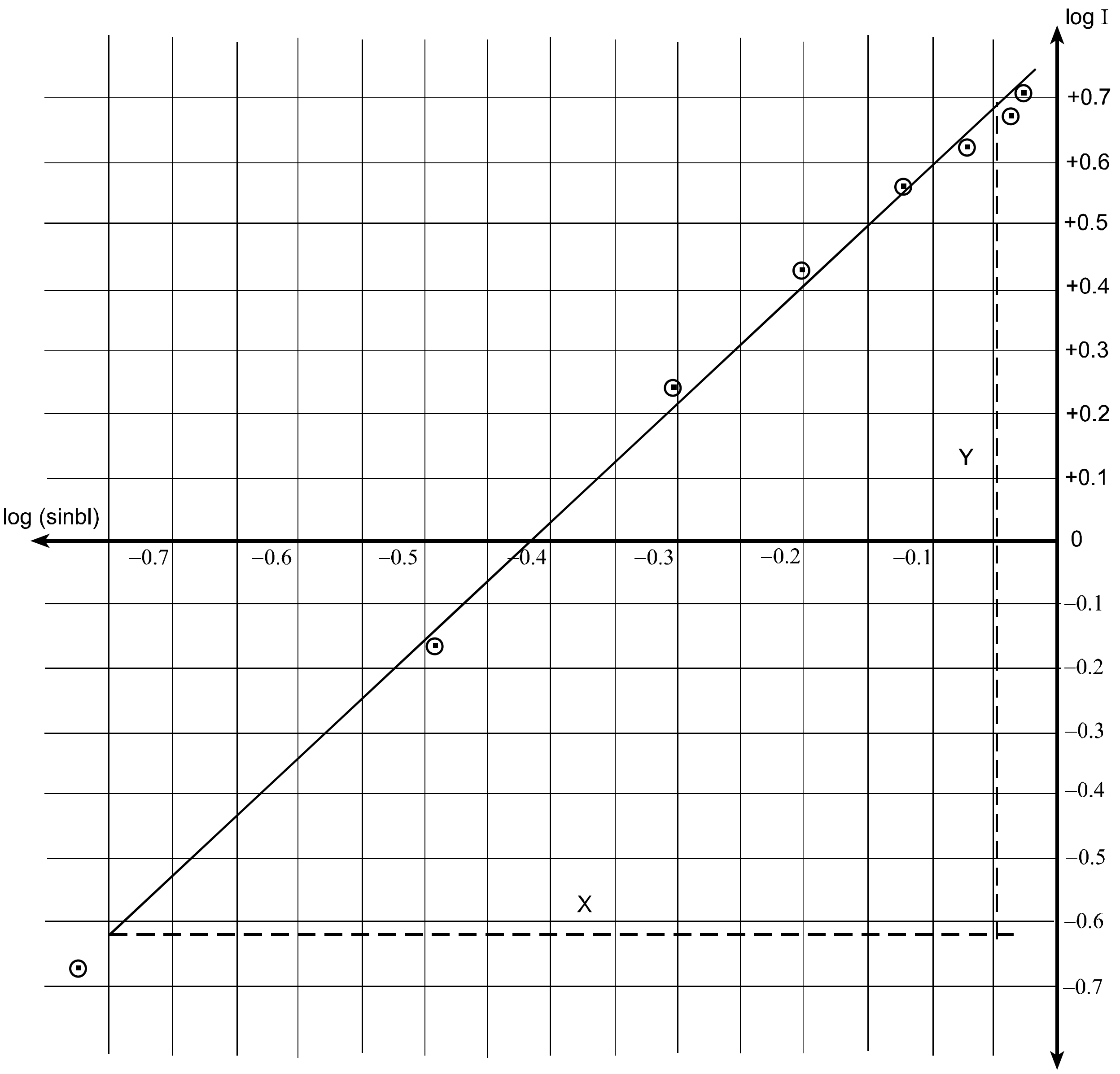


Figure 2-3-4: Graph of Log I versus Log sin βl

$$\text{gradient } n = \frac{Y}{X} = \frac{1.3}{0.7} \cong 1.86$$

CONTENT

The concept of impedance in waveguides and the use of the Smith Chart in impedance and matching calculations are introduced. The measurement of impedance of a waveguide component is carried out and the results used to determine the position of a capacitative probe to effect matching.

EQUIPMENT REQUIRED

Quantity	Identifying letter	Component description
1	---	Control console
1	A	Variable attenuator
1	B	Waveguide slotted line
1	C	Slotted line probe tuner
1	K	Resistive termination
1	P	X-band oscillator source
1	S	Probe-diode detector used with slotted line
1	R	Short-circuit plate

OBJECTIVES

When you have completed this assignment you

- Should understand the terms normalised impedance and admittance of a waveguide component
- Know how to measure normalised impedance by measuring the vswr and position of a electric field or voltage minimum in the standing wave pattern produced by the component
- Appreciate the use of the Smith Chart in aiding impedance determination
- Appreciate the need for matching
- Know how matching can be accomplished using a slotted line probe/screw tuner

KNOWLEDGE LEVEL

Before you commence this assignment you

- Should look at Assignment 2 and know the significance of vswr and how it can be measured
- Appreciate that a load terminating a waveguide or a transmission line will cause reflection and produce a standing wave unless matched

- Be familiar with the basic results of transmission line theory

INTRODUCTION

Summary of basic transmission line and standing wave results of a terminated line

(1) **Input impedance normalised impedances and reflection coefficient of a terminated line**

The impedance of a waveguide load is determined using the general result for the impedance of a terminated line given by transmission line theory. Since voltage and current cannot be measured in a waveguide, the actual measurement of impedance is deduced from the standing wave electric field pattern set up by the load.

From transmission line theory, we have for the input impedance of a line of characteristic impedance Z_0 when terminated in a load of impedance, Z_T, as indicate in Figure 2-4-1.

$$Z_{in}(l) = \frac{V(l)}{I(l)} = Z_0 \frac{Z_T + j Z_0 \tan\beta l}{Z_0 + j Z_T \tan\beta l} \qquad (1)$$

where l = line length

$\beta = 2\pi/l_g$, the phase constant of the line

l_g = guide wavelength

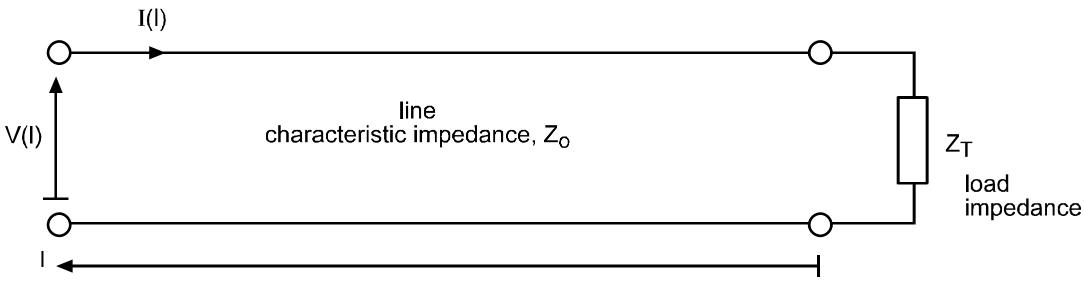


Figure 2-4-1

The result given by (1) is general and can be used to quantify waveguide loads as impedances in much the same way as impedances are used at lower frequencies and in coaxial and parallel wire type lines. In these types of lines voltage and current can be defined and their characteristic impedance uniquely determined. Typical Z_0 values for coaxial lines used in RF work are 50 Ω and 75 Ω.

The characteristic impedance of waveguides, however, cannot be uniquely specified and so to avoid this difficulty impedance levels are relative to Z_0 as normalised impedances.

$$z_{in}(l) = \frac{Z_{in}(l)}{Z_0}$$

$$= \frac{Z_T + jZ_0 \tan \beta l}{Z_0 + jZ_T \tan \beta l}$$

$$= \frac{Z_T / Z_0 = j \tan \beta l}{1 + jZ_T / Z_0 \tan \beta l}$$

$$= \frac{z_T + j \tan \beta l}{1 + jz_T \tan \beta l} \qquad (2)$$

where $z_T = Z_T/Z_0$ = normalised load impedance.

The reflection coefficient at the load is given by:

$$\Gamma = \frac{Z_T - Z_0}{Z_T + Z_0}$$

$$= \frac{z_T - 1}{z_T + 1} \qquad (3)$$

so unless $Z_T = Z_0$ or equivalently $Z_T = 1$ reflection will occur. If the load is matched to the line, that is $Z_T = Z_0$, then G = 0 and no reflection occurs.

(2) **Admittance**

As well as working in impedances it will be also convenient to work in admittances. This is especially so in matching applications where shunt elements are employed (elements connected in parallel).

Admittance is the reciprocal of impedance and so formulae (1), (2) and (3) can be expressed in terms of admittance as

$$Y_{in}(l) = \frac{I(l)}{V(l)} = Y_0 \frac{Y_T + jY_0 \tan\beta l}{Y_0 + jY_T \tan\beta l} \qquad (1^1)$$

$$y_{in}(l) = \frac{y_T + j\tan\beta l}{1 + jy_T \tan\beta l} \qquad (2^1)$$

$$\Gamma = \frac{Y_0 - Y_T}{Y_0 + Y_T} = \frac{1 - y_T}{1 + y_T} \qquad (3^1)$$

2. Measurement of normalised impedance of a waveguide load

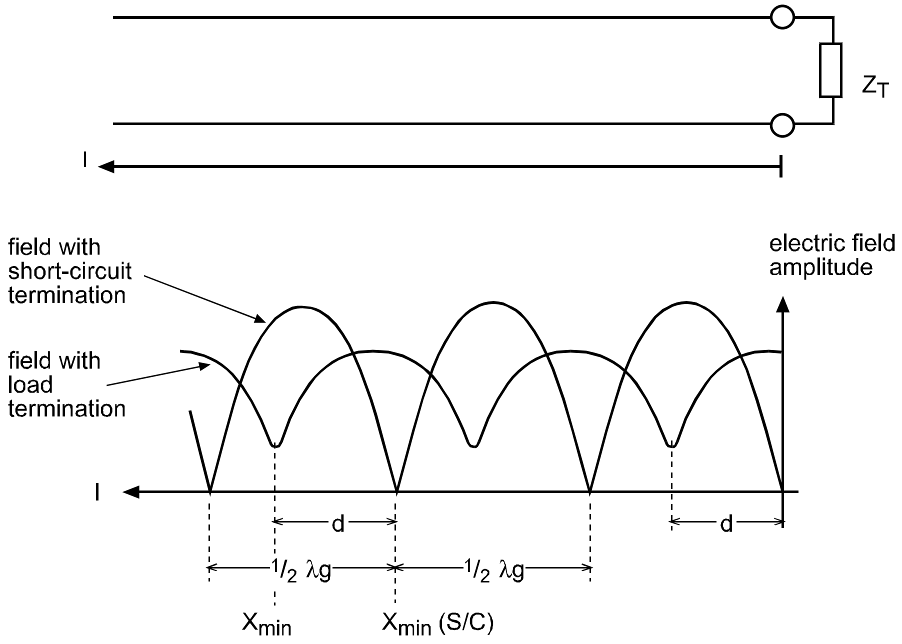


Figure 2-4-2

Figure 2-4-2 illustrates a typical standing wave pattern set up by a load terminating a line when a mismatch between line and load occurs. The plot of electric field amplitude versus distance from the load shows a variation between points of maximum E_{max} and minimum E_{min}. By measuring the voltage standing wave ratio, the VSWR, S, and the position of the first minimum, d, from the load, the normalised load impedance Z_T can be determined.

At a voltage or electric field minimum, the incident and reflected waves are in anti-phase and the normalised input impedance of the line at this point is resistive and equal to 1/S. Thus:

$$\text{when } l = d \qquad z_{in}(d) = \frac{1}{s} = \frac{z_T + j\tan\beta d}{1 + jz_T \tan\beta d} \qquad (3)$$

$$\text{where} \qquad S = \frac{E_{max}}{E_{min}} \text{, the VSWR}$$

and $d = \text{distance from load of first E - field minimum}$

We solve (3) for the normalised load impedance:

$$Z_T = \frac{1 - jS\tan\beta d}{S - j\tan\beta d}$$

and expressing z_T in terms of its real and imaginary parts, we have

$$\text{real or resistive part of } z_T, \quad r_T = \frac{S(1 + \tan^2 \beta d)}{S^2 + \tan^2 \beta d} \qquad (4)$$

$$\text{imaginary or reactive part of } z_T, \quad x_T = \frac{(1 - S^2)\tan\beta d}{S^{12} + \tan^2 \beta d} \qquad (5)$$

In the practical experiment the vswr is measured by measuring the detector current I_{max} at a maximum and at a minimum I_{min} using the slotted line:

$$S = \frac{E_{max}}{E_{min}} = \sqrt{\frac{I_{max}}{I_{min}}}$$

The distance, d, is best measured by recording the positions of minimum with load and short-circuit terminations, then with reference to Figure 2-4-2

$d = x_{min}$ (load) - x_{min} (short-circuit)

The calculations to determine z_T expressed by the formulae (4) and (5) can be accomplished by use of a calculator or with the aid of a Smith Chart. The procedure using a Smith Chart is explained in the Typical Results section at the end of this assignment.

3. Matching a load using a reactive stub

The slotted line probe tuner, component C, shown in Figure 2-4-3, acts as a shunt reactance element and is a useful component in effecting matching. For small penetrations of the probe, capacitative susceptance can be introduced which increases the penetration, but eventually changes sign and becomes inductive. In most matching applications, units of this type consist of either a probe, post or screw and are inserted to produce capacitative susceptance.

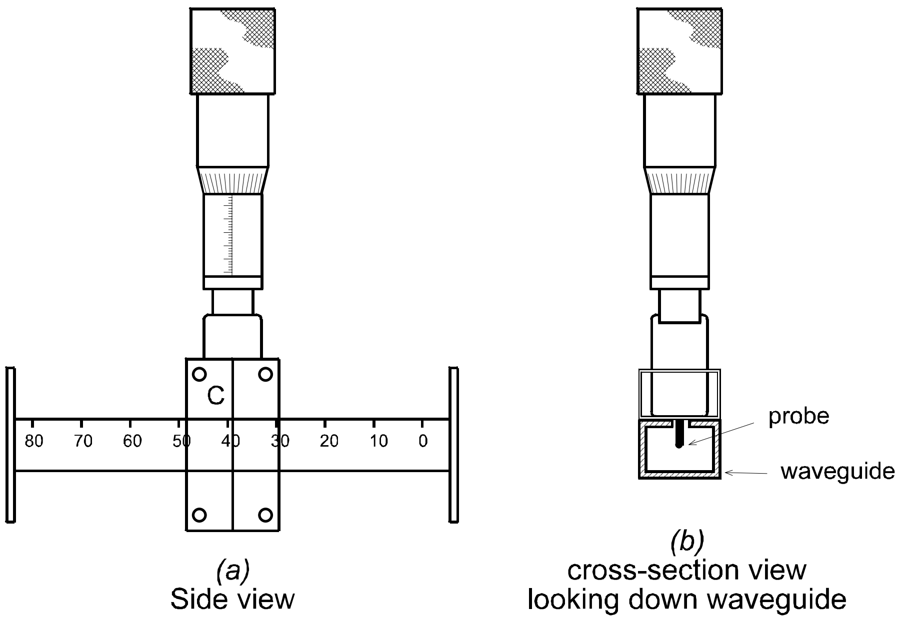


Figure 2-4-3 Slotted Line Probe Tuner

The principle of shunt stub matching may be explained with reference to Figure 2-4-4(a). We look for a point at a distance from the load where the real part of the input admittance is unity, i.e. where:

$$y_{in}(l) = 1 \pm jb$$

Then if at this point we connect in a shunt stub of susceptance equal in magnitude but opposite in sign we can cancel out the load susceptance so that $y_{in} = 1$ and thus effect matching to the feed line.

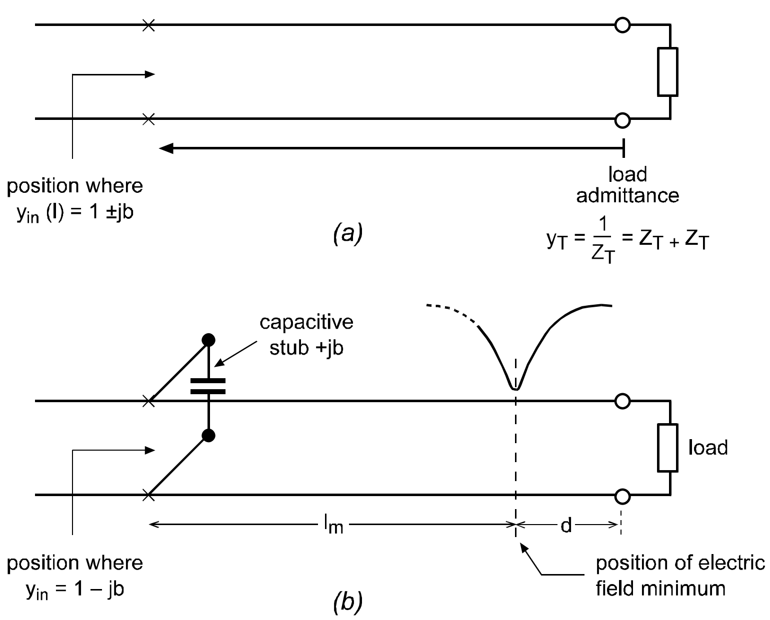


Figure 2-4-4 Shunt Stub Matching

We can locate the position where yin(l) = ±jb by finding the position d of the first electric field minimum and then using the known admittance at this point as a new load:

$$y_T = S \text{ at a minimum}$$

in the general formula for transmission line input admittance. We solve this for the condition that the real part of $y_{in} = 1$, i.e:

$$y_{in}\,(l_m) = \frac{y_T + j\tan\beta lm}{1 + jy_T \tan\beta lm} \quad \text{where } y_T = S$$

solving for the real part of yin and equating to unity, we obtain the condition:

$$\tan^2 \beta l_m = l/S$$

so $$l_m = \frac{1}{\beta}\tan^{-1}\ (l/S) = \frac{\pm\lambda g}{2\pi} = \tan^{-1}\ (l/S)$$

We have in fact two values, one positive and one negative, for lm, so there are two positions either side of the voltage minimum position satisfying real $[y_{in}]$ = 1. The negative value of lm corresponds to yin = 1 + jb and would require inductive susceptance for matching, the positive value used in this assignment requires capacitative susceptance as yin = 1 –jb.

EXPERIMENTAL PROCEDURE

WARNING:

Microwave radiation can be harmful, especially to eyes.
NEVER look into energised waveguide.

Measurement of the normalised impedance of a waveguide

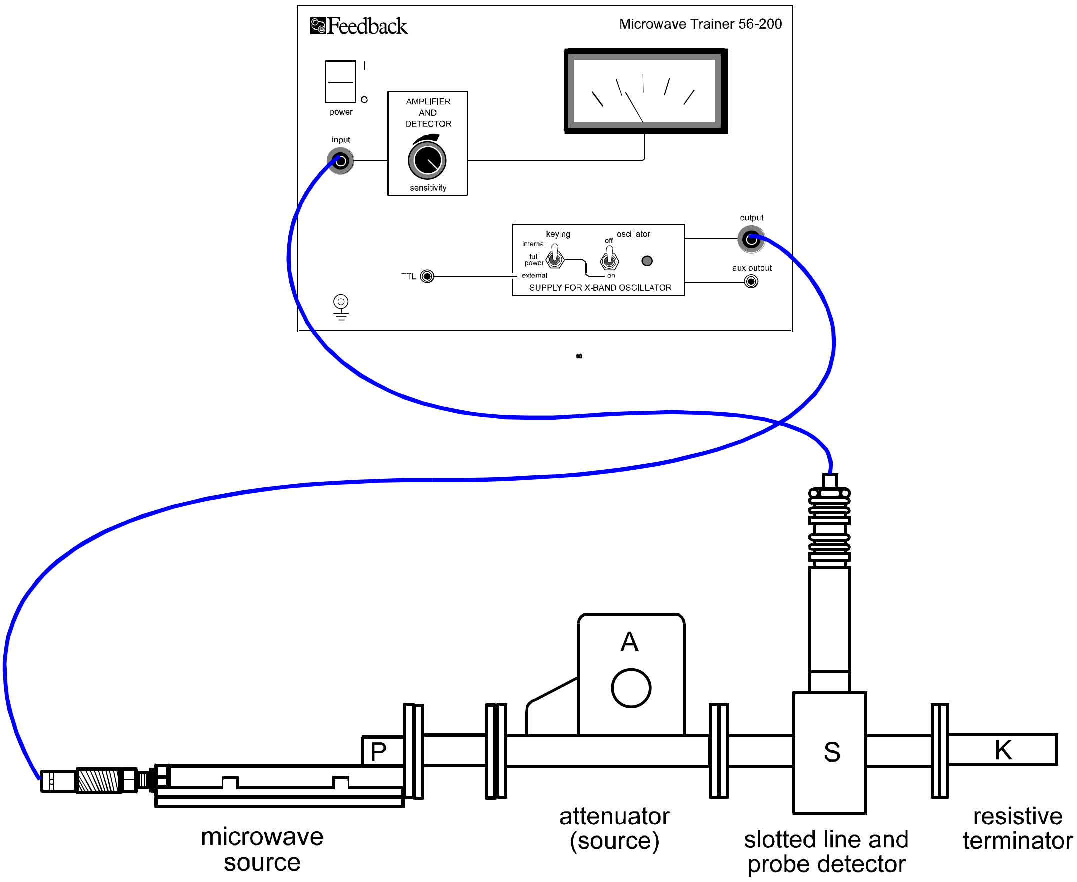


Figure 2-4-5

1. Set up the equipment as shown in Figure 2-4-5. The waveguide load whose impedance we are to measure is connected to the slotted line section. Our test load in this experiment is the resistive termination, component K.

Note:
The probe of the probe-detector S mounted on slotted line should be adjusted to penetrate between 1 to 2 mm through the slot into the waveguide. This should ensure sufficient coupling to measure the electric field in the waveguide without disturbing the standing wave set up by the load under

2. Set the attenuator to approximately 20° to provide a fairly good value of buffer attenuation, switch the X-band source to internal keying and meter to detector output. Switch on the console power supply and the X-band oscillator.

3. Move the detector carriage along the slotted waveguide section to locate a position of an electric field maximum. Adjust the sensitivity of the detector-amplifier on the console to provide a meter reading of 3 to 4 mA. If necessary adjust the attenuator setting.

4. The microwave bench is now set up to measure the VSWR of the resistive load. Move the probe-detector carriage to a position closest to the load and then moving away from the load record positions of electric field minima and values of detector output current at positions of both minima and maxim. Record results in a table similar to that given in Figure 2.-4-6 below.

With resistive load		With short-circuit
Positions of max. & min.	Detector Current	Positions of max. & min.
x_{1min} =	I_{min1} =	x_1 s/c min =
x_{2max} =	I_{max2} =	
x_{3min} =	I_{min3} =	x_3 s/c min =
x_{4max} =	I_{max4} =	
x_{5min} =	I_{min5} =	x_5 s/c min =

Figure 2-4-6: Table for Recording Results for Impedance Measurement

5. Remove the resistive load and replace by the short-circuit plate R so that the slotted line is now terminated in a short-circuit.

Locate and record positions of electric field minimum as the probe-detector carriage is moved along the slotted line starting closest to the short-circuit and moving progressively along the slot towards the source.

6. Calculate from your results the VSWR S of the load, the guide wavelength λ_g and the distance d of first electric field minimum from the load input.

Remember:

$$\text{VSWR } S = \sqrt{\frac{I_{max}}{I_{min}}}$$

$$\lambda_g = 2\left(x_{3s/c} - x_{1s/c}\right)$$

$$= x_{5s/c} - X_{1s/c}$$

(note: use short-circuit positions for l_g since these are well-defined nulls and hence can be measured very accurately)

$$d = x_1 = x_{1s/c}$$

$$= x_3 - x_{3s/c}$$

7. Finally calculate the normalised impedance of the load, $z_T = r_T + j\, x_T$ where the resistive part r_T and imaginary part x_T are given by the formulae (4) and (5) in the **Introduction** section.

8. Check your result using the Smith chart.

Matching a waveguide load using a slotted waveguide probe tuner

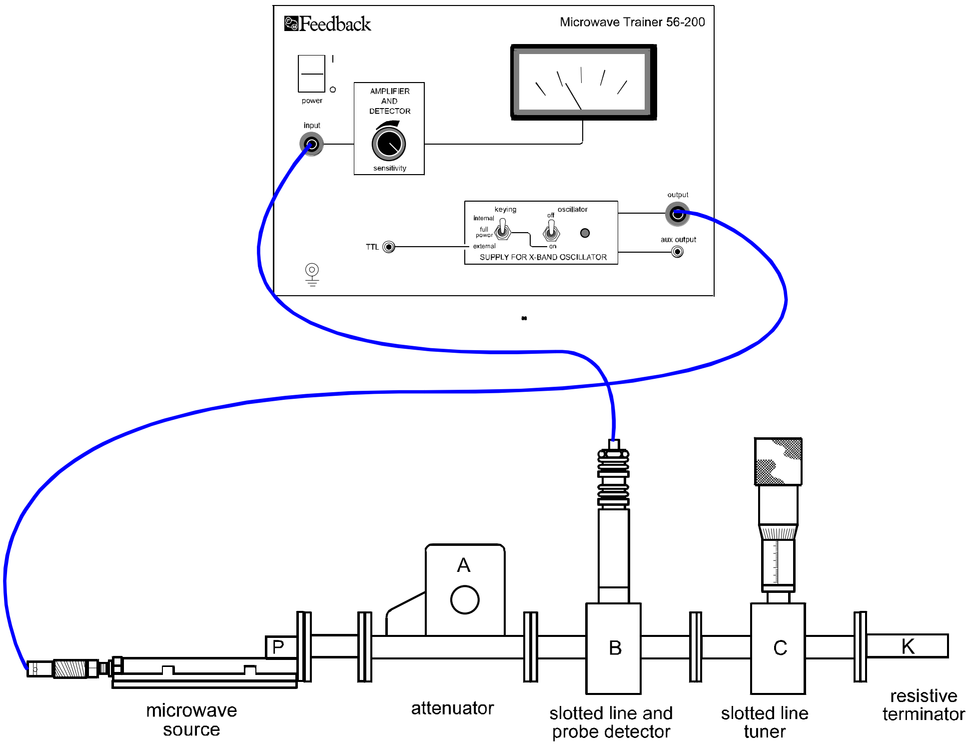


Figure 2-4-7

1. Set up the equipment as shown in Figure 2-4-7 with the probe tuner C inserted between the slotted line and the resistive termination K.

   The probe-tuner unit is to be used to match the resistive load and improve its VSWR, ideally to unity.

2. The probe position is to be located where the real part of the normalised input admittance of the load is unity and the imaginary part (the susceptance) is inductive, i.e. when

   $$y_{in}\ (l) = 1 - jb$$

   The inductive susceptance -jb can then be cancelled by introducing the probe at this point and adjusting the depth of penetration until the capacitive susceptance of the probe +jb cancels -jb. The resultant admittance,

$$y_{in} = 1 - jb + jb = 1$$

and so matching is achieved.

3. The desired position of the probe-tuner is given by the formula in the **Introduction** section:

$$l = d + l_m + \tfrac{1}{2}n\, l_g$$

where d = distance of first electric field minimum from load

$$l_m = \frac{\lambda_g}{2\pi} \tan^{-1}\left(\frac{1}{\sqrt{S}}\right)$$

$$S = \text{load VSWR}$$

$$n = 0,\ 1,\ 2,\ 3...$$

Using the results obtained in the previous impedance measurement section, calculate l_m and locate the probe at the required position l.

Notes:

(i) The scale of the probe-tuner carriage unit zero is at 10mm from its right-hand flange, so the probe should be positioned at (l - 10) mm relative to this scale.

(ii) The minimum practical value of l should be selected for most effective matching; the condition n = 0 is not usually physically obtainable so select n = 1 giving:

$$l = d + \tfrac{1}{2}l_g$$

4. Having set the probe at the position (l - 10) relative to the tuner scale gradually increase the probe penetration measuring the VSWR, at each position. With small penetration depths little improvement in VSWR is obtained. However, as the matching condition is approached where the capacitive susceptance of the probe cancels the load input susceptance a sharp decrease in VSWR can be observed.

Repeat the best value of VSWR and compare with the load VSWR before matching in terms of the percentage of the power reflected and percentage of power transmitted.

5. Check the values of l and l_m obtained above using a Smith chart.

SUMMARY

Methods of impedance measurement and matching have been investigated experimentally. The normalised impedance of a waveguide load has been measured and also matched using a shunt stub in the form of a probe tuner.

Results may be processed analytically using transmission line formulae and also with the aid of a Smith chart.

Notes

Measurement of resistive load impedance

VSWR S = 1.63

position of minima with load

21.9, 41.0, 59.2 mm

positions of minima with short-circuit

27.9, 46.3, 65.1 mm

guide wavelength (average value)

λ_g = 37.2 mm

Position of first minimum from load

d = 41 –27.9 = 13.1 mm

= 59.2 –46.3 = 12.9 mm

average value d = 13 mm

$$\frac{d}{\lambda_g} = \frac{13}{37.2} = 0.35$$

$$\beta d = \frac{2\pi}{\lambda_g} d = 2\pi\,(0.35)\ \text{rad or } 126°$$

Calculation of normalised load impedance

tanßd = tan 126° = -1.38; S = 1.63, S^2 = 2.66

Using formulae (4) and (5) in **Introduction** section:

$$r_T = \frac{S(1 = \tan^2 \beta d)}{S^2 + \tan^2 \beta d} = \frac{1.63(1+1.9)}{2.66+1.9} = 1.03$$

$$x_T = \frac{\tan \beta d\,(1 - S^2)}{S^2 + \tan^2 \beta d} = \frac{-1.38(1-2.66)}{2.66+1.9} = 0.5$$

Calculation of normalised load impedance using a Smith chart

1. Draw in the VSWR circle with S = 1.63, that is the circle centre point M (1, 0) on the chart cutting the vertical ρ axis at the point P = 1.63. See Smith chart of Figure 2.4.8.

2. To locate normalised impedance:

   (i) point Q where r = 1/S = 1/1.63 = 0.61 represents the input impedance of the load at electric field minimum.

   (ii) Move d/l_g = 13/37.2 = 0.35 wavelengths from the point Q in the counter clockwise direction, that is the direction marked towards load on the chart.

   (iii) Locate 0.35 on outer wavelength scale, point K. Join K to centre point M.

   (iv) The point Z where the line KM cuts the VSWR circle provided the coordinates of the normalised impedance of the load:

   z = (1.05, 0.5) so r_T = 1.05, x_T = 0.5

   and $z_T = 1.05 + j\,0.5$

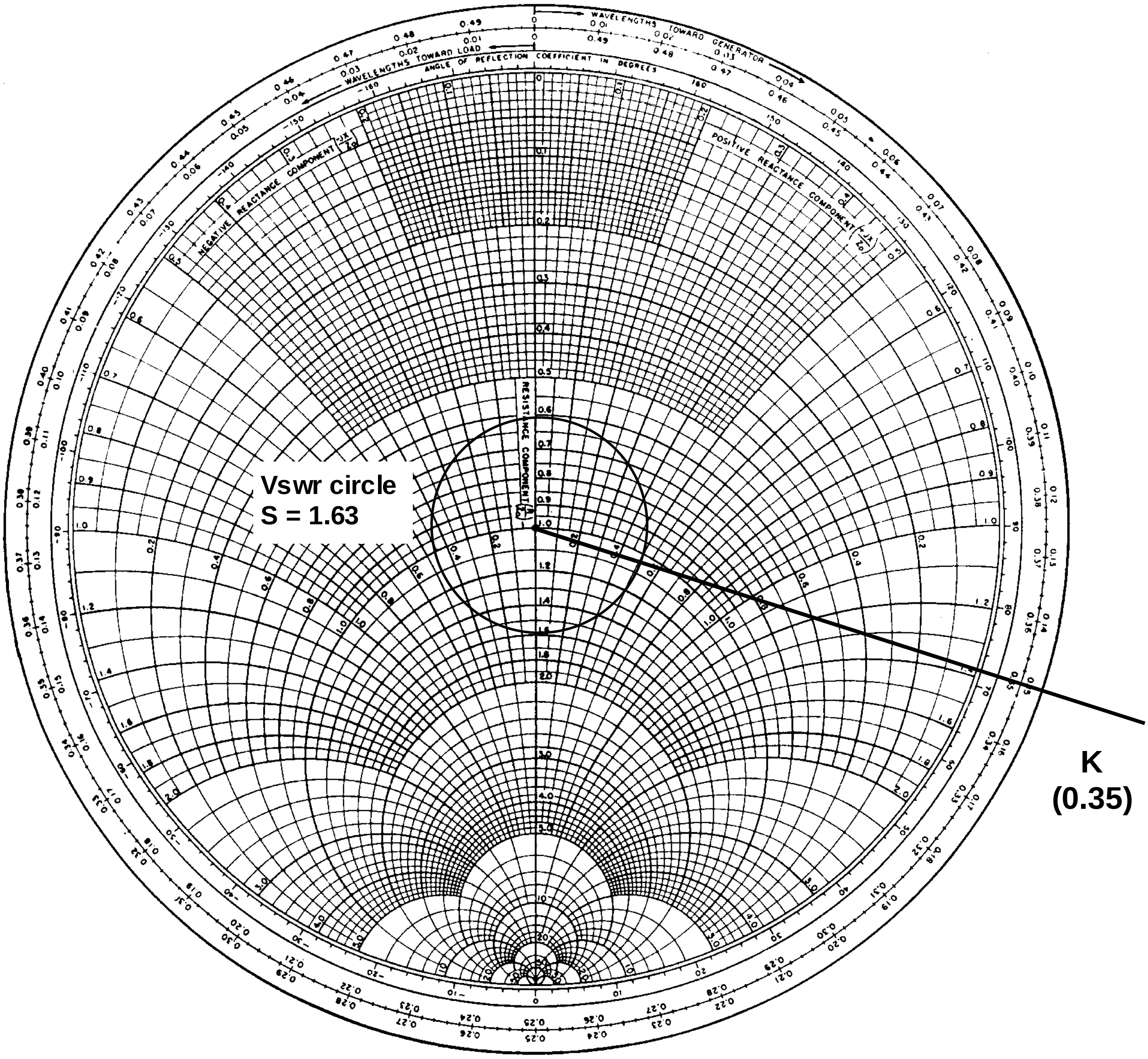


Figure 2-4-8:
Determination of load impedance using a Smith Chart

Matching results: matching resistive load using slotted guide probe tuner

Calculation of probe tuner from load impedance measurements

Normalised load impedance, calculated from previous results, is given in Figure 2-4-9.

Distance of first minimum,	d = 13 mm
Guide wavelength	l_g = 37.2 mm
	d/l_g = 0.35
Load VSWR	S = 1.63

distance from minimum where input admittance is y_{in} = 1 - jb, that is at the plane where matching is to be effected is given by:

$$l_m = \frac{\lambda_g}{2\pi} \tan^{-1} \frac{1}{\sqrt{S}}$$

$$= \frac{\lambda_g}{2\pi} \tan^{-1} \frac{1}{\sqrt{1.63}}$$

$$= \frac{\lambda_g}{2\pi} \tan^{-1}(0.783) = 0.1\lambda_g$$

Thus total distance from load for matching plane:

$l = l_m + d$

$= 0.1\ l_g + 0.35\ l_g = 0.45\ l_g$

and this is repeated at further ½ l_g intervals in the direction towards generator.

Hence matching positions are:

$l = 0.45\ l_g = 0.45 \times 37.2 = 16.7$ mm

$l = 0.45\ l_g + 0.5\ l_g = 0.95\ l_g = 35.3$ mm

$l = 0.45\ l_g\ 1.0\ l_g = 1.45\ l_g = 53.9$ mm and allowing for 10 mm displacement of tuner scale zero, the tuner probe should be placed at 6.7 mm or 25.3 mm of 43.9 mm to effect matching. Selecting position 25.3 mm the following improvement in VSWR was obtained:

probe penetration referenced to micrometer scale: 16.96 mm
VSWR measured, S = 12

Thus matching has improved VSWR from 1.63 to 1.2:

without matching:

VSWR S = 1.2

reflection coefficient G = S-1/S+1 = 0.3

percentage of power reflected, G^2 x 100 = 9%

with matching:

VSWR S = 1.2

G = S-1/S+2 = 0.091

percentage of power reflected, less than 1%

Matching using Smith Chart

The steps described below refer to the Smith chart of Figure 2.4.9 used to execute the matching calculations.

1. Draw VSWR circle with S = 1.63 and locate load impedance by moving d/l_g = 0.35 wavelengths in the direction towards the load from the position Q on the r-axis, r = 1/S = 0.61.

2. Load impedance is at point Z,

z_T = 1.05 + j0.5

Load admittance is diagonally opposite at point Y,

y_T = 0.78 - j0.36

3. Next locate point on the VSWR circle where this circle cuts the resistive/conductive unity of 1.0 circle, i.e. point T. Draw in the line MT.

At point T the normalised input admittance,

$y = 1 - jb$

has a conductance of 1.0 and an inductive susceptance -jb which can be determined from reading off the Smith chart at T: $-jb = -j\ 0.48$ for this particular case. The inductive susceptance is cancelled by the probe tuner introducing an equal and opposite sign capacitive susceptance.

4. The position of the probe of the tuner relative to the plane of the load is determined from the Smith chart:

reference position of load admittance, point Y : 0.044 wavelengths

matching position of tuner, point T : 0.356 wavelengths

so distance of tuner from load = $(0.044 + 0.356)\ l_g$

$= 0.40\ l_g$

5. Note that the matching position is repeated at subsequent intervals of $\frac{1}{2}\ l_g$ from T towards the generator. Correct must also be made for the zero of the tuner scale which is situated 10mm from the plane of the load.

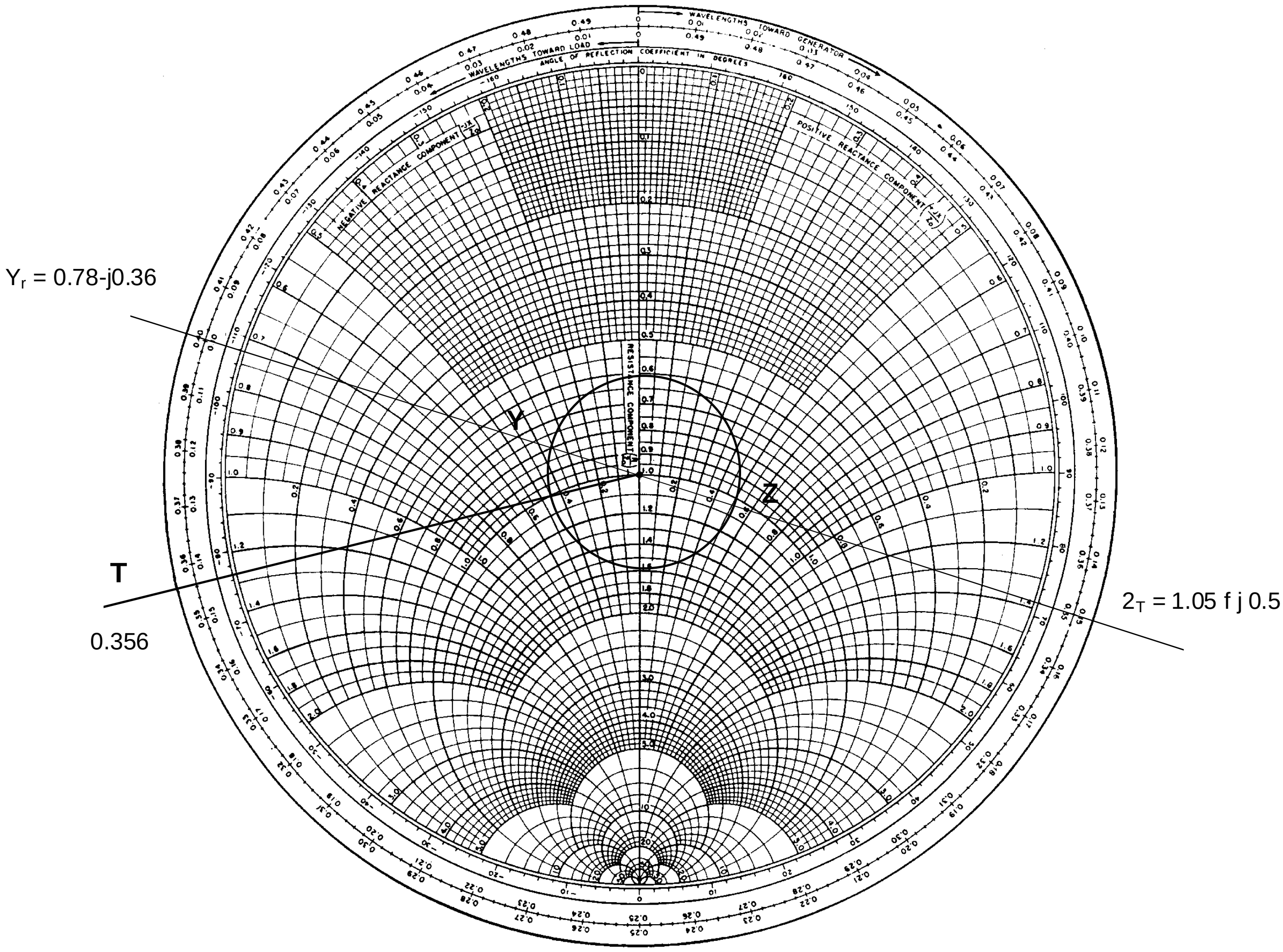


Figure 2-4-9:
Determination of matching position for shunt stub
using a Smith chart

Notes

CONTENT

A test bench is set up to investigate the radiation pattern of a rectangular waveguide pyramidal horn antenna. Polar radiation diagrams are plotted and the 3 dB beamwidth of the antenna is determined.

EQUIPMENT REQUIRED

Qty	Identifying Letter	Description
1	-	Control console
1	A	Variable attenuator
1	K	Resistive load termination
1	M	Diode detector
2	N	Horn antenna
1	P	X-band oscillator
4	-	Supports

OBJECTIVES

When you have completed this assignment you

- Will appreciated the directional radiation characteristics of a horn antenna and know how to plot its radiation pattern
- Know how to measure the beamwidth and gain of an antenna

KNOWLEDGE LEVEL

Before you start this assignment it would be an advantage

- To be familiar with the operation of the microwave bench
- Know that microwave signals can be detected using a diode detector and for low-level signals, that detected output is proportional to power
- To appreciate the role of antennas in the transmission and reception of radio waves

INTRODUCTION

Antennas are essential components in the transmission and reception of radio waves. In the microwave range, highly directive antennas capable of producing the narrow beams required for the line-of-sight links and satellite communications can be designed. The horn antenna, whose radiation characteristics are investigated in this assignment, plays an important role as a radiator of microwave energy in its own right and also as a primary feed for reflector antennas employed in microwave radio links and radar.

The directional characteristics of an antenna - the directions it radiates energy into space - can be visualised graphically by plotting radiated power density versus angular direction. These polar plots are known as far-field radiation diagrams, the latter qualifying the condition that measurements are taken at a sufficiently far distance from the antenna to represent the characteristics as dependent primarily on angular direction. Close to an antenna the radiation pattern is very complex and seldom used in practice. The far-field conditions are satisfied at distances,

$$r \geq 2D^2 / \lambda$$

where D = largest antenna dimension

λ = transmitted wavelength

To fully describe the directional properties of an antenna two radiation diagrams are normally required: one in the horizontal plane, in the case of the horn antenna this would be the H-plane with respect to the horn sketched in Figure 2-5-1, and one for the vertical plane, the E-plane in Figure 2-5-1.

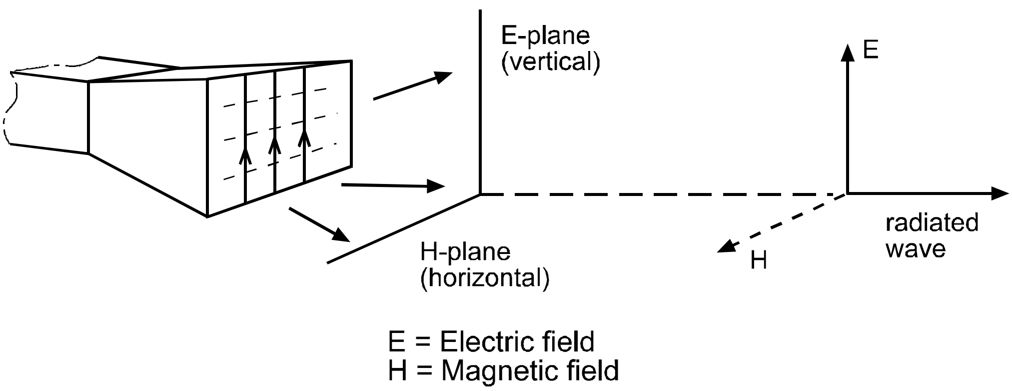


Figure 2-5-1: Pyramidal Horn Antenna

A typical radiation diagram for an antenna with directional radiation (and receiving) properties such as a horn or a parabolic reflector fed by a horn is shown in figure 2.5.2. The angular spread on the main beam between points where the power drops to one-half or by 3 dB from the maximum is known as the 3-dB beamwidth and is an important measure of an antenna's directivity. Not all radiation is confined to the main beam and subsidiary beams at lower power levels and known as side-lobes occur. side-lobes and spill-over radiation can cause interference in microwave radio systems and their levels must be carefully controlled by antenna designers.

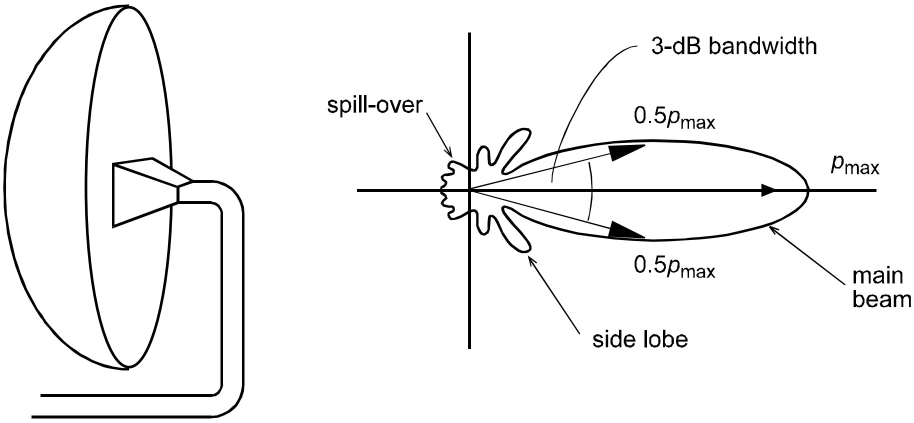


Figure 2-5-2: A Typical Radiation Diagram for a Directional Microwave Antenna

One of the most important parameters of an antenna is its gain. Antenna gain compares the power radiated in the direction of the main beam with that of a hypothetical antenna radiating equally in all directions. Antenna gain is defined as:

G = p/pi

where p = power density W/m^2 radiated by antenna in given direction

pi= power density radiated equally in all directions

= P/4pr^2

P= total power radiated

r= distance from antenna

4pr^2= surface area of sphere radius r

For aperture type antenna such as horns and parabolic reflectors, gain is given by the formula:

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

where A = area of antenna aperture

η = aperture illumination efficiency (typically between 0.5 and 0.8)

Gain is very often expressed in decibels, dB:

G dB = 10 log (p/pi) dBi

The i qualifying, isotropic, the fact that the reference antenna radiates isotropically (equally in all directions). The product of antenna gain G and radiated power P is known as the effective isotropic radiated power normally abbreviated to EIRP:

EIRP = G x P

The power received in a line-of-sight radio link can be expressed in terms of transmitted power and antenna gains; the received power

$$P_R = P_T \; G_T \times \left(\frac{\lambda}{4\pi r}\right)^2 \times G_R$$

where P_T = transmitter radiated power

G_T = gain of transmitter antenna

G_R = gain of receiver antenna

l = wavelength

r = link distance

The above formula is extremely useful in power budget calculations for microwave radio links. It may also be used to measure experimentally the gain of an antenna. If P_R is measured for a given transmitted power and a reference antenna is used over a link of known length, G_T can be determined:

$$G_r = \frac{\left(\frac{P_R}{P_T}\right) \times \left(\frac{4\pi r}{\lambda}\right)^2}{G_R}$$

or in dB $$G_r\ dB = 10 \log \left(\frac{P_R}{P_T}\right) + 20 \log \left(\frac{4\pi r}{\lambda}\right) - G_R\ dB$$

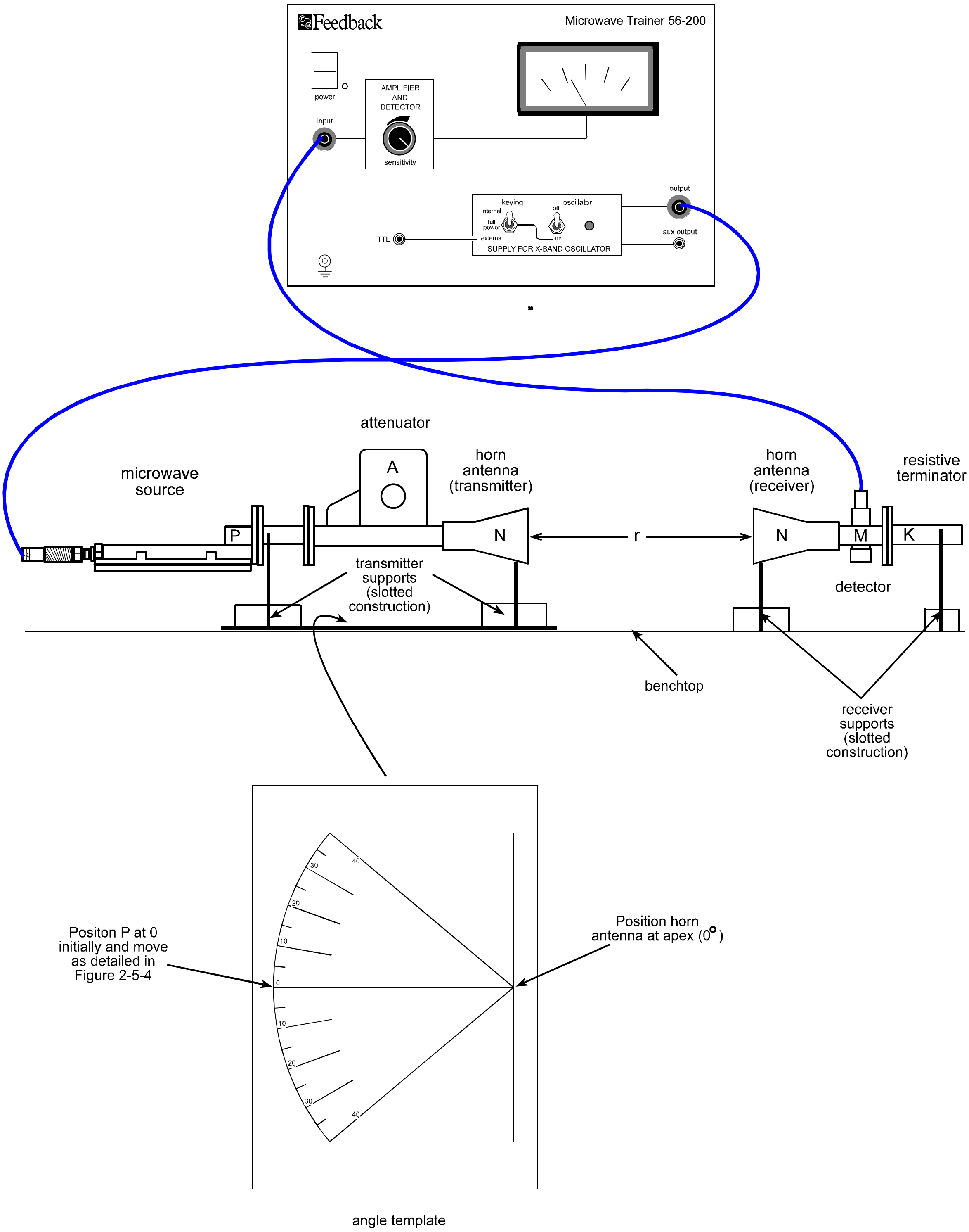

Feedback
Microwave Trainer 56-200
power
input
AMPLIFIER AND DETECTOR
sensitivity
output
keying
oscillator
internal
full power
external
off
on
TTL
aux output
SUPPLY FOR X-BAND OSCILLATOR
attenuator
A
microwave source
horn antenna (transmitter)
horn antenna (receiver)
resistive terminator
P
N
r
N
M
K
transmitter supports (slotted construction)
detector
benchtop
receiver supports (slotted construction)
Positon P at 0 initially and move as detailed in Figure 2-5-4
Position horn antenna at apex (0°)
angle template

Figure 2-5-3

EXPERIMENTAL PROCEDURE

WARNING:

Do **NOT** look directly into the transmitter antenna when taking measurements in this experiment.
Remember, although the power levels produced by the microwave source in the Trainer are low, microwave radiation can cause harm and eyes are particularly sensitive.

1. Connect up the equipment as shown in Figure 2-5-3. Switch the X-band source to internal keying and the meter to detector output.

2. Ensure that the distance r between transmitter and receiver horns is about 30 cm. This distance partially satisfies the conditions that measurements are taken in the far-field, whilst providing sufficient scope for the received signal levels to be detected. The far-field condition is

$$r \geq \frac{2D^2}{\lambda}$$

where D = maximum dimension of horn aperture = 8 cm

λ = wavelength = 2.88 cm for f = 10.425 GHz

so $r = 2D^2/\lambda \cong 44$ cm

Thus the condition is not quite satisfied. However, reasonably accurate results for plotting a radiation diagram can be obtained with r = 30 cm. It is also important to ensure that the radio path between the antennas and their surrounds are free from obstacles, particularly metallic structures, which could cause reflections into the antennas and give rise to false results.

3. Switch on the console power supply and X-band oscillator source. Set the attenuator to a low attenuation setting, typically 40° on the attenuator scale, and turn the amplifier-detector control up to maximum sensitivity.

Align the antennas for the line-of-sight 0° position. In this position the transmitter antenna will be radiating directly in line with the receiver and correspond to maximum antenna gains and maximum received signal.

Adjust attenuator and detector-amplifier sensitivity to obtain a meter reading close to full scale deflection. Record this reading.

4. The radiation diagram for the transmitter horn can now be obtained by rotating the transmitter section from the 0° position through steps of 5° up to 40° either side of the 0° position, using the template provided (on the Manual CD or from the Feedback Website: www.fbk.com/).

Record measurement of received signal level as indicated on the detector meter in a table such as that given below.

The angle of rotation θ° can be set by use of the template, aligning the supports with the template lines as indicated on the template.

Angular direction Θ°	Diode detector meter reading I mA	Angular direction Θ°	Diode detector meter reading I mA
0°		0°	
+5°		–5°	
10°		10°	
15°		15°	
20°		20°	
25°		25°	
30°		30°	
35°		35°	
40°		40°	

Table for logging radiation diagram results

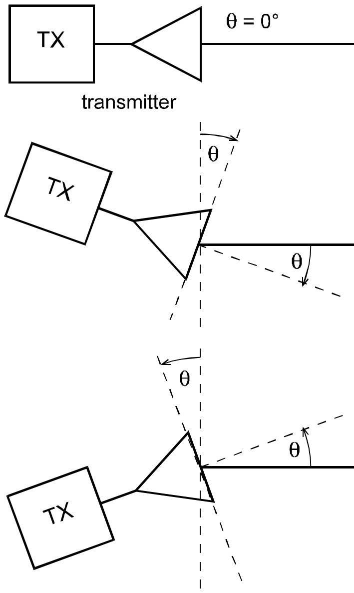


Figure 2-5-4: Measurement of Radiation Polar Diagram of an Horn Antenna

5. Plot the polar radiation diagram of the horn antenna on polar graph paper, an example of which is shown below in Figure 2-5-5. Θ° is the angular direction and I, the diode detector output current, is directly proportional to power for small signal levels. Thus a polar plot of I versus Θ represents the power radiation diagram.

From the radiation diagram, determine the beamwidth between half-power point (3 dB) levels; that is, the angle between points on the polar curve where the power drops to half of the maximum gain of the Θ = 0° position.

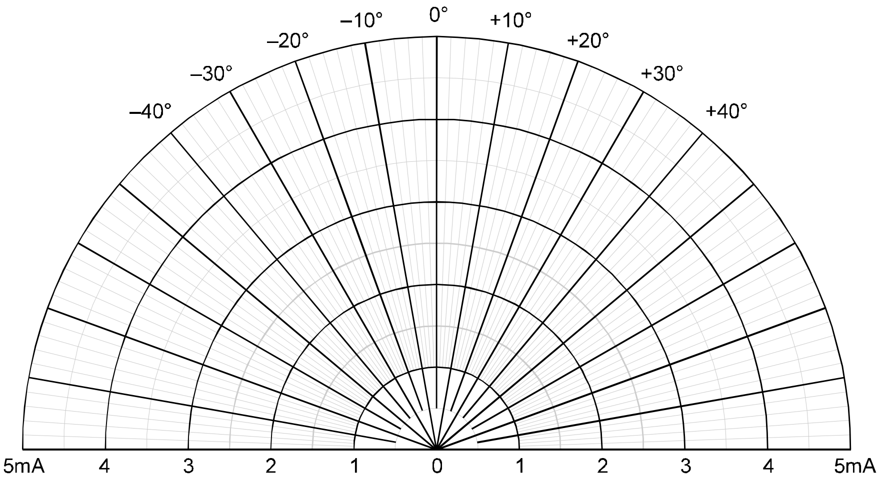


Figure 2-5-5 Polar Graph Paper for Plotting Radiation Diagram

6. Interchange transmitter and receiver antennas so their respective roles are changed and repeat the experiment to obtain the radiation diagram of the second antenna. Compare the two diagrams and their 3 dB beamwidth.

7. The gain of an aperture type antenna is given by the formula:

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

where η = aperture efficiency

A = aperture area

Assuming that the aperture efficiency of the horn antenna is 60%, estimate the gain of the horns used in the experiment at the 10.425 GHz source frequency.

8. The gain of the antennas may be measured experimentally by determining:

P_T = total power transmitted

P_R = power received

or the ratio P_R/P_T

and utilising the formula:

$$P_R = P_T \; G_T \times \left(\frac{\lambda}{4\pi r}\right)^2 G_R$$

where G_T, G_R = gain of transmitter, receiver antennas which in our case can be assumed equal.

Outline how you would measure gain G of the horn antennas and calculate G for the case

P_T = 2.0 mW

P_R = 0.05 mW

r = 0.5 m

f = 10.425 GHz

SUMMARY

The radiation diagram of horn antennas has been investigated experimentally using a basic microwave test bench. The polar radiation diagram has been plotted and the 3 dB beamwidth of the antenna determined. The measurements taken were in the H-plane and indicated the horn antenna to have a directive radiation pattern.

Notes

Angular direction Q°	Detector I mA	Angular direction Q°	Detector I mA
0°	4.0	0°	4.0
+ 10°	3.1	- 10°	3.1
+ 20°	1.4	- 20°	1.5
+ 30°	0.38	- 30°	0.4
40°	0	- 40°	0

Specimen Results

The radiation diagram is plotted in Figure 2-5-6

Half power points occur at -18° and +16°

so 3-dB beamwidth = 18 + 16 = 34°

Gain: dimension of horn aperture are:

7.3 cm x 4.8 cm

so Area A = 7.3 x 4.8 = 35.0 cm^2

and l = 3 x 10^8/10.425 x 10^9

= 0.0288 m = 2.88 cm at f = 10.425 GHz

Hence estimate of gain using

$$G = \eta \frac{4\pi}{\lambda^2} A \quad \text{with } \eta = 0.60$$

$$= 0.6 \times 4\pi \times \frac{35}{2.88^2} = 31.8$$

or in dB 10 log G = 15.0 dBi

Gain using formula:

$$P_R = P_T \; G_T \times \left(\frac{\lambda}{4\pi r}\right)^2 G_R$$

and assuming $G_T = G_R = G$

provides

$$G = \sqrt{\left(\frac{P_R}{P_T}\right) x \left(\frac{4\pi r}{\lambda}\right)^2}$$

$$= \sqrt{\left(\frac{0.05}{2}\right) x \left(\frac{4\pi \text{ x } 0.5}{0.0288}\right)^2} = 34 \text{ or } 15.4 \text{ dBi}$$

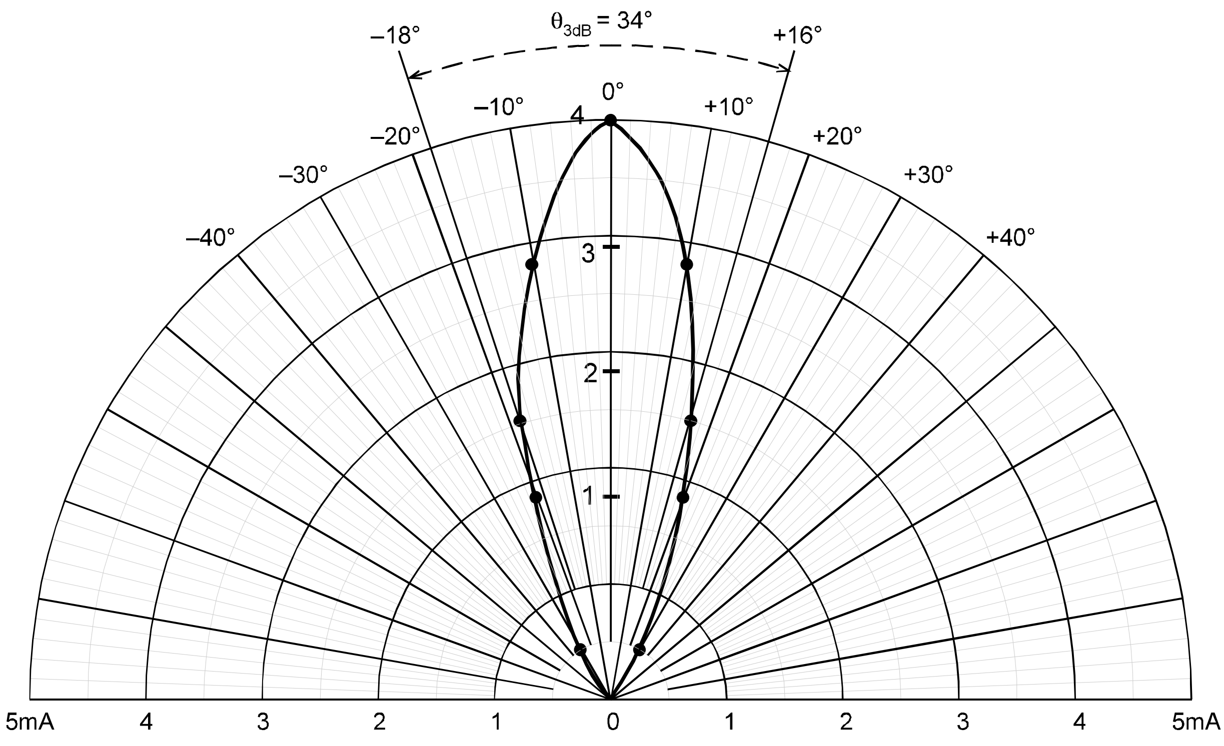


Figure 2-5-6: Radiation Diagram for Horn Antenna

CONTENT

The properties of a waveguide directional coupler are investigated. It is confirmed that microwave power flowing in the forward direction (from source to load) and power flowing in the reverse direction (reflected from the load) may be independently detected using the directional coupler. In the practical assignment, the directional coupler is employed to measure the ratio of forward to reflected power for three different waveguide components and their VSWRs determined from these measurements.

EQUIPMENT REQUIRED

Qty	Identifying Letter	Description
1	-	Control console
2	A	Variable attenuators
1	F	Directional coupler
1	S	Probe diode-detector
1	N	Horn antenna
1	K	Resistive termination
1	R	Short-circuit termination plate

OBJECTIVES

When you have completed this assignment you:

- Will appreciate the properties of directional couplers and their applications in microwave transmission and measurement systems
- Know how a directional coupler may be used to monitor power flowing in forward and reverse direction
- Know how to measure the voltage standing wave ratio (VSWR) of a waveguide component using a directional coupler

KNOWLEDGE LEVEL

Before you start this assignment it would be an advantage:

- To be familiar with the operation of the microwave bench
- To have read and done Assignment 2:- Measurement of Voltage Standing Wave Ratio (VSWR).
- To know that microwave signals can be detected using a diode detector and for low-level signals that the detector output is proportional to power.

INTRODUCTION

A directional coupler consists essentially of two coupled transmission line designed to couple a certain fraction of energy from one line to the other and in a direction dependent on the direction of power flow. These properties are best defined with reference to the diagram of Figure 2-6-1.

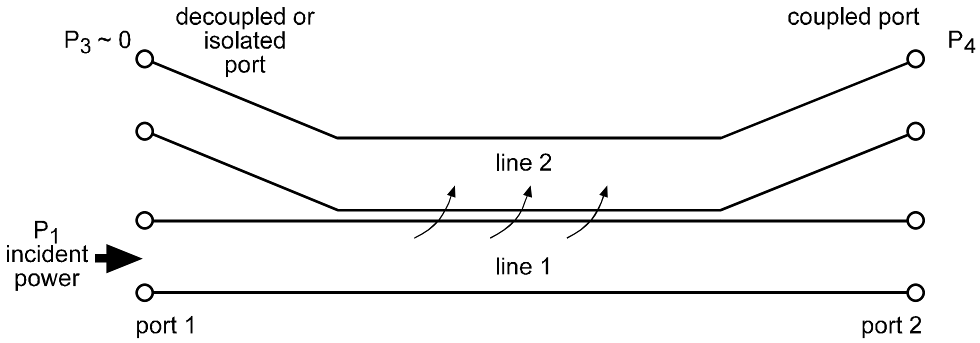


Figure 2-6-1

A fraction of the incident power P_1 entering port 1 and flowing in line 1 is coupled into line 2. The power coupled P_3 emerges at port 3 only with ideally zero power emerging at port 4. The balance of the power P_2 in line 1 emerges at port 2.

The coupling coefficient of the device,

$$C = \frac{\text{power to coupled port}}{\text{incident power}} = \frac{P_3}{P_1}$$

is usually expressed in decibels as

$$10 \log P_1/P_3 \text{ dB}$$

the reciprocal ratio P_1/P_3 rather than P_3/P_1 being used to avoid a negative quantity e.g. if the incident power $P_1 = 10$ mW and the power coupled to port 3, $P_3 = 0.1$ mW:

coupling C = 0.1/10 = 0.01

coupling in dB = 10 log10/0.01 = 20 dB

In practice, there is a small leakage of power to the 'isolated' or 'decoupled' port, port 4. The directional coupling quality of a coupler the directivity is defined in terms of the ration of coupled to decoupled port powers i.e.

$$\text{directivity } D = \frac{\text{power to coupled port}}{\text{power to isolated port}} = \frac{P_3}{P_4}$$

and is usually expressed in decibels,

$$10 \log P_3/P_4$$

Well-designed couplers typically have directivities in excess of 30 dB so that the ration of powers coupled in wanted to unwanted directions is usually better than 1000 in such devices.

Directional couplers find important application in power monitoring, power splitting and in microwave measurement systems. Some examples of their application are illustrated in Figure 2-6-2.

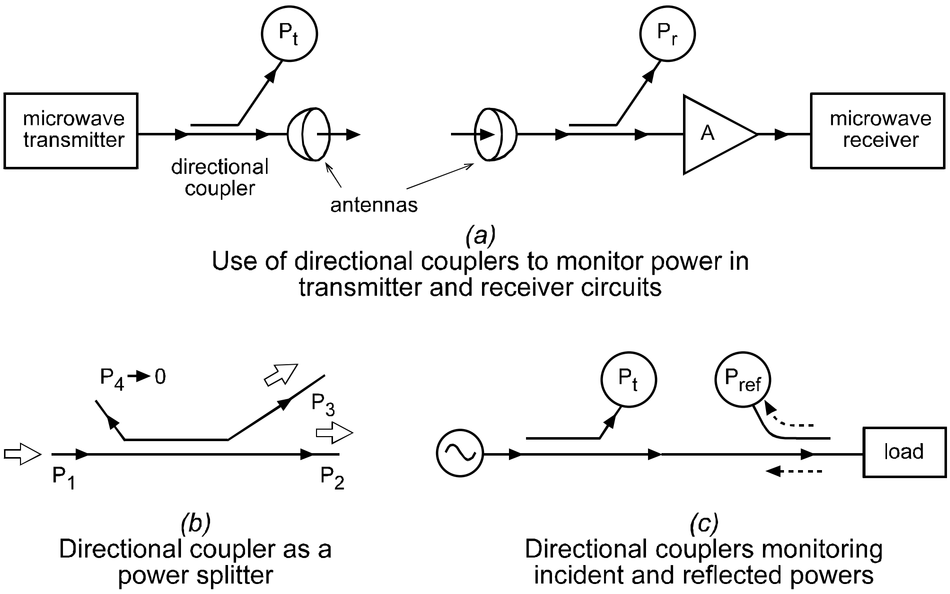
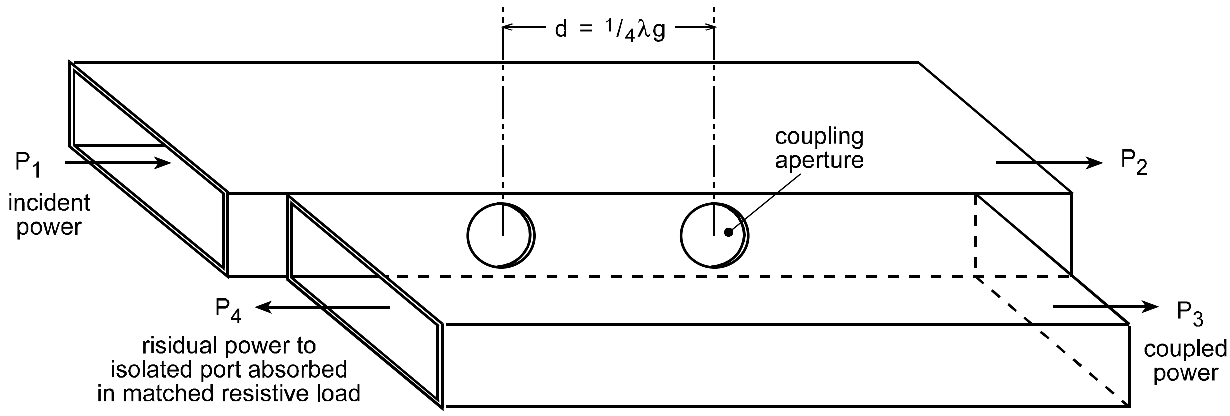
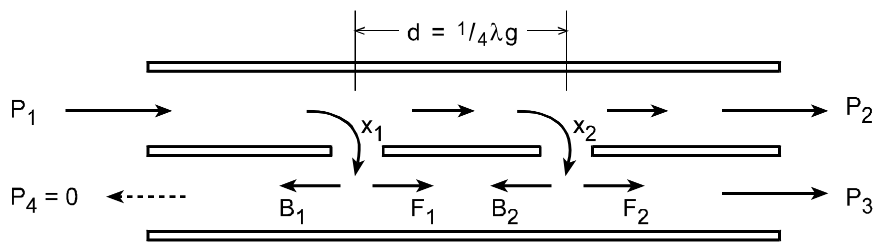


Figure 2-6-2: Directional Coupler Applications

The type of coupler used in this assignment is a side wall waveguide coupler where coupling is produced by cutting apertures in the side wall of two parallel waveguides. Microwave energy is coupled from one guide to the other via these apertures. The strength of coupling is determined by the position and area of the apertures and the directional properties by their axial spacing. A sketch of a simple 2-hole coupler of the type used in the assignment is shown in Figure 2-6-3. Its action is explained in the following paragraph.



(a)



(b)

Figure 2-6-3: Sidewall Coupler

Consider waves incident at port 1 and suppose coupling through the first aperture produces waves of amplitude F_1 and B_1 in forward and reverse directions as indicated in Figure 2-6-3(b). Suppose at the second aperture coupling produces forward and reverse waves of amplitude F_2 and B_2. The components F_1 and F_2 from the apertures in the forward direction emerging at port 3 are in-phase since both have travelled the same distance d. Components B_1 and B_2 travelling in the reverse direction and emerging at port 4, however, differ in phase since B_2 has travelled an additional distance of 2 d when it combines with B_1 at port 4, i.e. from x_1 to x_2 and x_2 to x_1. Since d is designed to be ¼ lg, 2d = ½ lg and hence the components are 180° out of phase and therefore tend to cancel. If the coupling through the apertures is such that $B_1 = B_2$, then the cancellation will be complete and no power will emerge at port 4.

EXPERIMENTAL PROCEDURE

WARNING:

Always avoid looking directly into an antenna or open waveguide components when energised. Microwave radiation can cause harm and eyes are particularly sensitive.

1. Connect up the equipment with the horn antenna as the effective load and with the directional coupler in the forward coupling position as shown in Figure 2-6-4.

The probe diode-detector S detects the power coupled in the forward direction. The depth of penetration can be adjusted by the nut and should be close to maximum.

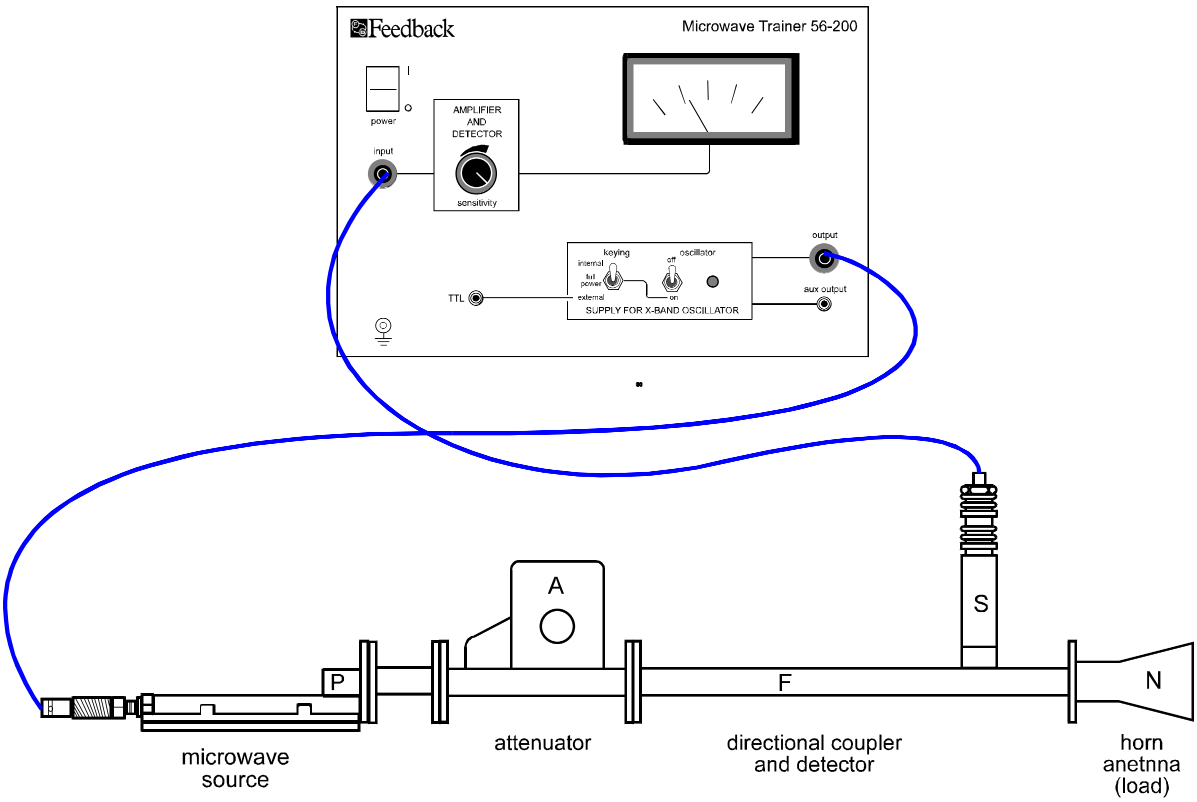


Figure 2-6-4

2. Set the attenuator to approximately 40°, switch the X-band source to internal keying and the meter to detector output. Switch on the console power supply and X-band oscillator source.

3. Adjust the amplifier-detector sensitivity control and if necessary the attenuator to obtain a reasonable (mid-scale) reading on the meter. Record the detector current in a table similar to that given above. This result provides a reference directly proportional to the power incident on the horn antenna load.

Detector meter current, mA	HORN ANTENNA	RESISTIVE TERMINATION	ATTENUATOR PLUS SHORT-CIRCUIT
forward direction, I_f reverse direction, I_r			
power reflection coeff. $P_r/P_i = I_r/I_f = \Gamma^2$			
voltage reflection coefficient, $\Gamma = \sqrt{(I_r/I_f)}$			
VSWR, $S = \dfrac{1 + \Gamma}{1 - \Gamma}$			

Figure 2-6-5: Table to Record Results

4. Next reverse the directional coupler to measure the power reflected from the horn antenna. The directional coupler is now positioned to couple power flowing in the reverse direction. The set-up is shown in Figure 2-6-6.

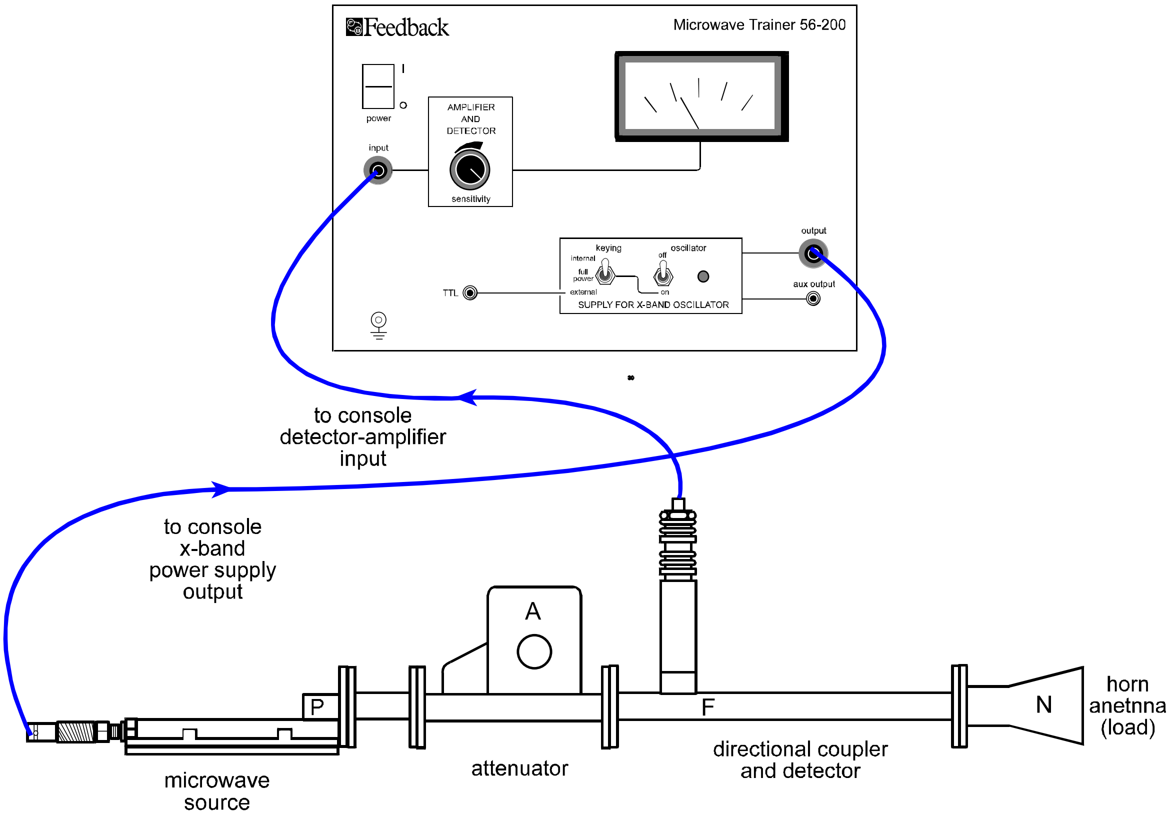


Figure 2-6-6

5. Measure the detector current output for the reverse direction case and record in the table under the HORN ANTENNA section.

6. Disconnect the horn antenna and replace with the resistive load termination. Measure detector current output for both forward and reverse coupling direction as indicated in 3, 4, 5 above.

7. Repeat with the above measurements for the case of a highly reflective load using a variable attenuator set at 80° (very low attenuation) terminated by a short circuiting plate.

8. For each of the 3 loads work out the following:

power reflection coefficient, $\frac{P_r}{P_i} = \frac{I_r}{I_f} = \Gamma^2$

voltage reflection coefficient, $\Gamma = \sqrt{\frac{P_r}{P_i}} = \sqrt{\frac{I_r}{I_f}}$

voltage standing wave ratio, VSWR S = $\frac{1+\Gamma}{1-\Gamma}$

and record your results in the table

10. Comment on the results obtained for each component as regards the degree of matching it presents to its waveguide feed line.

SUMMARY

The directional properties of a waveguide coupler have been investigated by measuring the ratio of powers coupled in forward and reverse directions for 3 different components. From these results, the VSWRs of the components were determined. Applications of directional couplers and construction for a typical coupler have also been considered.

Results for the 3 components are summarised in the table below:

horn antenna N

resistive load termination K

attenuator A set at 80° and short-circuited at far end with plate R

Detector meter current	HORN ANTENNA	RESISTIVE LOAD	ATTENUATOR AT 80° + S/C
forward direction, I_f reverse direction, I_r	2 mA 0.1 mA	2.4 mA 0.2 mA	2.0 mA 1.6 mA
power reflection coefficient, $P_r/P_i = I_r/I_f = [\,\Gamma\,]^2$	0.1/2 = 0.05	0.2/2.4 = 0.083	1.6/2.0 = 0.8
voltage reflection coefficient, $[\,\Gamma\,]$	$\sqrt{0.05} = 0.224$	$\sqrt{0.083} = 0.29$	$\sqrt{0.8} = 0.89$
VSWR, $S = \dfrac{1 + \lvert\Gamma\rvert}{1 - \lvert\Gamma\rvert}$	1.6	1.8	17

NOTE: we assume the diode-detector used in the directional coupler operates in its square-law range so:

I_f is directly proportional to the forward or incident power P_i

I_r is directly proportional to the reflected (reverse) power P_r

The results show that the horn antenna presents quite a good match with a VSWR of 1.6 and only 5% of the incident power reflected; the resistive load presents a fair match but this could be improved by better design (good quality terminations have a VSWR better than 1.05); the attenuator is set at a very low loss and with the short-circuit presents a highly reflective load, confirmed by a very high VSWR.

Notes

CONTENT

The properties and applications of three important waveguide junctions: the H plane, the E-plane and hybrid Tees, are investigated.

EQUIPMENT REQUIRED

Qty	Identifying Letter	Description
1	-	Control console
1	A	Variable attenuator
1	B	Slotted line
1	S	Probe diode-detector unit
1	P	X-band oscillator
1	M	Waveguide mounted diode detector
1	K	Resistive termination
1	E	H plane or shunt Tee
1	G	E plane or series Tee
1	H	Hybrid Tee
2	N	Horn antennas

OBJECTIVES

When you have completed this assignment you will

- Appreciate the properties and applications of the three basic wavegeuide junctions: shunt, series and hybrid Tees
- Have investigated experimentally their power dividing properties

KNOWLEDGE LEVEL

Before your start this assignment it would be an advantage

- To be familiar with the operation of the microwave bench by at least having completed Assignments 1 and 2
- To know that microwave signals can be detected using a diode detector
- To know how VSWR may be measured using a slotted line

INTRODUCTION

In many microwave system applications it is required to divide a waveguide into two branches or conversely combine two waveguides into a single line.

This may be accomplished in rectangular waveguide by using either a shunt-type H-plane or series-type E-plane Tee junction; the form of the junctions and their equivalent circuits are shown in Figure 2-7-1. The H-plane junction is so called because all 3 branches are in the plane of the magnetic field of the H_{10} mode; the E-plane since the 3 branches lie in the electric field of the H_{10} mode.

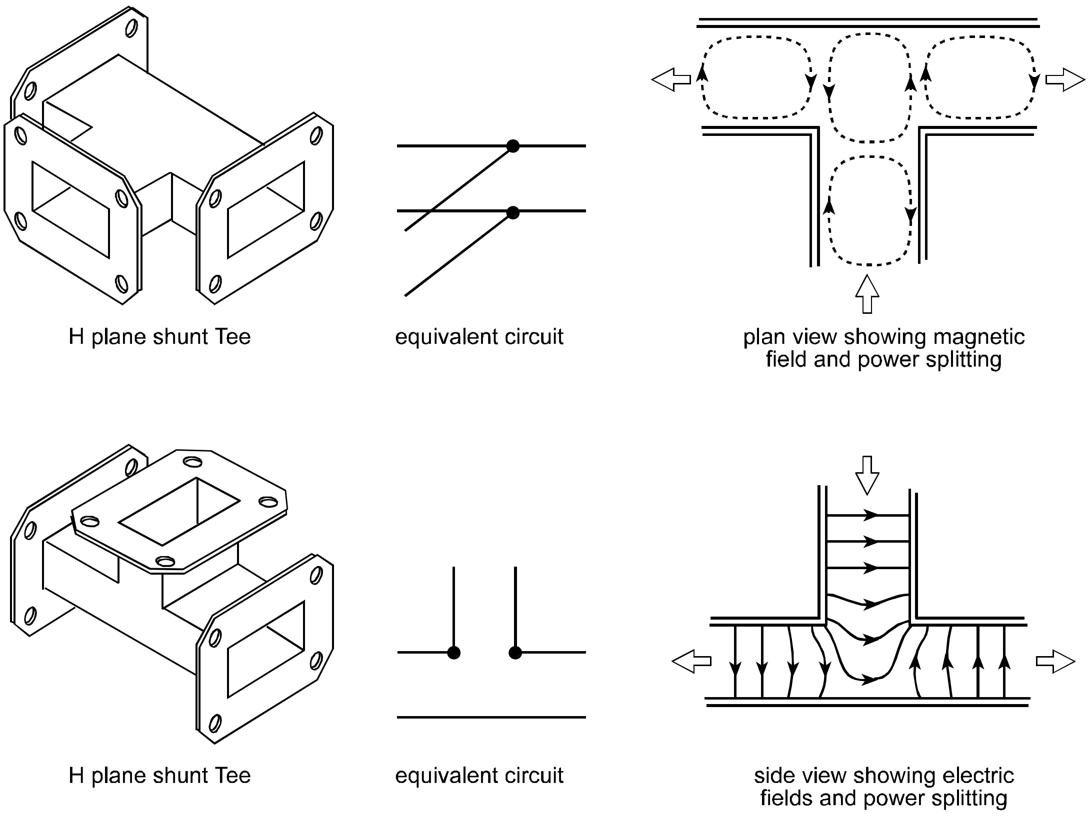


Figure 2-7-1

Power incident in the single line branch divides equally between the other two branches with waves in-phase in the case of the H-plane Tee but in anti-phase for the E-plane Tee. The two shunt lines present impedances in parallel at the junction which combine to give an impedance one-half the single guide impedance. In the E-plane case the input impedance at the junction seen by the single line is two impedances in series. Thus there will be a mismatch for both types of junction and in practice these are compensated by matching elements.

In addition to power splitting/power dividing applications, Tee junctions are used as shunt and series stub matching elements and in radar applications as microwave switches.

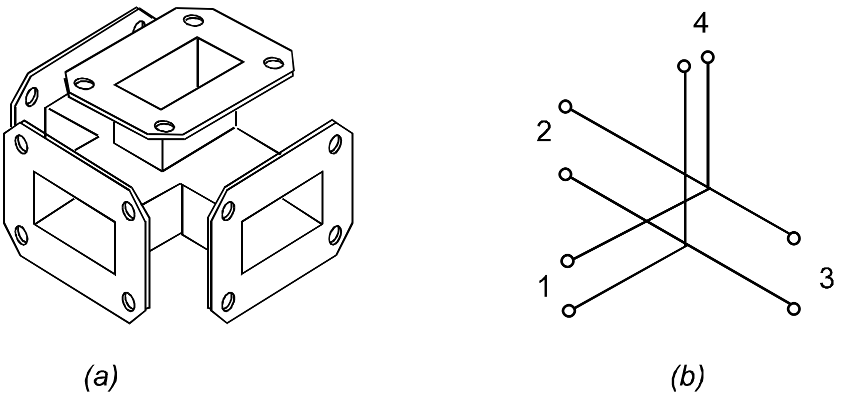


Figure 2-7-2

The hybrid-Tee junction shown in Figure 2-7-2 is essentially a combination of a shunt and series Tee junctions. The hybrid-Tee, assuming each branch is matched, has the property that power incident in any branch divides equally between the two adjacent branches but with no power coupled to the opposite branch. For example, referring to the hybrid-Tee diagram of Figure 2-7-2(b):

(i) power incident at 1 divides equally to 2 and 3; no power to 4

(ii) power incident at 4 divides equally to 2 and 3; no power to 1

(iii) power incident at 2 divides equally to 3 and 4; no power to 2

The hybrid-Tee finds important application as a duplexer for common use of a single antenna in radars, as a balanced mixer in receivers and as a bridge in microwave measurements.

EXPERIMENTAL PROCEDURE

WARNING:

Microwave radiation can be harmful, especially to eyes. **NEVER** look into an energised waveguide.

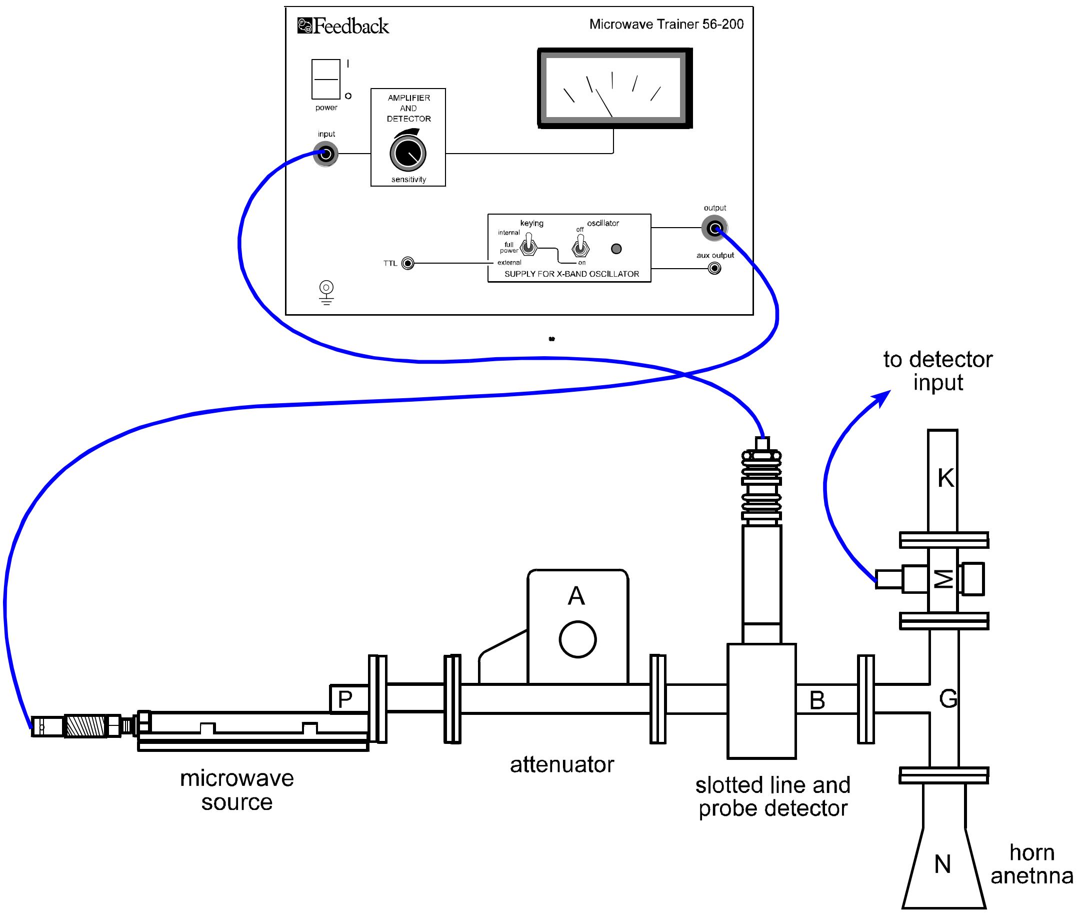


Figure 2-7-3

E- and H-plane investigations

1. Connect up the equipment as shown in Figure 2-7-3. The slotted line and probe-diode detector is to be used to measure the VSWR in the input waveguide to the Tee, the horn antenna is connected in one Tee arm to act as a good-quality load whilst the power in the other Tee arm is monitored by the diode-detector terminated in the resistive load.

2. Set the attenuator to approximately 60° (low attenuation), switch the X-band source to internal keying and meter to detector output. Switch on the console power supply and X-band oscillator source.

3. First measure the VSWR in the input line to the Tee using the slotted line and with the probe detector unit output connected to the console detector input. Move the probe diode-detector unit to locate a maximum of electric field. Adjust the detector-amplifier sensitivity on the console to obtain a mid-to-full scale reading. Record the meter current I_{max}. Next locate an electric field minimum and record the detect current I_{min}.

4. Disconnect the lead from the slotted line probe detector and connect to the waveguide diode detector in the left-hand Tee arm. Measure the detector current and record its value I_{LH}.

5. Interchange positions of horn antenna and diode-detector plus resistive termination and record the detector current I_{RH}.

6. Assuming the detector to be operating in its square-law range, I_{LH} and I_{RH} are directly proportional to the power flowing in the left-hand and right-hand arms of the junction. Thus the tracking performance of the Tee junction can be evaluated. Tracking in this context means the degree in which power splits equally into the two arms.

7. Repeat for the second Tee junction. Results for both Tees can be recorded in a table similar to that given in Figure 2-7-4.

8. Evaluate for each Tee:

   Voltage standing rations (VSWR);

   Reflection coefficient, Γ;

   Tracking ratio: power LH arm/power RH arm

Measurement Results	E-plane	H-plane
VSWR measurements detector current at		
maximum I_{max}		
minimum I_{min}		
$\text{VSWR, S} = \sqrt{\frac{I_{max}}{I_{min}}}$		
reflection coefficient, Γ = S - 1/S + 1		
Tracking: detector current		
Left-hand arm I_{LH}		
Right-hand arm I_{RH} Ratio of powers I_{LH}/I_{RH}		

Figure 2-7-4 Table for Recording Results of E- and H-Plane Tees

HYBRID-TEE INVESTIGATED

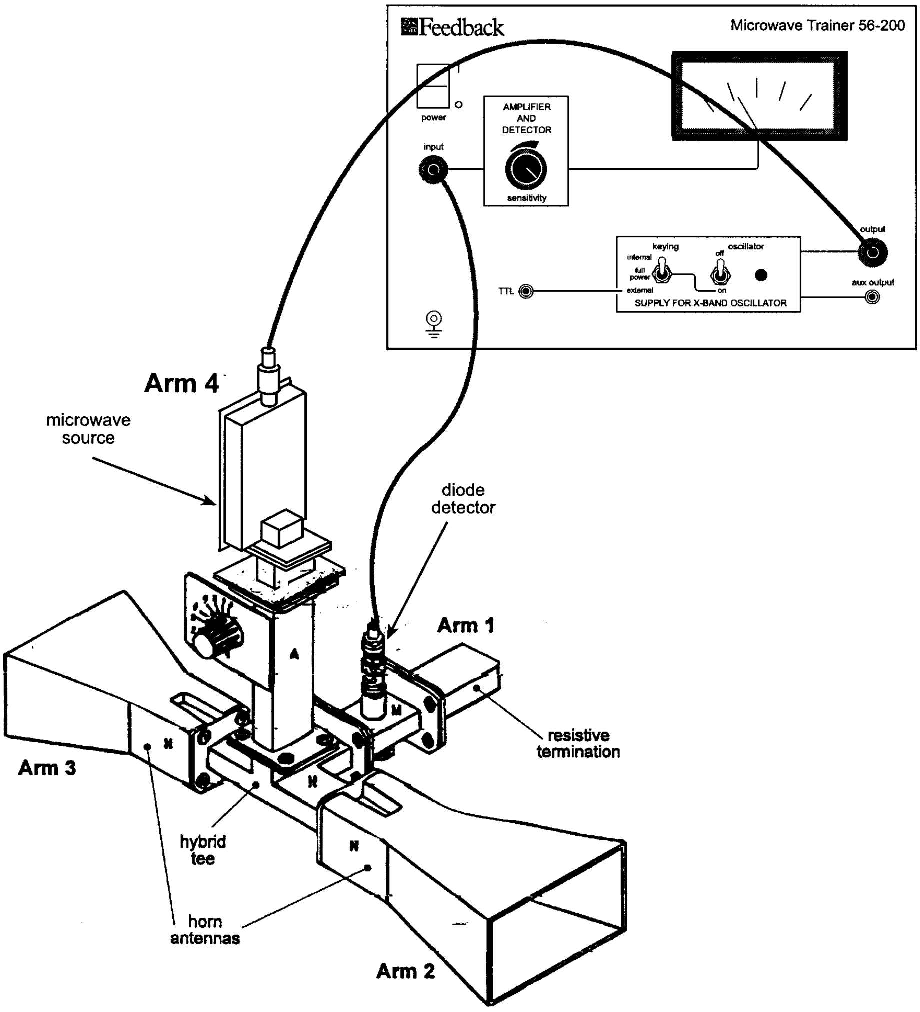


Figure 2-7-5

1. Set up the equipment as shown in Figure 2-7-5 with the attenuation set at about 60° producing low attenuation. Switch to internal keying if not already set and switch on power supply and the X-band oscillator source.

2. Measure the detector current in the position shown. Note this will be zero or very close to zero since the detector is located in the isolated arm of the hybrid-Tee, that is the supply ARM 4 and ARM 1 containing the detector are decoupled.

3. Next investigate the power supplied to ARMS 2 and 3 which currently supply the two horn loads. Disconnect the horn in ARM 2 and interchange with the detector and resistive termination so in the new set-up ARM 2 is terminated in a horn and ARM 2 has the detector to measure power. Record the detector output current I_2, which assuming square-law detecting operation will proportional to the power P_2 in ARM 2.

4. Interchange detector in ARM 2 and horn in ARM 3 to measure detector current I_3, which is directly proportional to the power in ARM 3, P_3.

5. Return system to the original set-up with the horns termination ARMS 2 and 3 and insert the slotted line and diode-probe detector unit in ARM 4. Measure the VSWR by recording detector current at positions of an electric field maximum and minimum.

SUMMARY

Three basic waveguide junctions; shunt, series and hybrid Tees have been investigated experimentally,. The shunt H-plane and series E-plane junctions have been shown to divide input equally to the two output branches.

The properties of the hybrid-Tee in dividing power and in presenting isolation between two branches are also confirmed.

H-plane Tee

Input VSWR measurements line 1:

$$I_{max} = 1.4\,\text{mA} \quad I_{min} = 0.2\,\text{mA}$$

$$\text{VSWR, } S = \sqrt{\frac{I_{max}}{I_{min}}} = \sqrt{\frac{1.4}{0.2}} = 2.6$$

$$\text{reflection coefficient} \quad \Gamma = \frac{S-1}{S+1} = 0.44$$

Power tracking measurements:

Port input to single line branch, port 1

Power output to RH branch, port 2 : I_2 = 2.8 mA

Power output to LH branch, port 3 : I_3 = 2.8 mA

showing 1 : 1 tracking

E-Plane Tee

Input VSWR measurements line 1:

$$I_{max} = 1.4 \qquad I_{min} = 0.4$$

$$\text{VSWR, } S = \sqrt{\frac{I_{max}}{I_{min}}} = \sqrt{\frac{1.4}{0.4}} = 1.9$$

$$\text{reflection coefficient, } \quad \Gamma = \frac{S-1}{S+1} = 0.31$$

Hybrid Tee

Input arm 4 VSWR » 14 = 3.7

Reflection coefficient Γ = 0.58

Ratio of power to arms 2 and 3:

I_2 = 3.0 mA, I_3 = 3.0 mA

$P_2/P_3 = 1$

Results indicate good tracking in all 3 cases but poor matching with relatively high VSWRs.

Notes

CONTENT

The means of transferring microwave power from waveguide to coaxial line by coaxial-to-waveguide transformer is investigated experimentally.

EQUIPMENT REQUIRED

Qty	Identifying letter	Description
1	-	Control console
1	A	Variable attenuator
1	B	Slotted line
1	S	Probe diode-detector
1	P	X-band oscillator
2	J	Waveguide-to-coaxial transformer
1	M	Waveguide mount diode-detector unit
1	K	Resistive termination
1	C	Slotted-line probe tuner
1	-	Length of coaxial cable with N-type connectors

OBJECTIVES

When you have completed this assignment you will:

- Appreciate that both coaxial lines and rectangular waveguides are two primary lines used in transporting microwave signals
- Know how power transfer between waveguides and coaxial lines may be achieved
- Investigated the performance of a practical waveguide-to-coaxial transformer
- Appreciate the need for matching in microwave systems for maximum power transfer

KNOWLEDGE LEVEL

Before you start this assignment it would be an advantage:

- To be familiar with the operation of the microwave bench by, at least, having completed Assignments 1 and 2
- To know that microwave signals can be detected using a diode detector
- To know how VSWR may be measured using a slotted line, see Assignment 2
- To appreciate the need for matching and how this may be achieved in waveguide, see Assignment 4

INTRODUCTION

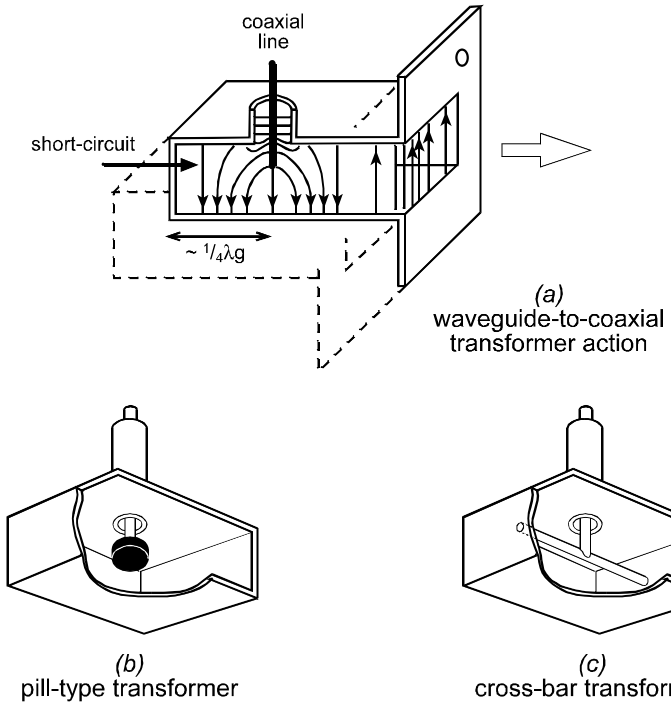


Figure 2-8-1

Figure 2.8.1(a) shows a cut-away sketch illustrating the means of coupling power form a coaxial line to a waveguide. The inner conductor of the coaxial line penetrates into the waveguide and acts as a probe antenna radiating energy into the guide. The probe is situated $\frac{1}{4}l_g$ from the short-circuit end to provide very high impedance conditions for 'unwanted' power travelling to the left.

Two practical devices are shown in Figure 2-8-1(b) and (c). With effective matching waveguide-to coaxial transformers can provide excellent power transfer with VSWRs better than 1.5 over a complete frequency range, e.g. over X-band, 8.5 - 12 GHz, relatively inexpensive transformers are available with VSWR as low as 1.06 on centre-band but better than 1.5 over the whole band.

EXPERIMENTAL PROCEDURE

WARNING:

Microwave radiation can be harmful, especially to eyes. **NEVER** look into an energised waveguide.

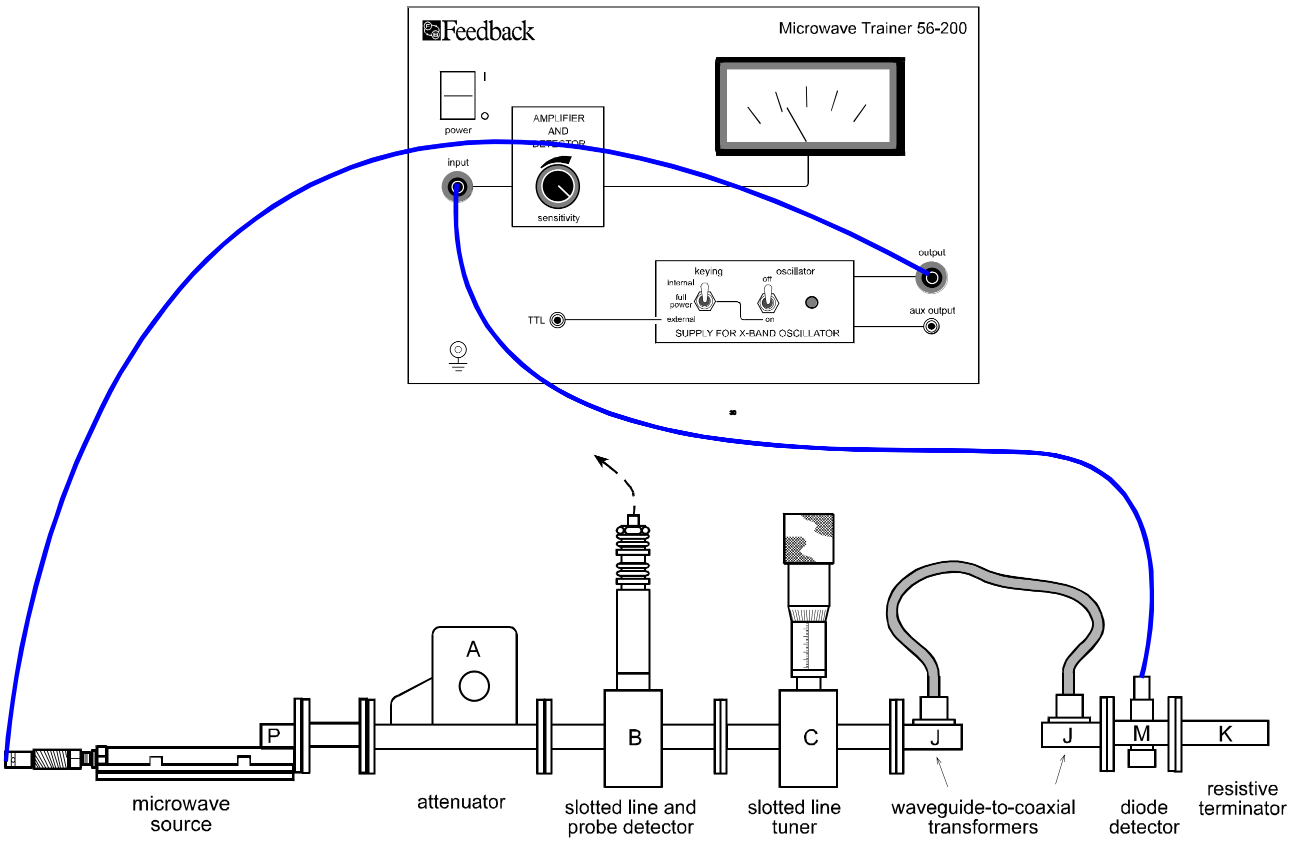


Figure 2-8-2

1. Set up the equipment as shown in Figure 2-8-2. The slotted line and probe detector unit is to be used to measure the VSWR and the slotted line probe tuner to match the transformer. Ensure initially that the probe position in the tuner is fully withdrawn, beyond 25 mm on the micrometer scale.

   Connect X-band source to console power supply and check that the diode-detector M monitoring the output power from the second transformer is connected via its measuring lead to the detector input on the console.

2. Set the attenuator to approximately 50° and the sensitivity of the detector-amplifier on the console to mid-position. Switch the X-band source to internal keying and the meter to detector output.

   Switch on the console power supply and the X-band oscillator source.

3. Microwave power from the source enters the first waveguide-to-coaxial transformer to travel down the coaxial cable length and to re-enter the waveguide via the second transition. The output power is detected by the diode detector M. Record the output detector current.

   Note:

   This may be a relatively low value of the order of 0.5 mA at mid-range sensitivity of the detector-amplifier control on the console due to high reflection losses.

4. Transfer the measuring cable lead to the probe-detector on the slotted line and measure the VSWR by measuring detector current I_{max} at an electric field maximum and the current I_{min} at an electric field minimum.

   emember, VSWR $S = \sqrt{\dfrac{I_{max}}{I_{min}}}$

   Note:

   Without matching the VSWR is a high value indicating the transformer presents a poor match.

5. Place the probe of the slotted tuner at a position d = 20mm referred to the tuner scale. Use this position, which was determined from following the impedance matching procedure in Assignment 4, or if time permits carry out your own measurements to determine d.

Change the measuring cable lead back to diode-detector M to monitor output power as tuning is effected.

6. Screw in the probe of the tuner at position d and observe the effect on output power as measured by the output of diode-detector M.

At first no change will occur but at scale readings less than 16 mm where the penetration of the probe within the guide is considerable, a sharp increase should be observed. Adjust probe penetration carefully to obtain maximum power output and record corresponding detector current.

7. Transfer coaxial measuring lead back to probe detector on slotted line and re-measure VSWR.

8. Determine the new value of VSWR after matching. Compare the percentage of power transferred by the transformers prior to and after matching was effected.

SUMMARY

The performance of waveguide-to-coaxial transformers has been investigated experimentally. The effect of matching shows power transfer can be dramatically increased.

Case 1: no matching, probe of slotted line tuner fully withdrawn

Detector current output, $I_1 = 0.5$ mA

so power P_1 µ 0.5

VSWR $S = \sqrt{2.0/0.04} = = \sqrt{50} = 7.0$

Reflection coefficient, $\Gamma = S\text{-}1/S\text{+}1 \approx 0.86$

Percentage incident power transferred, $(1 - \Gamma^2) \times 100\% = 26\%$

Case 2: with matching

Detector current output, $I_2 = 2.3$ mA

P_2 µ 2.3

VSWR, $S = \sqrt{1.6/1.05} = 1.23$

$\Gamma = 0.10$

Percentage of power transmitted, $P_2 = 99\%$

Ratio $P_1/P_2 = I_1/I_2 = 0.5/2.3 = 0.22$

CONCLUSION

The coaxial-to-waveguide transitions in the Trainer Kit transfer microwave power from coaxial line to waveguide by simple probe coupling. Their efficiency, however, is poor as exhibited by their very high VSWR. A high proportion of incident energy is reflected. By matching using a probe-tuner, power transfer can be greatly improved.

CONTENT

A basic microwave radio link is set up and its transmission characteristics are investigated.

EQUIPMENT REQUIRED

Qty	Identifying Letter	Description
1	-	Control Console
1	A	Variable attenuator
1	B	Slotted line
1	S	Probe diode-detector
1	P	X-band oscillator source
2	N	Horn antennas
1	M	Waveguide mount diode-detector
1	K	Resistive terminator
4	-	Supports
1	-	Metal plate (optional)

OBJECTIVES

When you have completed this assignment you will:

- Appreciate the main factors determining power received in a microwave radio line-of-sight link.
- Investigated the operation of a radio link and seen the effect of link distance, polarisation and reflection interference on the level of power received.
- Know the importance of microwave radio links as long distance communication carriers.

KNOWLEDGE LEVEL

Before you start this assignment it would be an advantage:

- To be familiar with the components in the Microwave Trainer.
- To have completed Assignments 2 (Measurement of VSWR) and Assignment 5 (Radiation diagrams of microwave horn antennas).
- To know how VSWR is measured and how microwave signals are detected.

INTRODUCTION

Microwave line-of-sight radio links are extensively used in long distance communication networks and currently cater with satellite communications for some 40% of the global traffic. Terrestrial microwave radio and satellite networks form complementary systems with optical fibre networks with the former possessing advantages of flexibility in installation and coverage. A typical microwave radio network is shown diagrammatically in Figure 2-9-1.

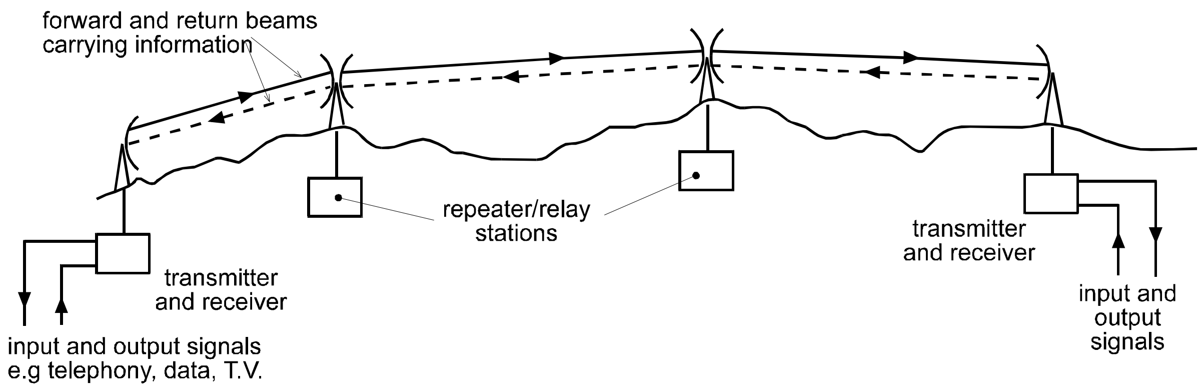


Figure 2-9-1

The power received in a line-of-sight radio link (see also Assignment 5) can be expressed as:

$$P_R \quad G_T \text{ x } (\lambda / 4\pi r)^2 \text{ x } G_R / L$$

where P_T = transmitter radiated power

G_T = gain of transmitter antenna

G_R = gain of receiver antenna

l = c/f = wavelength

f = frequency

c = velocity of e.m. waves = 3×10^8 m/s

L = system losses (feed, atmospheric losses)

This formula is extremely valuable in power budget calculations. The formula may also be express in decibels (dB and decibel-watts dBW or decibel-milliwatts dBm):

P_R dBW or dBm

= P_T dBW or dBm + G_T dB + G_R dB - L dB - FSL dB

where FSL = 20 log (4pr/l) dB is known in the free space loss.

In practice, transmitter power P_T are in the order of 0.5 to 1 W. Antennas are parabolic reflectors fed by horn antenna. Diameters are typically 1 to 3 m and antenna gains are usually in the range 30 to 50 dB. Individual link distances vary but seldom exceed 80 km. The free space loss term contributes the largest loss term by far with values up to 150 plus dB.

Example for a typical link

P_T = 0.5 W or -3 dBW

f = 10 GHz, l = 0.03 m

G_T = G_R = 35 dB

r = 40 km

FSL = 20 log (4pr/l) = 144.5 dB

L = 3.5 dB (feed losses = 3 dB, atmospheric 0.5 dB)

so P_R = -3 + 35 + 35 - 3.5 –144.5

= -81 dBW or 7.9 x 10^{-9} W

Typical received power are the order of nano-watts or greater but with threshold levels down to a few pico-watts.

EXPERIMENTAL PROCEDURE

WARNING:

Do **NOT** look directly into the transmitter antenna or any energised waveguide component. Although the power levels produced by the Trainer are low (below 10 mW), microwave radiation can cause harm and eyes are particularly sensitive.

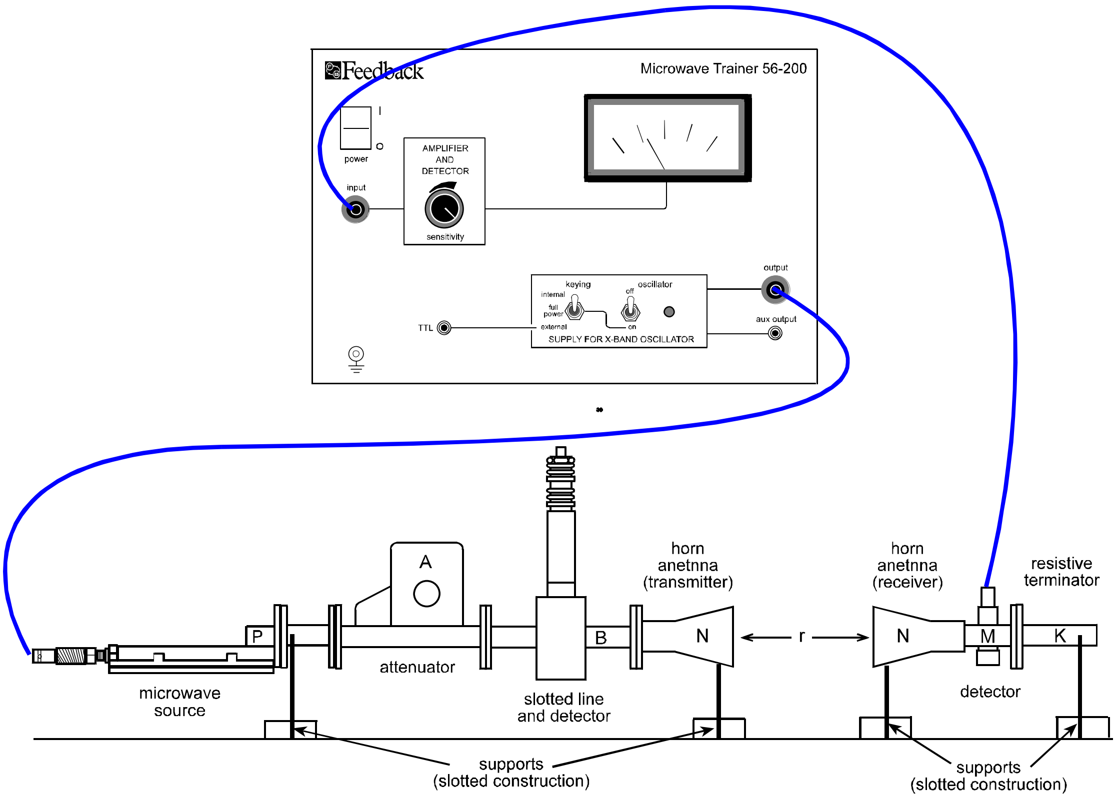


Figure 2-9-2

1. Connect up the equipment as shown in Figure 2-9-2 with the horns in line-of-sight and spaced 35 cm apart. The assemblies should be mounted on the supports, as shown. Ensure that the radio path between the two antennas and the surrounds are free from obstacles, particularly metal, which could cause reflections and upset measurements. Note r = 35 cm is close to the minimum distance for far field operation; virtually all communication applications operate in the far field radiation zones.

2. Switch the X-band source to internal keying. This essentially acts to switch on and off the microwave signal and acts in our case of a microwave link as a crude source of modulation. In practical link modulation would occur at a lower frequency. The baseband signal would modulate an intermediate frequency signal and this modulated signal would be attached to a microwave carrier in an up-converter mixer circuit.

3. Set the attenuator to 40° to provide a modest amount of attenuation; switch the meter on the console to detector output and adjust the sensitivity of the detector-amplifier to mid-position.

Switch on the console power supply and the X-band oscillator source. Adjust attenuator and sensitivity to obtain a meter reading close to full scale deflection.

4. Record detector meter current (which assuming square law detector operation, is directly proportional to received power) and link distance r in a table as given below.

Link distance r metre	Detector output current, I mA (proportional to received power P_R)
0.35 m (35 cm)	
0.40 m	
0.45 m	
0.50 m	
0.55 m	
0.60 m	
0.65 m	
0.70 m	
0.75 m	

5. Repeat measurements as the link distance r is increased in 5 cm steps from r = 35 cm to r = 75 cm.

6. Plot a graph of I representing received power P_R versus r link range to check inverse square law relationship predicted by received power budget formula.

7. Reset equipment to a mid-range position say with link range r = 40 cm.

8. Measure input VSWR to test the match of the horn antenna using the slotted line and probe diode detector. This may be accomplished by measuring the detector output current on the console meter at a position of electric field maximum, I_{max}, and at an electric field minimum, I_{min}, (see Assignment 2).

$$\text{VSWR} \quad S = \sqrt{\frac{I_{max}}{I_{min}}}$$

Calculate the percentage of incident power reflected.

9. Multipath interference due to ground reflection may be simulated by placing a metal plate (typical size 15 x 10 cm) as indicated in Figure 2-9-3.

Record the variation of detector current measured by diode detector M as the vertical distance y is varied.

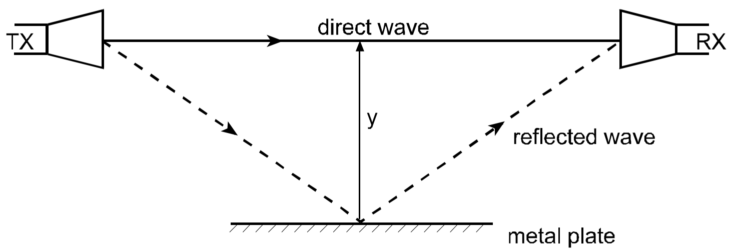


Figure 2-9-3: Interference due to Reflection

10. The microwave signal radiated is polarised with its electric field vector parallel to the narrow waveguide and horn dimensions. The receiving horn must be correctly aligned to receive this polarisation and so ensure transfer of the signal into the receiving waveguide by exciting the H_{10} mode with the correct direction of the electric field vector as indicated in Figure 2-9-4.

Test the effect of rotating the receive section on the received signal strength as evidenced by the diode detector M current output. At 90° orientation the received signal strength should be zero.

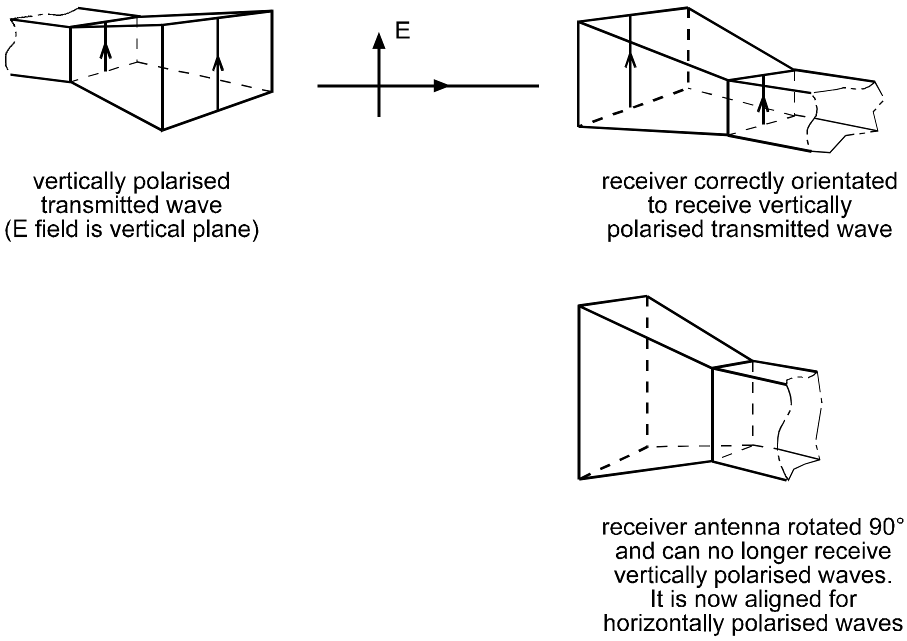


Figure 2-9-4: Alignment of Antennas for Vertical and Horizontal Polarisation of Waves

SUMMARY

A simple microwave radio link has been set up and the power level received versus link distance has been investigated. Interference between direct and reflected waves has been simulated and the received power variation observed as the path difference was varied. The effect of antenna rotation with respect to the polarisation of the transmitted wave was also investigated.

Notes

VSWR measured for a link distance r = 40 cm:

VSWR S = 1.17

reflection coefficient

Γ = S-1/S+1 = 0.08

percentage of power reflected, Γ^2 x 100 = 0.6%

Results for received signal level as measured by detector output current versus link distance r:

Link distance r cm	35	40	45	50	55	60	65	70	75
Output detector (mA) current $I \mu P_R$	2.0	1.5	1.2	1.0	0.7	0.55	0.4	0.3	0.2

A graph I versus r^{-2} , i.e. P_R versus $1/r^2$ gives an approximate straight line fit, indicating that the received power varies inversely with the square at the link distance.

Notes

3 Introduction to Microwaves and their Applications

3.1 Summary

This chapter provides an introduction to the nature of electromagnetic waves focusing on the high frequency end of the radio frequency spectrum. This top end range where the wavelength of the radio waves is small, typically centimetres to millimetres, is known as the microwave range.

The microwave wavelength and frequency ranges are given in relation to the complete electromagnetic spectrum. The characteristics and merits of microwaves are considered together with their major applications in microwave terrestrial radio links, satellite communications, radar, cooking and heating. Subsidiary uses and the future developments into millimetric applications are also noted.

3.2 Microwaves: Frequency Range, Wavelength and Bands

Microwaves correspond to waves at the uppermost end of the radio wave spectrum and derive their name from their very small wavelengths.

Special techniques and devices were required to handle such small wavelengths and these have been developed from the early days of radar through microwave radio and satellite links over the past 60 plus years. Microwave techniques are now employed over a wide range of frequencies which embrace:

frequency:	0.3 GHz to 3000 GHz
where	1 GHz = 1000 MHz = 10^9 Hz
wavelength	100 cm to 0.1 mm

The overall range from the top end of the radio frequency spectrum to the beginning of the far infra-red spectrum is shown in Figure 3-1.

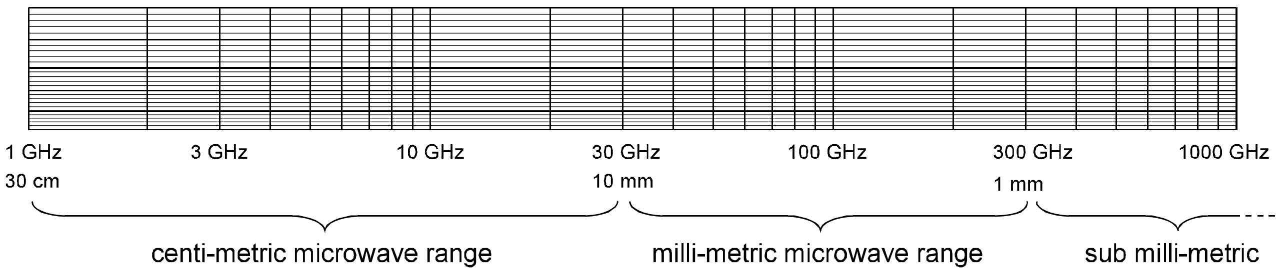


Figure 3-1: Microwave Frequency and Wavelength Range

In general, however, the majority of microwave applications are in the 1 to 30 GHz range with more recent developments into the millimetric range up to about 100 GHz. Thus the term 'microwaves' is normally used to define radio waves in the centimetre and millimetre wavelength ranges:

wavelength : 30 cm to 3.mm

frequency : 1 GHz to 100 GHz

The microwave range is further broken up to a number of sub-bands denoted by letters. This helps to define components, devices, waveguides, specific applications etc... much more conveniently. The designations most commonly used by microwave engineers is given in the table of Figure 3-2.

	Frequency range	**Wavelength range**
L - band	1 –2 GHz	30 –15 cm
S - band	2 –4 GHz	15 - 7.5 cm
C - band	4 –8 GHz	7.5 - 3.75 cm
X - band	8 –12 GHz	3.75 - 2.50 cm
J or K_u - band	12 –18 GHz	25 - 16.67 mm
K - band	18 –27 GHz	16.7 - 11.1 mm
Q or K_a - band	27 –40 GHz	11.1 - 7.5 mm
V - band	40 –60 GHz	7.5 - 5.0 mm
O or E - band	60 –90 GHz	5 - 3.33 mm
W - band	75 –170 GHz	4 - 1.76 mm

Figure 3-2: Letter Designation often used for Microwave Bands

3.3 Electromagnetic waves and their properties

Microwaves are located at the high frequency end of the radio wave spectrum, the latter covering wavelengths from several kilometres to fraction of a millimetre. The radio wave or frequency spectrum is itself a part of a vast spectrum of electromagnetic waves. This family is termed electromagnetic to qualify the fact that all waves in the family carry electromagnetic energy contained in oscillating electric and magnetic fields.

The nature of an electromagnetic wave propagating in free space is shown in Figure 3-3. The wave could be equally a radio wave, a microwave or even light; they are all members of the common electromagnetic family. The wave is composed of electric and magnetic fields oscillating in planes at right angles to each other but with both fields transverse to the direct the wave transports its energy. This type of wave is known as a transverse electromagnetic wave, usually abbreviated to TEM, and is typical of the nature of radio waves or light travelling in free space.

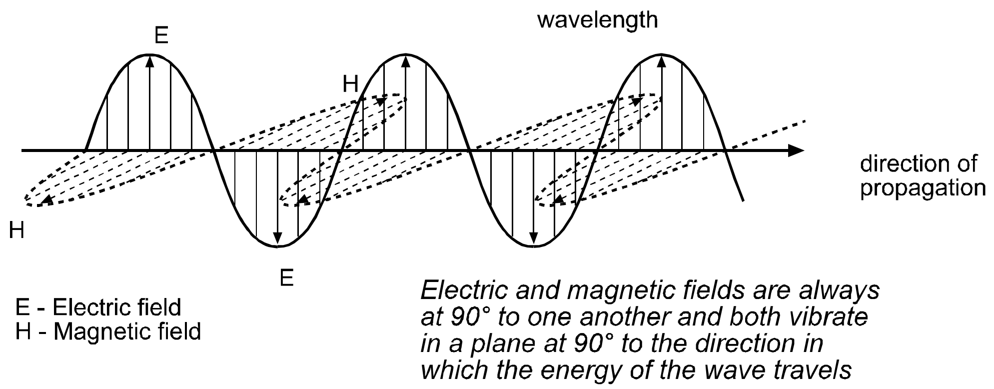


Figure 3-3: Nature of a Plane Electromagnetic Wave in Free Space

Electromagnetic waves propagate in free space at the velocity of light, usually denotes by c:

$\underline{c}$ $= 2.9979 \times 10^8$ metres per second

although c $= 3 \times 10^8$ m/s

$= 186{,}000$ miles per second

are very good approximations.

The energy transported by a TEM wave in terms of power per unit area of the wave is given by:

p $= \frac{1}{2}$ E x H watts per square metre

where E = peak amplitude of electric field component

H = peak amplitude of magnetic field component

These components are also related:

$\sqrt{\varepsilon}\, E = \sqrt{\mu}\, H$

where ε, μ = permittivity, permeability of medium through which waves are travelling

$\varepsilon = \varepsilon_r \varepsilon_o$, $\mu = \mu_r \mu_o$

ε_r = dielectric constant, μ_r = relative permeability

$\varepsilon_o = 8.842 \times 10^{-12}$ F/m, the permittivity of free space

$\mu_o = 4p \times 10^{-7} = 1.256 \times 10^{-6}$ H/m, the permeability of free space

the ratio,

$$\frac{E}{H} = \sqrt{\frac{\mu_r}{\varepsilon_r}} \times \sqrt{\frac{\mu_r}{\varepsilon_r}}$$

$$= 377\sqrt{\frac{\mu_r}{\varepsilon_r}}$$

= 377 for free space as $\varepsilon_r = \mu_r = 1$ in this case.

The velocity of electromagnetic waves is given generally by:

$$v = \frac{c}{\sqrt{\varepsilon_r}} = \frac{c}{n} \text{ m/ s}$$

where $n = \sqrt{\varepsilon_r}$ is the refractive index of the medium.

For air $n = 1.000273$ so:

$$v = \frac{c}{n} = c$$

to an excellent degree of approximation.

The periodic time T, frequency f and wavelength λ are related by:

λ = distance travelled by wavefront in one cycle

= velocity x time = v x T

but $T = 1/f$

so $\lambda = v/f$ and $v = f\lambda$

Thus over the radio frequency range from very low frequencies (VLF) to the limits of the microwave range at 1000 GHz or so, wavelengths vary from 100 km to the order of fractions of a millimetre.

3.4 The Electromagnetic Wave Spectrum

Radio and microwaves form the first part, the lower frequency section, of the vast spectrum of electromagnetic waves. The complete spectrum with frequency and wavelength ranges is shown in Figure 3-4.

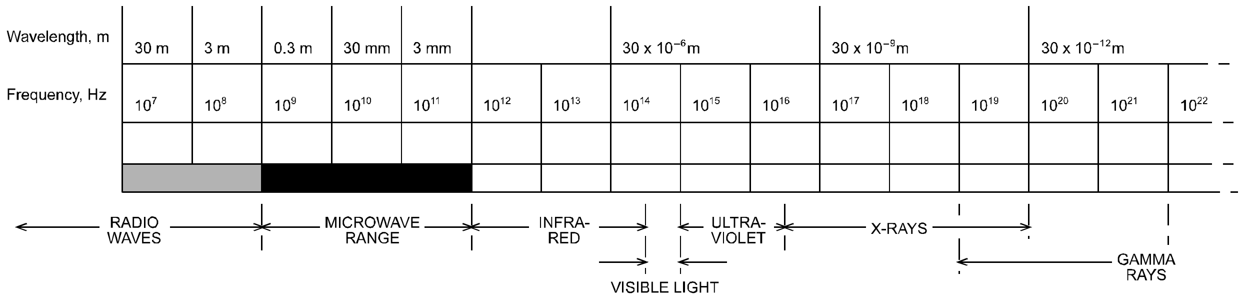


Figure 3-4: Electromagnetic Wave Spectrum

The spectrum is divided into the following groups:

Radio frequency (RF) waves	Frequencies range from a few kHz to 100+GHz; waves propagate with low loss in Earth's atmosphere; microwave frequency range 1 – 100 GHz used extensively for radio communications, radar with special applications to microwave cooking (2.45 GHz), heating and medical thermo-therapy.
Infra-red	Heat radiation waves extending from top end of microwave range to just above visible light range; loss in Earth's atmosphere limits range, although infra-red extensively used in night-vision devices and also in short distance radio communication links; near infra-red with wavelengths in the range of 1.6 - 0.77 mm (mm = 10^{-6} m) used in optical fibre communication systems.
Light	the visible part of the electromagnetic spectrum extending from 700 nm (red) to 400 nm (violet), (1 nm = 10^{-9} m); low-loss propagation in Earth's atmosphere.
Ultra-violet	wavelength range from below 400 nm to about 10 nm; harmful radiation to skin; plays an important role in fluorescence and photo-electronic effect.
X-rays	wavelength below 1 nm to about 0.01 nm; important radiation in medical and material diagnostics.
Gamma rays	top-end frequency part of the electromagnetic spectrum; highly dangerous radiation emitted from radioactive sources and atomic interactions.

3.5 Relatives Merits of Microwaves

Although the history of microwaves can be traced back to well before the 1940's, it was the need for the radio detection of aircraft and shipping in World War 2 that provided the real spur to the development of special techniques and devices to produce the interrogating tools, the centimetric microwave radio beams. The invention of the magnetron, which provided a reliable high-power source of microwaves, was one of the most significant contributions to the development of radar in this period. It is still today perhaps the most widely used microwave device as the power source in microwave ovens for fast food cooking!

Since the 1940's, microwave communications has been developed hand-in-hand with radar, and today microwave terrestrial radio and satellite communication links alongside optical fibre networks form the basis of long distance global telecommunications.

Apart from the heavy demands made on the radio spectrum inevitable pushing use upwards in frequency, what special characteristics or advantage does the microwave range possess? Why has it gained so much prominence in communications, and why is it the range almost exclusively used for radar? The basic reasons are straightforward and are considered below.

3.5.1 Small Wavelength for Efficient Antenna Design

The centimetre and millimetre wavelengths obtained at gigahertz frequencies enable relatively compact, highly efficient antennas to be designed. Microwave radiation can be focused and shaped into well defined patterns ideally suited to the production of highly directive, pencil-type, beams required in microwave radio and Earth station-to-satellite uplinks. Specially shaped beams required for tracking, search, surveillance in radar and spot, zone and global beams required in satellite downlinks are all readily achievable in the microwave range. An example of such beams for a satellite-to-Earth downlink is illustrated in Figure 3-5.

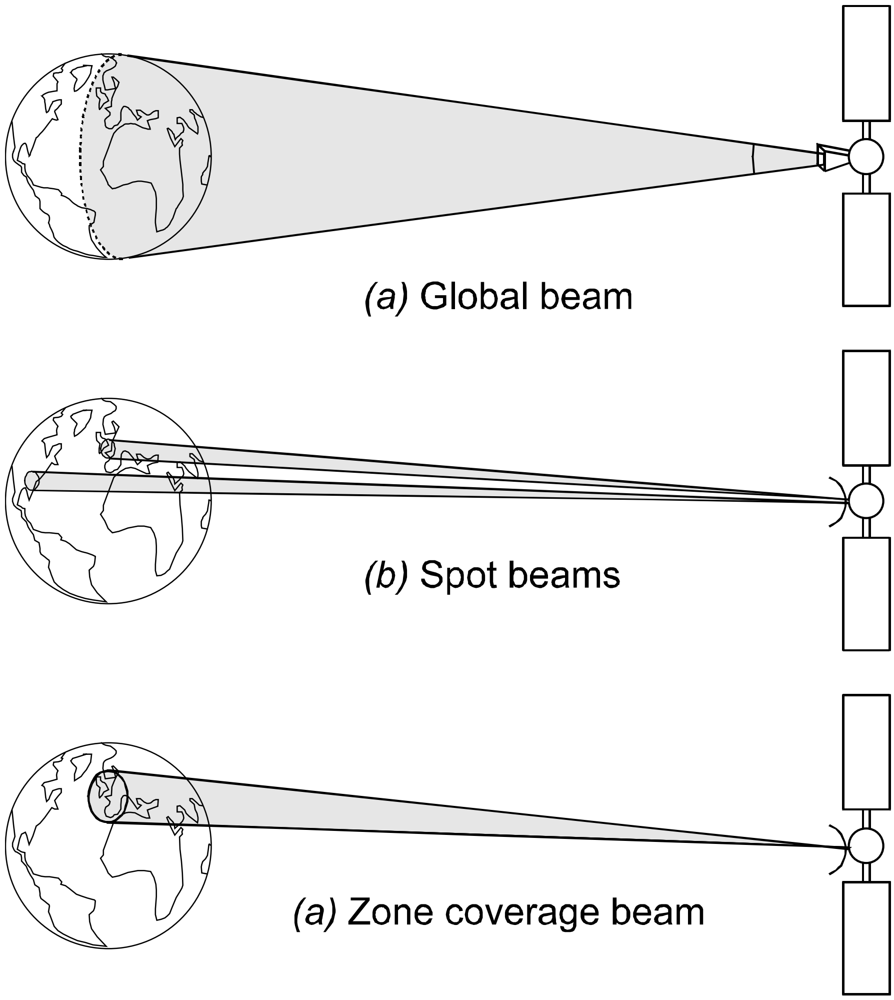


Figure 3-5: Example of the Ability to Shape and Focus Beams in the Microwave Range

3.5.2 Good Bandwidth Availability

Relatively large frequency bandwidths are available in the microwave bands. A single microwave carrier can transport a high capacity of information and over a given bandwidth range several independent carriers can be employed on a single link. Figure 3-6 shows an example of 10 carriers each bearing 34 Mbit/s of information on a medium capacity multi-channel system of 300 MHz total bandwidth.

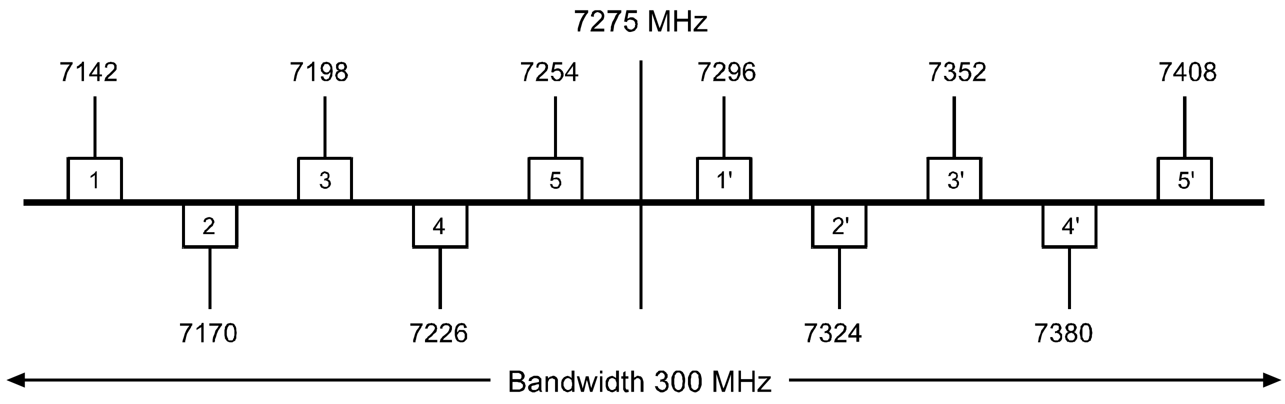


Figure 3.6: Example of a frequency plan showing 10 carriers each bearing 34 Mbits/s of information in a multi-channel microwave radio system

3.5.3 Relatively Low-Loss Window in the Atmosphere

Radio waves propagate through the Earth's atmosphere with little loss and fortunately for communications purposely are not normally greatly affected by adverse conditions such as fog, rain or snow. As the frequency of transmission increases into the microwave range losses climb but under clear dry conditions are still relatively low and do not present any severe difficulties for microwave radio and satellite communication links, at least up to X and J bands in the range of up to 20 GHz. Beyond 30 GHz attenuation in the atmosphere increases significantly and the effects of rain, fog and snow and specific oxygen and water vapour resonances place very restrictive limitation on range and scope of transmissions.

Figure 3-7 gives a guide to practical distance for terrestrial link; Figure 3-8 provides curves showing attenuation versus frequency illustrating the effects of fog, gas absorption resonances and rain.

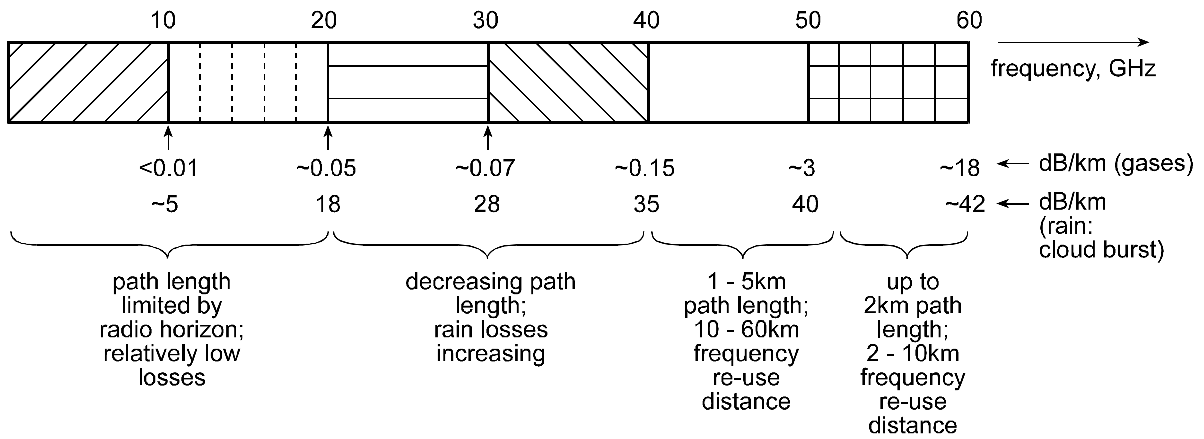


Figure 3-7: Single-link Microwave Radio Propagation Distances

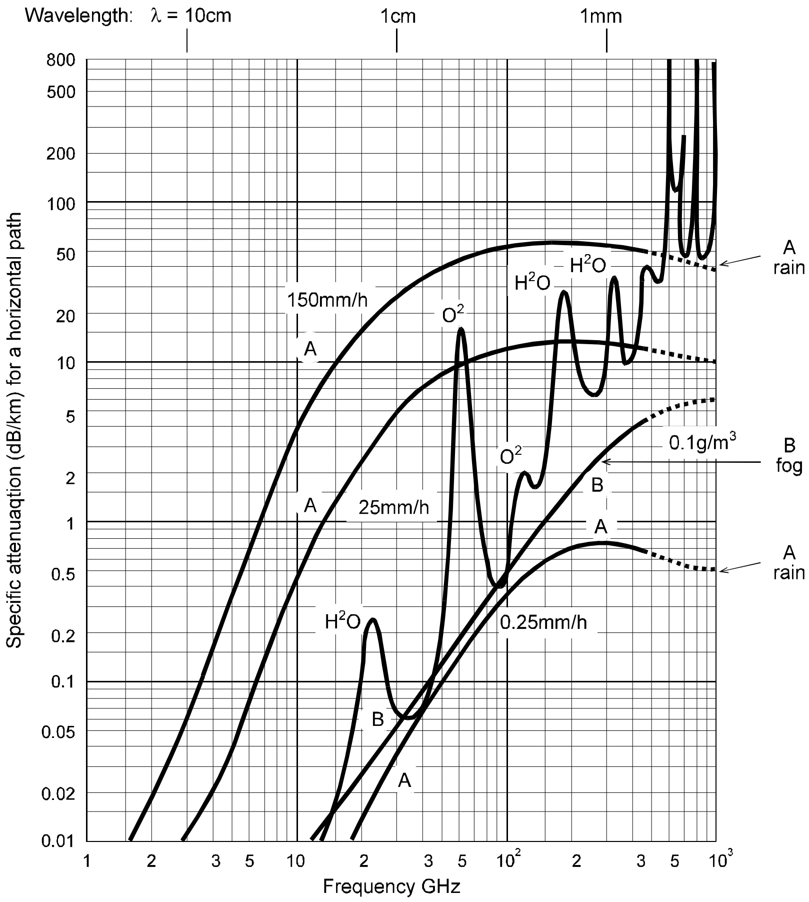


Figure 3-8: Attenuation of Microwaves in Atmosphere

3.5 Principal Applications of Microwaves

3.5.1 Telecommunications

Microwaves are extensively used as carriers of information in both terrestrial and satellite systems and together with optical fibre systems provide the means of national and international communications.

Microwave communications systems embrace:

- **Microwave terrestrial radio** Line-of-sight systems for telephony, television relay, data and personal communications; an example of a microwave link network is sketched in Figure 3-9.

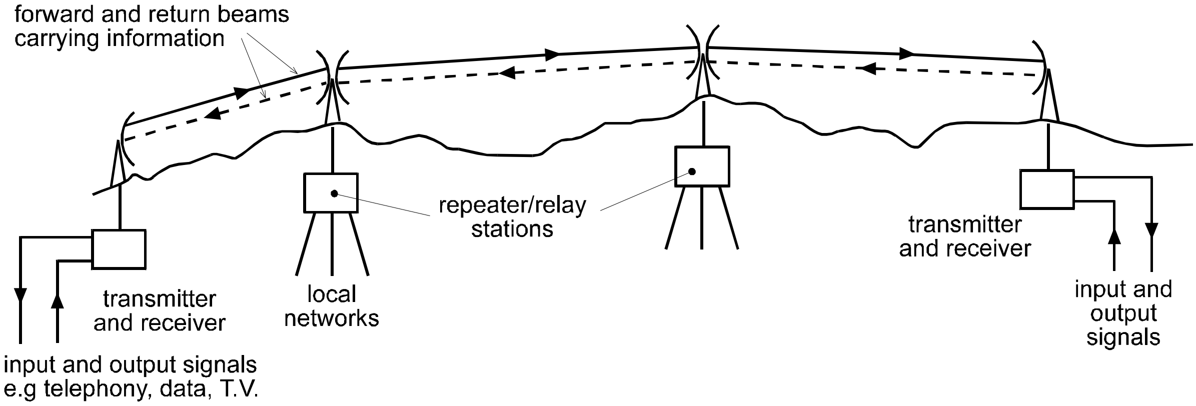


Figure 3-9: Microwave Radio Link Network

- **Satellite communications and television** Technological miracle come true and developed from the early 1960’s to act as a major provider for international telephony, data communications and television; an example of the wide variety of services available via satellite-Earth station links is sketched in Figure 3-10.

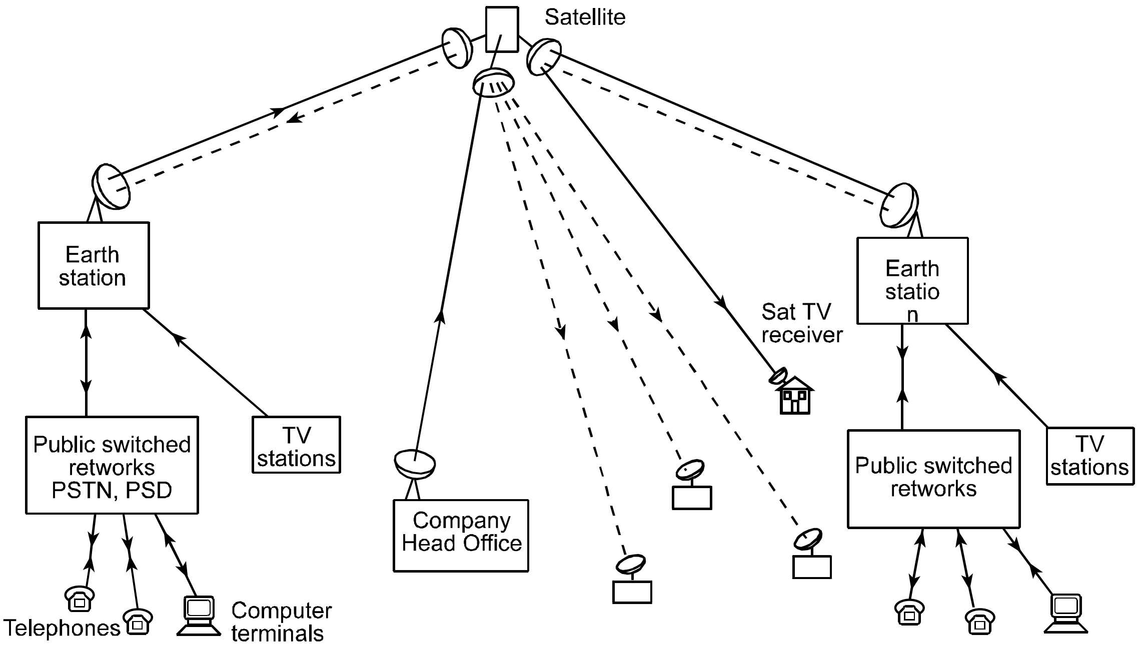


Figure 3-10: Examples of Services offered by Satellite Communications and Television

- **Tropospheric scatter communications** High-power microwave transmitters communication via scatter mechanisms from tropospheric layers, useful for linking remote locations e.g. widely distributed islands, oil rigs.

3.5.2 Radar and Navigational Aids

Radar is an acronym for radio detection and ranging but is now understood to define the devices, techniques and systems used for determining the range, position, velocity etc... of distant objects by means of radio waves.

With a few exceptions the radio waves used are in the microwave bands, principally centred on 3 GHz and 10 GHz but more recently moving up into the millimetre waves regions for specialist applications such as anti-collision and automatic tolling systems for cars and high resolution ground radars for tracking local movements in airports.

Radar systems for detecting position and range operate by transmitting a beam of radio waves in the direction of the object and measuring the time for the reflected wave, the echo, to return to the radar source, as indicated in Figure 3-11.

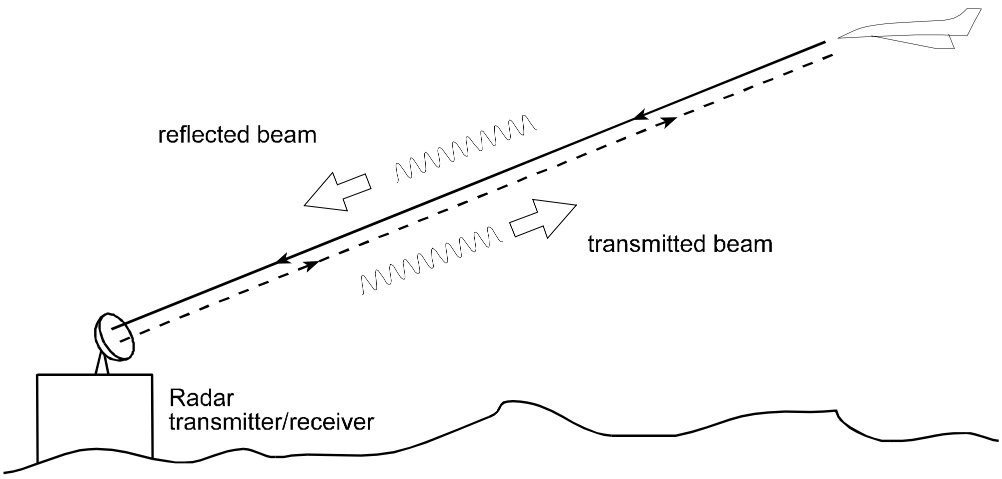


Figure 3-11: Radar, Radio Detection and Ranging

There are a wide diversity of radar applications related to the detection and location of targets, providing information of target's position, course, speed, to such applications as security and alarm systems. The principal application areas are listed below:

- Aircraft and airports: surveillance, tracking, precision approach and navigational radars.
- Maritime radars.
- Radar altimeters.
- Global positioning (GPS) radars using satellites.
- Radar beacons, racons for navigation and position fixing.
- Weather forecasting and storm detection radars.
- Doppler radars for velocity measurement.
- MTI, moving target indicator radars for displaying moving targets by cancelling all echoes from stationary targets.

3.6 Industrial, Scientific and Medical Applications of Microwaves

In addition to the primary applications of microwaves in terrestrial and satellite communications and radar, there are a number of important applications in industrial, scientific, and medical fields. These include:

- **Microwave ovens at 2.45 GHz**: fast-food cooking, defrosting.
- **Industrial drying and heating**: drying of paper, printing inks, textiles, food, thermal treatment of materials, pharmaceutical products, vulcanisation of rubbers, silicones.
- **Surveillance and alarm systems.**
- **Microwave sensors, measurements and telemetry.**
- **High-energy scientific applications**: particle accelerators, heating plasma for thermonuclear fusion.
- **Medical applications**: treating of cancers by penetration and heating effect of microwaves; microwave cauterisation; medical sterilisation; microwave radiometry diagnosis.

3.7 Millimetric Waves and Applications

With the ever increasing demands for telecommunications the centimetre wavelengths part of the microwave spectrum is now reaching virtually full capacity. However, the millimetre waveband, 30 –300 GHz, provides a vast, largely unused, bandwidth available for exploitation by future communication users, particularly in the areas of fixed, mobile, broadcast, satellite and broadband integrated telecommunication services.

Millimetre waves are ideal for short-hop, high capacity systems. Such links present a lower overall cost than fibre-optic cable, and far better systems availability than can be achieved with infra red.

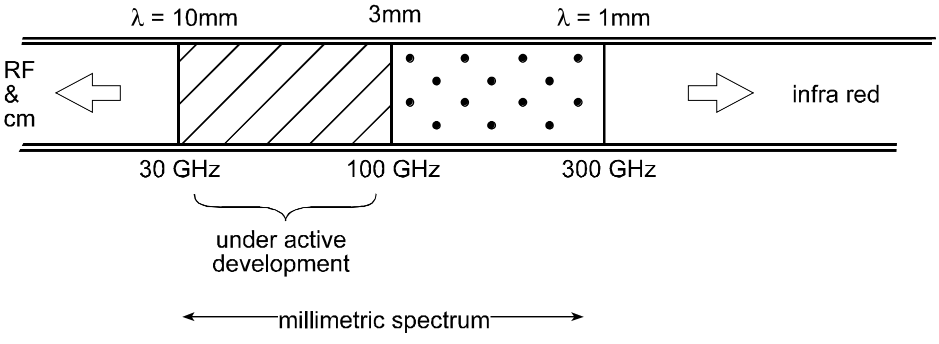


Figure 3-12

Equipment and antenna size is conveniently small, installation and reconfiguration is normally easy, and millimetric systems can operate with low rf power, e.g. 2 and 8 Mbit/s links in the 38GHz band operate very reliably with RF power below 5 mW.

Although millimetre waves are subject to much severer constraints due to atmospheric oxygen and water vapour and rain precipitation than centimetre waves, they 'see' through bad weather orders of magnitude better than infra red. Range limitation can be used to advantage in many mobile and fixed low-range communications applications by permitting frequency re-use.

The radio frequency spectrum in the millimetre waveband is still very largely unoccupied at present, and thus there is potentially available extensive bandwidth for high capacity transmission and specialised new services.

The relative merits, increasing telecommunication demands and potential millimetre wave application are listed below:

3.7.1 Relative Merits of Millimetre Waves

- Propagation characteristics superior to infra red.
- Lower and easier installation than cable.
- Ideal for short-haul links.
- Wide bandwidth enables high traffic capacity.
- Low rf power operation possible.
- Frequency re-use available.
- Small equipment and compact antennas.

3.7.2 Increasing Demands in Telecommunications for:

- Fixed, mobile, broadcast, terrestrial and satellite services.
- Broadband integrated telecommunications services.
- Short-haul business/data links.
- Personal communications.
- Video distribution.
- Give rise to a growing number of applications.

3.7.3 Applications

- Short-haul, point-to-point, and multi-point terrestrial links.
- Millimetric video distribution (MVDS).
- Electronic news gathering (EGN).
- Intra-building communications.
- Cordless LANs.
- Mobile radio systems: personal communications, pagers.
- Motor vehicle applications: automatic tolling, vehicle identification, auto-routing, anti-collision radar.
- Millimetric satellite communications.
- Broadband integrated services.

Notes

1 SI UNITS (INTERNATIONAL SYSTEM OF UNITS)

(I) The seven SI base units

Quantity	Name of unit	Unit abbreviation
length	metre	m
mass	kilogram	kg
time	second	s
electric current	ampere	A
thermodynamic temperature	kelvin	K
luminous intensity	candela	cd
amount of substance	mole	mol
plane angle	radian	rad
solid angle	steradian	sr

(ii) Some important SI derived units

Quantity	Name of unit	Unit abbreviation
area	square metre	m^2
volume	cubic metre	m^3
speed, velocity	metre per second	m/s
angular velocity	radian per second	rad/s
acceleration	metre per second squared	m/s^2
angular acceleration	radian per second squared	rad/s^2
density	kilogram per cubic metre	kg/m^3
frequency	hertz	Hz
force	newton	N
pressure	pascal, newton per square metre	Pa,N/m^2
torque, moment	newton metre	Nm
energy, work	joule	J
power	watt	W
electric charge	coulomb	C
electric potential	volt	V
electric field strength	volt per metre	V/m
magnetic flux	weber	Wb
magnetic flux density	tesla	T
magnetic field strength	ampere per metre	A/m
resistance	ohm	Ω
conductance	siemens	S
capacitance	farad	F
inductance	henry	H
permittivity	farad per metre	F/m
permeability	henry per metre	H/m

(iii) SI prefixes

Name	Symbol	Meaning	*(i.e. factor by which unit is multiplied)*
tera	T	x 10^{12}	(a million million times)
giga	G	x 10^{9}	(a thousand million times)
mega	M	x10^{6}	(a million times)
kilo	k	x10^{3}	(a thousand times)
hecto	h	x10^{2}	(a hundred times)
deca	da	x10^{1}	(ten times)
deci	d	x10^{-1}	(a tenth)
centi	c	x10^{-2}	(a hundredth)
milli	m	x10^{-3}	(a thousandth)
micro	μ	x10^{-6}	(a millionth)
nano	n	x10^{-9}	(a thousand millionth)
pico	p	x10^{-12}	(a million millionth)

2 PHYSICAL AND MATHEMATICAL CONSTANTS

Velocity of light in free space	c	=2.997925 x 10^{8} m/s
Velocity of light in dry air 20°c	c_{air}	=2.997 x 10^{8} m/s
Permittivity of free space	ε_o	=8.85416 x 10^{-12}
		$\cong(^1/_{36}\pi)$ x 10^{-9} F/m
Permeability of free space	μ_o	=4π x 10^{-7} =1.256637 x 10^{-6} H/m
Impedance of free space	Z_O	=376.7 $\cong$120π ohms
Charge of electron	e	=1.60202 x 10^{-19} c
Mass of electron	m	=9.1083 x 10^{-31} kg
Ratio of charge to mass of electron	$^e/_m$	=1.758857 x 10^{11} c/kg
Boltzmann's constant	k	=1.3804 x 10^{-23} J/K
Planck's constant	h	=6.547 x 10^{-34} Js

π=3.14159265 l/π=0.31831 e=2.7183 $\log_{10}e$=0.4343

Mean radius of Earth	Re	=6371 km

CLASSIFICATION OF RADIO WAVE RANGES

Classification	Frequency range	Wavelength range
Very low frequency (VLF)	3 - 30 kHz	100 - 10 km
Low frequency (LF)	30 - 300 kHz	10 - 1 km
Medium frequency (MF)	300 kHz - 3 MHz	1000 - 100 m
High frequency (HF)	3 - 30 MHz	100 - 10 m
Very high frequency (VHF)	30 - 300 MHz	10 - 1 m
Ultra high frequency (UHF)	300 - 3000 MHz	100 - 10 cm
Super high frequency (SHF)	3 - 30 GHz	10 - 1 cm
Extra high frequency (EHF)	30 - 300 GHz	10 - 1 mm

Letter designation for RF and Microwave bands

Band Descriptions: Old	Band Descriptions: New
HF 3 - 30 MHz	**A** 0 - 250 MHz
VHF 30 - 300 MHz	
	B 250 - 500 MHz
UHF 300 - 1000 MHz	**C** 500 - 1000 MHz
L 1 - 2 GHz	**D** 1 - 2 GHz
S 2 - 4 GHz	**E** 2 - 3 GHz
	F 3 - 4 GHz
C 4 - 8 GHz	**G** 4 - 6 GHz
	H 6 - 8 GHz
X 8 - 12 GHz	**I** 8 - 10 GHz
J / **Ku** 12 - 18 GHz	**J** 10 - 20 GHz
K 18 - 27 GHz	**K** 20 - 40 GHz
Q / **Ka** 27 - 40 GHz	
V 40 - 60 GHz	**L** 40 - 60 GHz
O / **E** 60 - 90 GHz	**M** 60 - 100 GHz
W 75 - 120 GHz	

Frequency GHz	Wavelength
0.1	3 m
0.15	2 m
0.2	1.5 m
0.3	100 cm
0.5	60 cm
0.75	40 cm
1.0	30 cm
1.5	20 cm
2	15 cm
3	100 cm
4	7.5 cm
5	6 cm
6	5 cm
8	3.75 cm
10	3 cm
15	2 cm
20	1.5 cm
30	10 cm
40	7.5 cm
50	6 mm
60	5 mm
75	4 mm
100	3 mm

5 Resistivity and conductivity of some common metals

Copper	1.7×10^{-8} Ωm 5.9×10^{7} S/m	Aluminum	2.9×10^{-8} Ωm 3.4×10^{7} S/m
Brass (70/30)	1.7×10^{-8} Ωm 5.9×10^{7} S/m	Gold	2.4×10^{-8} Ωm 4.2×10^{7} S/m
Silver	1.6×10^{-8} Ωm 6.3×10^{7} S/m	Platinum	1.2×10^{-8} Ωm 8.3×10^{6} S/m
Chromium	1.3×10^{-7} Ωm 7.7×10^{7} S/m	Tantalum	1.4×10^{-7} Ωm 7.1×10^{6} S/m

6 Skin Depth

At radio frequencies the current in a conductor is not uniformly distributed as at DC but tends to be concentrated at the surface and contained effectively within a depth δ, known as the skin depth.

$$d = 1/\sqrt{(\pi f \sigma \mu)}$$

where σ = conductivity
μ = permeability of conductor
f = frequency

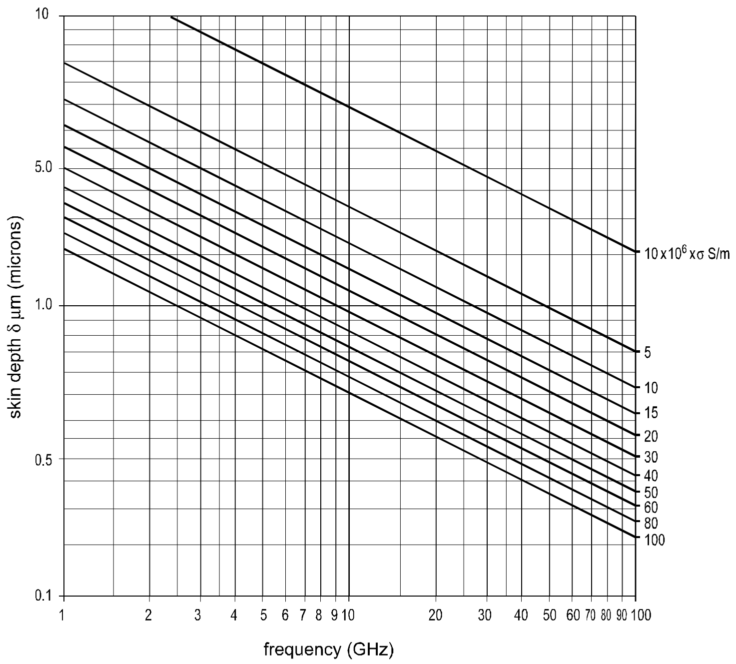
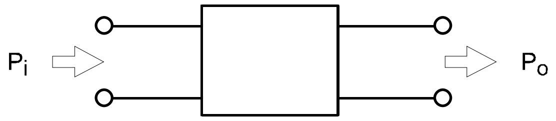


Figure A-1

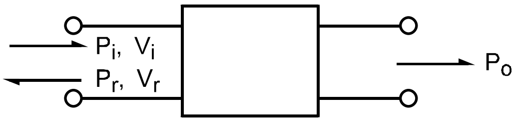
7 Table of Insertion Loss versus Fraction of Power Transmitted



Insertion Loss = | 10 log P_o/P_i | dB
Fraction of power transmitted = P_o/P_i

Insertion Loss 10log(P/P)dB	Fraction of Power Trasmitted
0 dB	1.0
0.1 dB	0.977
0.2 dB	0.955
0.3 dB	0.933
0.4 dB	0.912
0.5 dB	0.891
0.6 dB	0.871
0.7 dB	0.851
1.0 dB	0.794
2.0 dB	0.631
3.0 dB	0.501
4.0 dB	0.398
5.0 dB	0.316
7.0 dB	0.200
10.0 dB	0.1
20.0 dB	0.01
30.0 dB	0.001
40.0 dB	0.0001
50.0 dB	0.00001
60.0 dB	0.000001

8 Table of voltage standing wave ratio (VSWR), reflection coefficient, return loss and % of power reflected



$$\text{vswr} = \frac{1 + |r|}{1 - |r|}, \quad r = \frac{Vr}{Vr} \quad \text{... reflection coefficient}$$

return loss = 10 log P_r/P_i = 20 log | V_r/V_i | = 20 log | r | dB

reflected power $P_o = | r |^2 \; P_i$

VSWR	reflection coefficient	return loss	% power reflected
1	0	infinite	0
1.05	0.024	-32.3	0.06
1.1	0.048	-26.4	0.23
1.2	0.091	-20.8	0.83
1.3	0.13	-17.7	1.69
1.4	0.167	-15.6	2.78
1.5	0.20	-14.0	4.00
1.6	0.23	-12.7	5.29
1.7	0.26	-11.7	6.76
1.8	0.29	-10.88	8.16
1.9	0.31	-10.16	9.63
2.0	0.33	-9.54	11.11
3.0	0.50	-6.02	25.00
4.0	0.60	-4.44	36.00
5.0	0.67	-3.48	44.44
6.0	0.71	-2.97	51.02
10	0.82	-1.74	66.94
20	0.905	-0.86	81.9
30	0.935	-0.58	87.5

9 Decibel watts to watt (dBW to W) conversions

dBW = 10 log (P/1 watt)

i.e. P, the power in watts is expressed in decibels relative to 1W

$P = 10^{dBW/10}$ watts

e.g. 48 dBW represents a power , P $10^{4.8}$ = 63096 W

–35 dBW represents a power , P $10^{-3.5}$ = 0.00032 W

dBW	W (watts)
–120	1x10 W or 1 pico-watt
–110	1x10 W or 10 pW
–100	1x10 W or 100 pW
–90	1x10 W or 1 nano-watt
–80	1x10 W or 10 nW
–70	1x10 W or 100 nW
–60	1x10 W or 1 micro-watt
–50	1X10 W or 10 W
–40	1x10 W or 100 W
–30	1x10 W or 1 milli-watt
–20	1x10 W or 10 mW
–10	1x10 W or 100 mW
0	1 W
10	10 W
20	100 W
30	1000 W or 1 kilo-watt
40	10 kW

10 Decibel milliwatts to milliwatt (dBm to mW) conversions

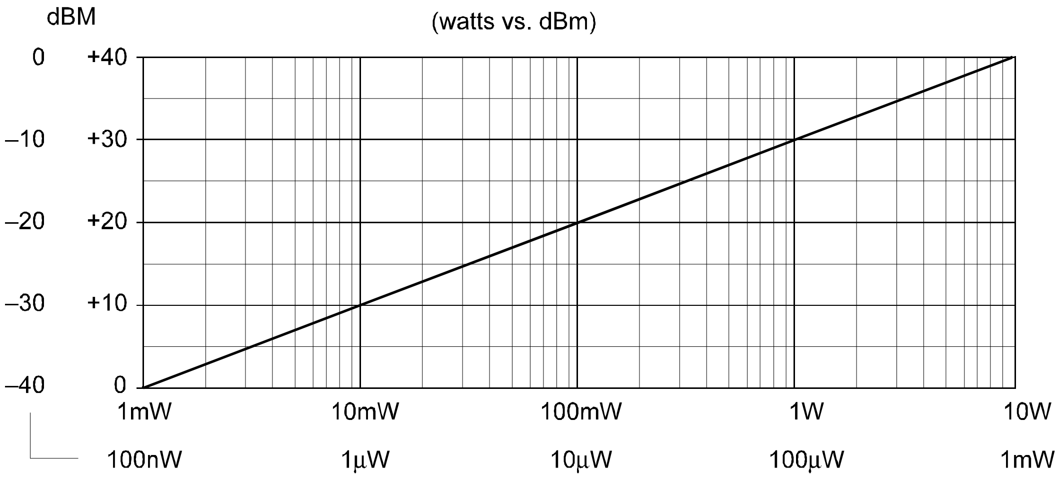
dBm = 10log (PmW/1mW)

1.e P in milliwatts expressed in decibels relative to 1mW

P = 10dBm/10 mW

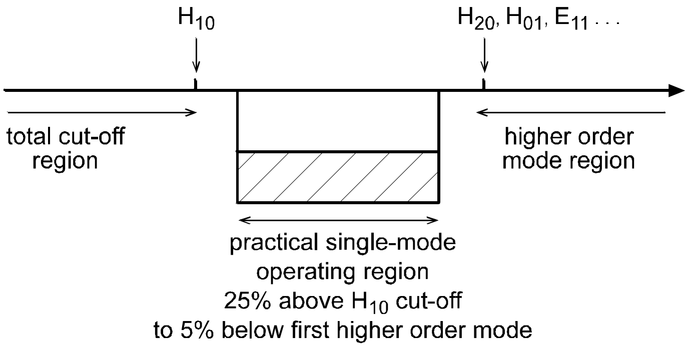
e.g. –15dBm represents P = 10 $^{-1.5}$ = 0.032 mW

+27dBm represents P = 10$^{2.7}$ = 501 mW



dBm	mW (milliwatts)
–30	0.001 mW
–20	0.01 mW
–10	0.1 mW
0	1 mW
10	10 mW
20	100 mW
30	1000 mW or 1 W
40	10000 mW or 10 W

11 Rectangular waveguide: sizes and operating ranges



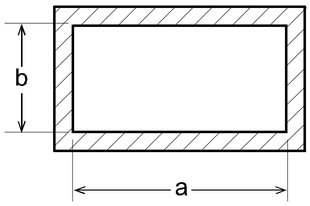
WAVEGUIDE DESIGNATION					
UK WG	US WR	International IEG-R	Internal dimensions mm	H cut-off GHz	Practical operating range GHz
00	2300	3	584.0 x 292.0	0.256	0.32 - 0.49
0	2100	4	533.0 x 267.0	0.281	0.35 - 0.53
1	1800	5	457.0 x 229.0	0.328	0.40 - 0.625
2	1500	6	381.0 x 191.0	0.393	0.49 - 0.75
3	1150	8	292.0 x 146.0	0.513	0.64 - 0.96
4	975	9	248.0 x 124.0	0.605	0.75 - 1.12
5	770	12	196.0 x98.0	0.766	0.96 - 1.45
6	650	14	165.0 x83.0	0.908	1.12 - 1.70
7	510	18	131.0 x65.0	1.157	1.45 - 2.20
8	430	22	109.0 x55.0	1.372	1.70 - 2.60
9A	340	26	86.0 x43.0	1.736	2.20 - 3.30
10	284	32	72.1 x34.0	2.080	2.60 - 3.95
11A	229	40	59.0 x29.0	2.577	3.30 - 4.90
12	187	48	47.5 x22.1	3.15	3.95 - 5.85
13	159	58	40.0 x20.0	3.711	4.90 - 7.05
14	137	70	35.0 x16.0	4.301	5.85 - 8.20
15	112	84	28.5 x12.6	5.260	7.05 - 10.0

Table continues overleaf

Rectangular waveguide: sizes and operational ranges (continued)

WAVEGUIDE DESIGNATION					
UK WG	US WR	International IEG-R	Internal dimensions mm	H cut-off GHz	Practical operating range GHz
16	90	100	22.86 x 10.16	6.557	8.20 - 12.40
17	75	120	19.0 x 9.5	7.868	10.00 - 15.00
18	62	140	16.0 x 7.9	9.468	12.40 - 18.00
19	51	180	13.0 x 5.8	11.574	15.00 - 22.00
20	42	220	11.0 x 4.3	14.074	18.00 - 26.50
21	34	260	8.6 x 4.3	17.328	22.00 - 33.00
22	28	320	7.1 x 3.6	21.081	26.50 - 40.00
23	22	400	5.7 x 2.9	26.342	33.00 - 50.00
24	19	500	4.8 x 2.4	31.357	40.00 - 60.00
25	15	620	3.8 x 1.9	39.863	50.00 - 75.00
26	12	740	3.1 x 1.6	48.350	60.00 - 90.00
27	10	900	2.4 x 1.3	59.010	75.00 - 110.00
28	8	1200	2.0 x 1.0	73.484	90.00 - 140.00
29	7		1.7 x 0.82	90.480	110.00 - 170.00
30	5		1.3 x 0.65	115.75	140.00 - 220.00
31	4		1.1 x 0.55	131.52	170.00 - 260.00
32	3		0.87 x 0.44	173.28	220.00 - 325.00

12 Rectangular waveguides: basic parameters



Guide wavelength, $\lambda_g = \dfrac{\lambda}{\sqrt{\left(1-\lambda^2/\lambda_c^2\right)}}$

Phase constant, $\beta = 2\pi/\lambda g$

Cut off wavelength, $\lambda_c = \dfrac{1}{\sqrt{\left[\left(\dfrac{m}{2a}\right)^2+\left(\dfrac{n}{2b}\right)^2\right]}}$

for Hmm and Emm modes, $\lambda_c = 2a$ for H_{10} mode

H-wave impedance, $Z_H = 377\ \lambda_g/\lambda$

E-wave impedance, Z_E $377\ \lambda/\lambda_g$

Phase velocity $V_p = \omega/\beta = f\lambda_g$

Power flow for fundamental H10 mode,
$P = 662 \times 10^{-6}\ ab\ \sqrt{(1 - \lambda^2/\lambda_c^2)}\ E_o^2$

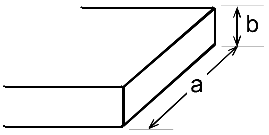
λ = free space wavelength

ω = $2\pi f$

f = frequency

E_o = peak electric field strength for H_{10} mode

13 Guide wavelength versus frequency for WG 16 over the X-band range 8.0 - 12.5 GHz



W16 (WR 90)

a = 0.9 inch = 2.286 cm

b = 0.4 inch = 1.016 cm

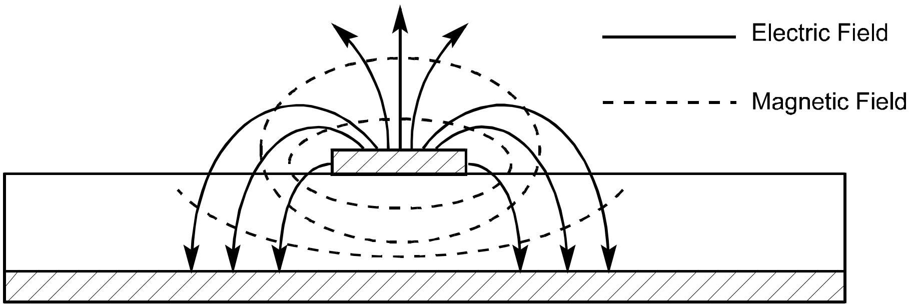
$\lambda_c = 2a = 4.572$ cm

$f_c = 6.56$ GHz for H_{10} mode

$$\text{Guide wavelength } \lambda_c = \frac{\lambda}{\sqrt{(1 - \lambda^2 / \lambda^2_c)}}$$

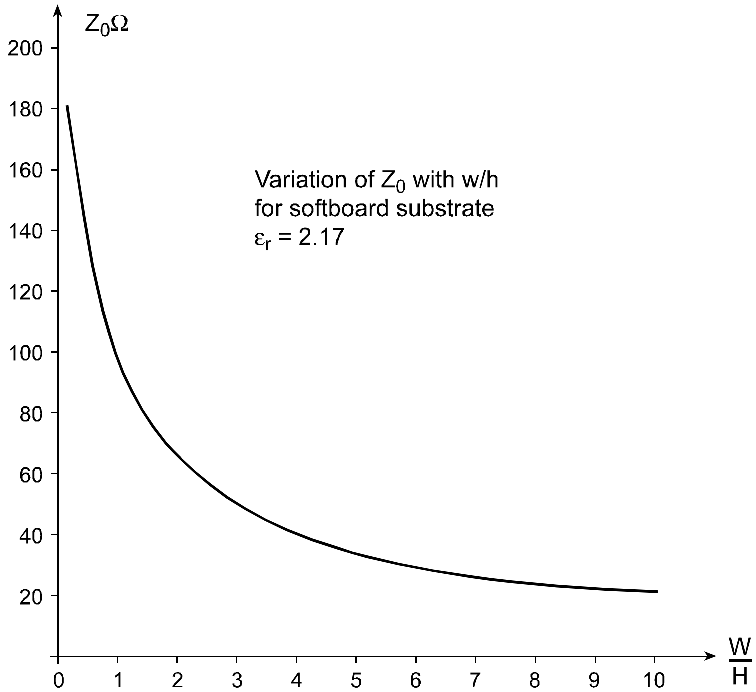
f_0	λg in cm										Subtract								
	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9
8.0	6.541	6.517	6.492	6.468	6.444	6.420	6.396	6.373	6.350	6.327	2	4	7	9	11	14	16	19	21
8.1	6.304	6.282	6.260	6.237	6.216	6.194	6.173	6.151	6.130	6.109	2	4	6	8	10	12	15	17	19
8.2	6.089	6.068	6.048	6.028	6.008	5.988	5.968	5,949	5.930	5.910	1	3	5	7	9	11	13	15	17
8.3	5.891	5.873	5.854	5.836	5.817	5.799	5.781	5.763	5,745	5.728	1	3	5	7	9	10	12	14	16
8.4	5.710	5.693	5.676	5.659	5.642	5.625	5.608	5.592	5.575	5.559	1	3	5	6	8	10	11	13	15
8.5	5.543	5.527	5.511	5.495	5.479	5.464	5.488	5.433	5.418	5.403	1	3	4	6	7	9	10	12	14
8.6	5.388	5.373	5.358	5.343	5.329	5.314	5.300	5.285	5.271	5.257	1	2	4	5	7	8	10	11	13
8.7	5.243	5.229	5.215	5.202	5.188	5.175	5.161	5.148	5.134	5.121	1	2	4	5	6	8	9	10	12
8.8	5.108	5.095	5.082	5.069	5.057	5.044	5.031	5.019	5.006	4.994	1	2	3	5	6	7	8	10	11
8.9	4.982	4.970	4.957	4.945	4.933	4.922	4.910	4.898	4.886	4.875	1	2	3	4	5	7	8	9	10
9.0	4.863	4.852	4.840	4.829	4.818	4.806	4.795	4.784	4.773	4.762	1	2	3	4	5	6	7	8	10
9.1	4.751	4.740	4.730	4.719	4.708	4.698	4.687	4.677	4.666	4.656	1	2	3	4	5	6	7	8	9
9.2	4.646	4.635	4.625	4.615	4.605	4.595	4.585	4.575	4.565	4.556	1	2	3	4	5	6	7	8	9
9.3	4.546	4.536	4.526	4.517	4.507	4.498	4.488	4.479	4.470	4.460	0	1	2	3	4	5	6	7	8
9.4	4.451	4.442	4.433	4.424	4.415	4.406	4.397	4.388	4.379	4.370	0	1	2	3	4	5	6	7	8
9.5	4.361	4.352	4.344	4.335	4.326	4.318	4.309	4.301	4.292	4.284	0	1	2	3	4	5	6	6	7
9.6	4.276	4.276	4.259	4.251	4.243	4.234	4.226	4.218	4.210	4.202	0	1	2	3	4	4	5	6	7
9.7	4.194	4.186	4.178	4.170	4.163	4.155	4.147	4.139	4.132	4.124	0	1	2	3	3	4	5	6	7
9.8	4.116	4.109	4.101	4.094	4.086	4.079	4.071	4.064	4.057	4.049	0	1	2	2	3	4	5	5	6
9.9	4.042	4.035	4.027	4.020	4.013	4.006	3.999	3.992	3.985	3.978	0	1	2	2	3	4	4	5	6
10.0	3.971	3.964	3.957	3.950	3.943	3.936	3.929	3.923	3.916	3.909	0	1	2	2	3	4	4	5	6
10.1	3.903	3.896	3.889	3.883	3.876	3.869	3.863	3.856	3.850	3.844	0	1	1	2	3	3	4	5	5
10.2	3.837	3.831	3.824	3.818	3.812	3.805	3.799	3.793	3.787	3.780	0	1	1	2	3	3	4	5	5
10.3	3.774	3.768	3.762	3.756	3.750	3.744	3.738	3.732	3.726	3.720	0	1	1	2	3	3	4	4	5
10.4	3.714	3.708	3.702	3.696	3.690	3.684	3.679	3.673	3.667	3.661	0	1	1	2	2	3	4	4	5
10.5	3.656	3.650	3.644	3.639	3.633	3.627	3.622	3.616	3.611	3.605	0	1	1	2	2	3	3	4	5
10.6	3.600	3.594	3.589	3.583	3.578	3.572	3.567	3.562	3.556	3.551	0	1	1	2	2	3	3	4	4
10.7	3.546	3.540	3.535	3.530	3.525	3.519	3.514	3.509	3.504	3.499	0	1	1	2	2	3	3	4	4
10.8	3.493	3.488	3.483	3.478	3.473	3.468	3.463	3.458	3.453	3.448	0	1	1	2	2	3	3	4	4
10.9	3.443	3.438	3.433	3.428	3.423	3.419	3.414	3.409	3.404	3.399	0	0	1	1	2	2	3	3	4
11.0	3.394	3.390	3.385	3.380	3.375	3.371	3.366	3.361	3.357	3.352	0	0	1	1	2	2	3	3	4
11.1	3.347	3.343	3.338	3.333	3.329	3.324	3.320	3.315	3.311	3.306	0	0	1	1	2	2	3	3	4
11.2	3.302	3.297	3.293	3.288	3.284	3.279	3.275	3.271	3.266	3.262	0	0	1	1	2	2	3	3	3
11.3	3.258	3.253	3.249	3.245	3.240	3.236	3.232	3.227	3.223	3.219	0	0	1	1	2	2	3	3	3
11.4	3.215	3.211	3.206	3.202	3.198	3.194	3.190	3.186	3.181	3.177	0	0	1	1	2	2	2	3	3
11.5	3.173	3.169	3.165	3.161	3.157	3.153	3.149	3.145	3.141	3.137	0	0	1	1	2	2	2	3	3
11.6	3.133	3.129	3.125	3.121	3.117	3.113	3.109	3.105	3.102	3.098	0	0	1	1	1	2	2	3	3
11.7	3.094	3.090	3.086	3.082	3.079	3.075	3.071	3.067	3.063	3.060	0	0	1	1	1	2	2	3	3
11.8	3.056	3.052	3.048	3.045	3.041	3.037	3.034	3.030	3.026	3.023	0	0	1	1	1	2	2	2	3
11.9	3.019	3.015	3.012	3.008	3.004	3.001	2.997	2.994	2.990	2.987	0	0	1	1	1	2	2	2	3
12.0	2.983	2.979	2.976	2.972	2.969	2.965	2.962	2.958	2.955	2.951	0	0	1	1	1	2	2	2	3
12.1	2.948	2.945	2.941	2.938	2.934	2.931	2.928	2.924	2.921	2.917	0	0	1	1	1	2	2	2	3
12.2	2.914	2.911	2.907	2.904	2.901	2.897	2.894	2.891	2.887	2.884	0	0	0	1	1	1	2	2	2
12.3	2.881	2.878	2.874	2.871	2.868	2.865	2.861	2.858	2.855	2.852	0	0	0	1	1	1	2	2	2
12.4	2.849	2.845	2.842	2.839	2.836	2.833	2.830	2.826	2.823	2.820	0	0	0	1	1	1	2	2	2
12.5	2.817	2.814	2.811	2.808	2.805	2.802	2.799	2.795	2.792	2.789	0	0	0	1	1	1	2	2	2

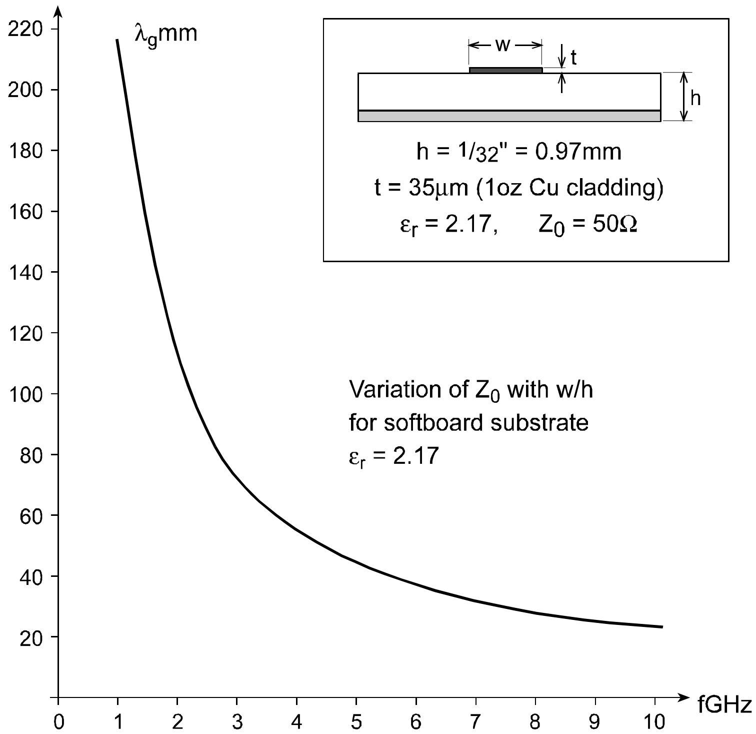
14 Dielectric substrate material characteristics



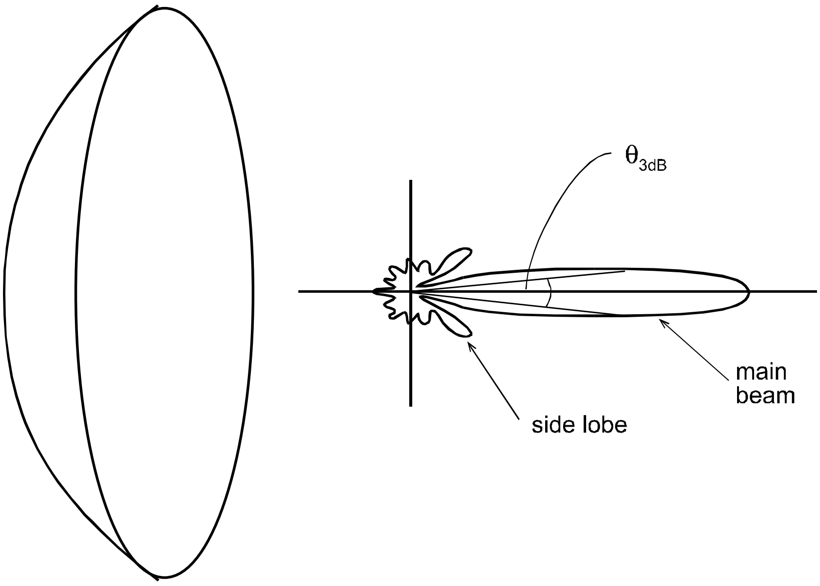
Substrate Material	**Dielectric Constant**	**Loss Tangent**	**Comments**
PTFE woven glass laminates	2.17 2.134 2.4 to 2.55	0.001 0.0015 0.002	Very good dimensional stability. Temperature range -27° to 260°C Medium Cost. Easy matching.
PTFE ceramic filled. e.g 'Epsilam 10'	10.2 ±0.25	0.002	Extremely useful for compacat circuits and a good alternative to alumina.
PTFE glass cloth ceramic powder filled.	6.0 ±0.5	0.0002	Typical loss for PTFE softboards: er / Freq (GHz) / Loss (dB/m) 2.17 / 2.0 / 0.8 2.17 / 10.0 / 4.0 10.0 / 2.0 / 2.4 10.0 / 10.0 / 8.0
Alumina	9.7 to 10.3	0.004	Excellent dimensional stability. Temperature range to 1600°C. Medium high cost. Difficult to machine
Sapphire	9.0	0.0001	Excellent dimensional stability. Temperature range -24°C to 371°C. Very high cost.
Glass bonded mica	7.5	0.002	Excellent dimensional stability. Medium cost.

15 Microstrip characteristics: characteristic impedance and guide wavelength for CU 217 substrate





16 Gain of parabolic reflector antennas



Antenna gain, $G = \eta\pi^2(D/\lambda)^2$

Where D = diameter of antenna aperture

l = wavelength

η = aperture illumination dfficiency (typically 0.55)

In terms of 3dB beamwidths:

$G \approx 30{,}000 / \theta_{3dB}\ \phi_{3db}$

Dish diameter m	Gain, dB, of parabolic dish antenna, η = 0.55			
	4 GHz	6 GHz	12 GHz	14 GHz
0.5	23.8	27.3	33.4	43.7
1.0	29.8	33.4	39.4	40.7
1.2	31.4	35.0	41.0	42.3
1.8	35.0	38.5	44.5	45.8
2.3	37.1	40.6	46.6	48.0
3.0	39.4	42.9	48.9	50.3
5.0	43.8	47.3	53.4	54.7
11.0	50.7	54.2	60.2	61.6
18.0	55.0	58.5	64.5	65.8
30.0	59.4	62.9	68.9	70.3
32.0	59.9	63.5	69.5	70.8

Notes

1 TRANSMISSION LINE EQUATIONS AND AC STEADY STATE SOLUTIONS

1.1 The distributed parameters of a line

In the circuit theory approach to transmission line theory, it is assumed that a uniform line can be characterised by four parameters which are distributed uniformly along the line length. The distributed parameters, also known as the line primary constants, are:

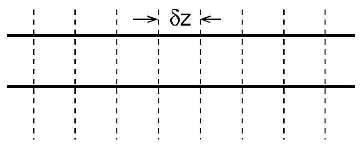
R = the series ac resistance per unit length of line (including both conductors)

L = the series inductance per unit length of line (including both conductors)

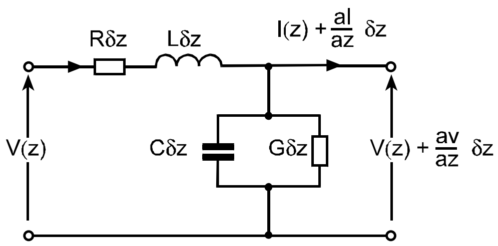
C = the shunt capacitance between the line conductors per unit line length

G = the shunt conductance between the line conductors per unit line length

1.2 Transmission line equations



(a)
Transmission lone showing it division into elemental sections of length $\delta z << \lambda$



(b)
Approximate equivalent circuit of an element of line, length $\delta z << \lambda$

Figure B-1

The representation of a uniform line as a cascade connection of a finite number of small elements of length $\delta z << \lambda$ is shown in Figure B-1.The transmission line equations may be derived from any of the elemental sections shown (by using Kirchhoff's laws and taking the limit as $\delta z \rightarrow 0$) and are given in general by:

$$\frac{\partial v}{\partial z} = -\left(Ri + L\frac{\partial i}{\partial t}\right) \quad .(1)$$

$$\frac{\partial i}{\partial z} = -\left(Gv + C\frac{\partial v}{\partial t}\right) \quad .(2)$$

which in the ac steady state,
i.e. assuming $v = V(z)e^{j\omega t}, i = I(z)e^{j\omega t}$ reduce to

$$\frac{dV}{dz} = -ZI \quad (3)$$

$$\frac{dI}{dz} = -YV \quad (4)$$

where v, V = voltage between line conductors at plane z

i,I = current flowing in line conductors through plane z

Z = (R+jωL) = series impedance of line per unit length

Y = (G+jωC) = shunt admittance of line per unit length

On eliminating I from (3) and (4), or V from (3) and (4) we obtain:

$$\frac{d^2V}{dz^2} - ZYV = 0 \quad (5)$$

$$\frac{d^2I}{dz^2} - ZYI = 0 \quad (6)$$

1.3 AC Steady state solutions

The general solutions of the steady state transmission line equations are

$$V = V^+e^{-yz} + V^-e^{+yz} \quad (7)$$

$$I = I^+e^{-yz} + I^-e^{-yz} = \frac{V+}{Z_c}e^{-yz} - \frac{V^-}{Z_c}e^{+yz} \quad (8)$$

where y = √ (ZY) known as the propagation constant of the line,

$$Z_0 = \sqrt{\left(\frac{Z}{Y}\right)} = \frac{V^+}{I^+} = \frac{V^-}{-I^-}$$

is known as the line characteristic impedance,

V^+, V^-, I^+, I^- are constants which may be evaluated when the conditions at the load and generator ends are given.

y and Z_0 may be expressed in terms of the line parameters as follows:

$$y=\sqrt{\left[(R+j\omega L)(G+j\omega C)\right]}=\alpha+j\beta$$

$$\alpha=\frac{1}{\sqrt{2}}\sqrt{\left[(RG-\omega^2LC)+\sqrt{\left\{(R^2+\omega^2L^2)(G^2+\omega^2C^2)\right\}}\right]}$$

is known as the attenuation constant; units, nepers per unit length

$$\beta=\frac{1}{\sqrt{2}}\sqrt{\left[(\omega^2LC-RG)+\sqrt{\left\{(R^2+\omega^2L^2)(G^2+\omega^2C^2)\right\}}\right]}$$

is known as the phase constant: - unit's radians per unit length

$$Z_0=\sqrt{\left(\frac{R+j\omega L}{G+j\omega C}\right)},\left|Z_0\right|=\left(\frac{R^2+\omega^2L^2}{G^2+\omega^2C^2}\right)^{1/4}$$

$$\angle Z_0=\frac{1}{2}\left[\tan^{-1}\frac{\omega L}{R}-\tan^{-1}\frac{\omega C}{G}\right]$$

For low-loss r.f. lines $\omega L >> R, \omega C >> G$ and the following approximations apply:

$$\alpha=\frac{1}{2}\left[R\sqrt{\frac{C}{L}}+G\sqrt{\frac{L}{C}}\right],\beta+\omega\sqrt{(LC)},Z_c=\sqrt{\frac{L}{C}}$$

The full time varying solutions, corresponding to cosine or sinusoidal sources, can be found by taking the real or imaginary parts of $V(z)e^{j\omega t}$, $I(z)e^{j\omega t}$, e.g. for a cosine variation

$$v(t,z) = Re(V(z)e^{j\omega t})$$

$$=Re[V^+e^{-az}e^{j(\omega t-\beta z)}+V^-e^{az}e^{j(\omega t=\beta z)}]$$

$$=I\ V^+\ Ie^{-az}\cos(\omega t-\beta z+\phi^+)$$

$$+I\ V^-\ Ie^{az}\cos(\omega t+\beta z+\phi^-) \qquad .(9)$$

assuming $V^+ = I\ V^+\ Ie^{j\phi+}$ and $V^- + I\ V^-\ Ie^{j\phi-}$

The V^+ and V^- terms in (7) and (9) may be interpreted as the incident and reflected wave components making up the total voltage across the line. Likewise with the I^+ and I^- components in (8).The incident wave component in (9) propagates down the line in the positive z direction at a velocity v_p, known as the phase velocity:

$$v_p = \frac{\omega}{\beta} = \frac{2\pi f}{2\pi / \lambda} = f\lambda \qquad .(10)$$

with an amplitude decaying with distance according to e^{-az}.The reflected wave propagates in the negative z axis direction at the same velocity and with the same exponential decay in amplitude.

In the case of the loss free line and the 'distortionless' line (condition RC = LG), $v_p = 1/\sqrt{(LC)}$ and is constant for all frequencies. However, in general v_p depends on frequency and signals at different frequencies will travel at different velocities on the line. Thus if a group of waves whose frequencies lie between ω and $\omega+d\omega$ is considered, then the resultant amplitude envelope of the group, which carries the energy contained in the signals, travels down the line at a velocity v_g different from v_p.v_g is known as the group velocity;

$$v_g(\omega) = \frac{d\omega}{d\beta} \qquad .(11)$$

2 CHARACTERISTICS AND PARAMETERS OF TERMINATED LINES

2.1 Reflection and transmission coefficients

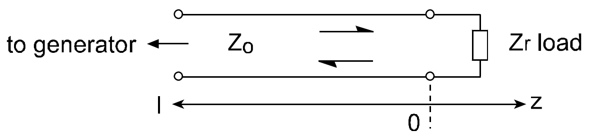


Figure B-2

Consider the conditions at the load end of line terminated in an impedance $Z_T \neq Z_0$. For convenience, let the z axis origin coincide with the plane of this load. Then the general solutions:

$$V(z) = V^+e^{-\gamma z} + V^-e^{+\gamma z}, I(z) = \frac{V^+}{Z_0}e^{-\gamma z} - \frac{V^-}{Z_0}e^{+\gamma z}$$

reduce to $V(0) = V^{+} + V^{-}, I(0) = \frac{V^{+} - V^{-}}{Z_0}$ at the load, z = 0;

but the ratio of voltage to current at the plane of the load equals the load impedance, thus:

$$\frac{V(0)}{I(0)} = Z_T = Z_0 \frac{V^{+} + V^{-}}{V^{+} - V^{-}} \qquad .(16)$$

We define the ratio of the reflected wave voltage V^- to the incident wave voltage V^+ at the load as the reflection coefficient, $\rho_0 = |\rho_0| e^{j\phi 0}$

Thus from (16): $$\frac{Z^T}{Z_0} = \frac{1 + \frac{V^-}{V^+}}{1 - \frac{V^-}{V^+}} = \frac{1+\rho_0}{1-\rho_0}$$

hence $$\rho_0 = \frac{V^-}{V^+} = \frac{Z_T - Z_0}{Z_T + Z_0} \qquad .(17)$$

is non-zero unless $Z_T = Z_0$

The ratio of the voltage across the load to the incident wave voltage is defined as the transmission coefficient, τ :

$$\tau = \frac{V^{+} + V^{-}}{V^{+}} = 1 + \rho_0 = \frac{2z_T}{Z_T + Z_0}$$

The ratio of the reflected to incident voltage waves at any distance ℓ from the load is

$$\rho(\ell) = \left[\frac{V^- e^{+YZ}}{V^+ e^{-YZ}}\right]_{Z=-\ell} = \frac{V^-}{V^+} e^{-2Y\ell} = \rho_0 e^{-2Y\ell}$$

$$= \left[|\rho_0| e^{-2a\ell}\right] e^{j(\phi 0 - 2\beta\ell)} \qquad .(18)$$

2.2 Standing waves and voltage standing wave ratio

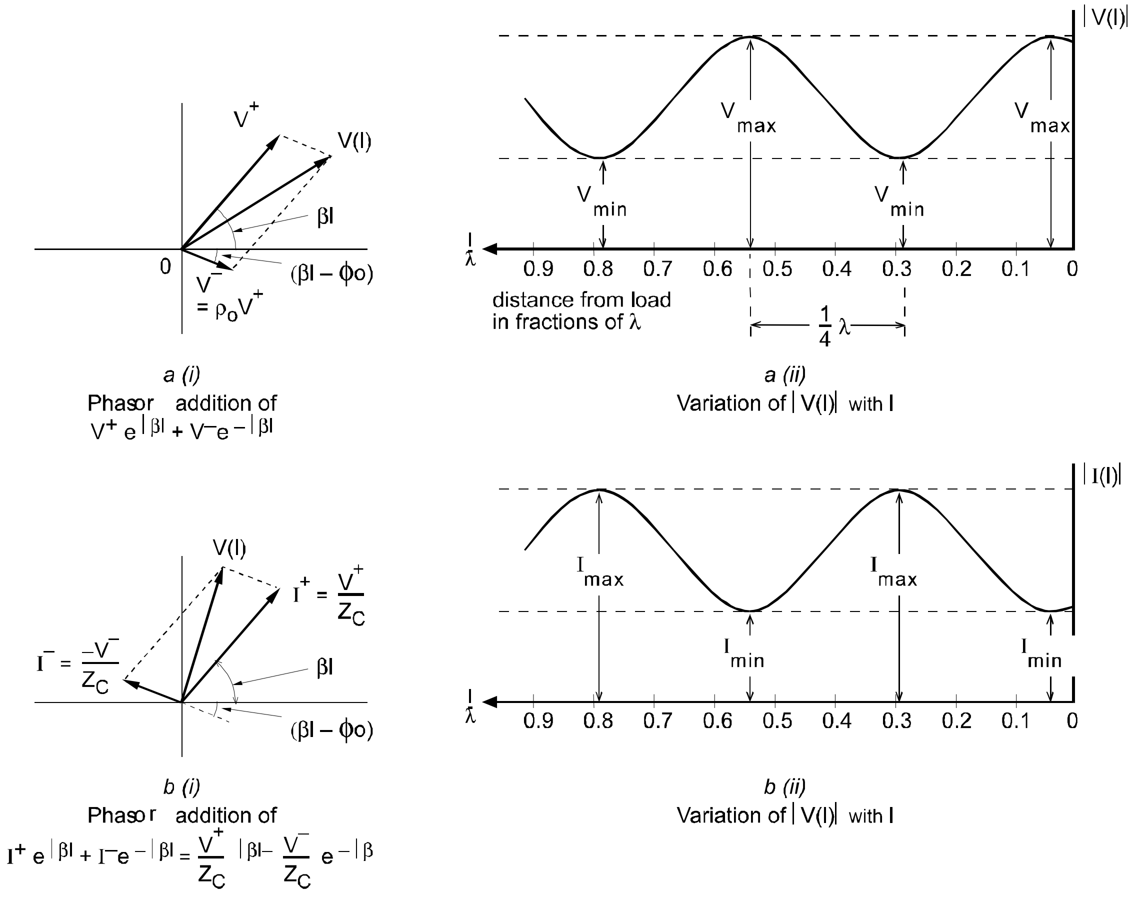


Figure B-3

The voltage and current solutions on a mismatched line terminated in $Z_T \neq Z_0$ at a distance / from the load are (neglecting loss in the line, i.e. $\gamma = j\beta$):

$$V(\ell) = V^{+}e^{j\beta\ell} + V^{-}e^{-j\beta\ell} = V^{+}\left(e^{j\beta\ell} + |\rho_0| e^{-j(\beta\ell-\phi 0)}\right) \qquad .(19)$$

$$I(\ell) = \frac{V^{+}}{Z_0}e^{j\beta\ell} - \frac{V^{-}}{Z_0}e^{-j\beta\ell} = \frac{V_{+}}{Z_0}\left(e^{j\beta\ell} - |\rho_0| e^{-j(\beta\ell-\phi 0)}\right) \qquad .(20)$$

where:

$$\frac{V^{-}}{V^{+}} = \frac{Z_T - Z_C}{Z_T + Z_C} = \rho_0 = |\rho_0| e^{j\phi 0}$$

is the reflection coefficient at the load.

The resultant amplitude of line voltage and current may be found by the phasor addition of the incident and reflected wave components as shown in Figure B-3 (a) and (b) respectively:

$$|V[\ell]| = V^{+}\left[1 + |\rho_0|^2 + 2|\rho_0|\cos\left(2\beta\ell - \phi_0\right)\right]^{1/2} \qquad (21)$$

$$|I[\ell]| = \frac{V^{+}}{Z_0}\left[1 + |\rho_0|^2 - 2|\rho_0|\cos\left(2\beta\ell - \phi_0\right)\right]^{1/2} \qquad (22)$$

The voltage amplitude has a maximum amplitude V_{max} and the current a minimum amplitude I_{min} at positions on the line where:

$$\cos(2\beta l - \phi_0) = +1,\ 2\beta l - \phi_0 = 2\pi m,\ m = 0, 1, 2, 3\ldots$$

i.e. when $l = \frac{1}{2\beta}(2\pi m + \phi_0 = \frac{1}{2}m\lambda + \frac{\lambda}{4\pi}\phi_0$ as $\beta = \frac{2\pi}{\lambda}$ (23)

and $V_{max} = V^{+}(1 + |\rho_0|)$, $I_{min} = V^{+}(1 - |\phi_0|)/Z_0$ (24)

Also when $\cos(2\beta - \phi_0) = -1$, $2\beta - \phi_0 = (2m + 1)\pi$

$$l = \frac{1}{2}m\lambda + \frac{1}{4}\lambda + \frac{\lambda}{4\pi}\phi_0 \qquad (25)$$

the voltage and current amplitudes are respectively at their minimum and maximum values:

$$V_{min} = V^{+}(1 - |\phi_0|),\ I_{max}(1 + |\rho_0|)/Z_0 \qquad (26)$$

The variation of voltage and current amplitudes with distance l as defined by (21) and (22) give rise to what is known as a standing wave pattern on the line. The term standing wave qualifies the fact that the positions of voltage and current maximum and minimum always occur at the same positions on a loss-less line. A pure standing wave occurs on the line when $|\rho_0| = 1$, in which case $V_{min} = V_{max} = 0$. If $|\rho_0| \neq 1$ a partial standing wave is said to exist on the line. When $|\rho_0| = 0$, i.e. $Z_T = Z_0$, there no standing waves set up since no reflection occurs at the load. In this case maximum power is transferred to the load and this is the condition we wish to achieve in practice.

A standing wave on a line may be quantified by 2 parameters. The first fixes the position of the standing wave pattern relative to the termination. This may be found experimentally by location a voltage minimum, for example. This second parameter is used as a measure of the degree of mismatch produced by the load, and this parameter is known as the Voltage Standing Wave Ratio, normally abbreviated to VSWR and denoted by the symbol S. It is defined as:

$$S = \frac{V_{max}}{V_{min}} = \frac{I_{max}}{I_{min}} \qquad .(27)$$

On substituting for Vmax and Vmin using (24) and (26), we have

$$S = \frac{1+|\rho_0|}{1-|\rho_0|} \qquad .(28)$$

$$\text{and } |\rho_0| = \frac{S-1}{S+1} \qquad .(29)$$

Also at positions on the line where the voltage is a maximum, the corresponding current is a minimum and we have:

$$V = V^+ (1 + |\rho_0| e^{-j(2\beta - \phi 0)}) e^{j\beta l} = V^+ (1 + |\rho_0|) e^{j\beta l} = V_{max} e^{j\beta l}$$

$$I = \frac{V^+}{Z_0} (1 - |\rho_0| e^{-j(2\beta - \phi 0)}) e^{j\beta l} = \frac{V^+}{Z_0} (1 - |\rho_0|) e^{j\beta l} = I_{max} e^{j\beta l}$$

since $(\beta l - \phi 0) = 2m\pi$, $e^{-j(2\beta - \phi 0)} = 1$ and therefore the effective input impedance looking in at these positions along the terminated line is:

$$Z_{in} = \frac{V}{I} = \frac{V_{max}}{V_{min}} = Z_0 \frac{1+|\rho_0|}{1-|\rho_0|} = Z_0 S \qquad .(30)$$

i.e. at positions of voltage maximum the input impedance is S times the characteristic impedance of the line.

Likewise at positions of voltage minimum, the current is at a maximum and the effective input impedance at these positions is:

$$Z_{in} = \frac{V}{I} = \frac{V_{max}}{V_{min}} = Z_0 \frac{1-|\rho_0|}{1+|\rho_0|} = \frac{Z_c}{S} \qquad .(31)$$

2.3 Input impedance of a terminated line

The input impedance of a length l of line terminated in an impedance Z_T is defined as:

$$Z_{in} = \frac{V(l)}{I(l)} \qquad (32)$$

where $V(l) = V^+(e^{\gamma l} + \rho_0 e^{-\gamma l})$ . . . the line voltage at l

$$I(l) = \frac{V^+}{Z_c}(e^{\gamma l} + \rho_0 e^{-\gamma l})$$ . . . the line current at l

and $\rho_0 = \frac{Z_T - Z_0}{Z_T + Z_0}$ is the reflection coefficient at the load.

On substituting the above expressions for V(l) and I(l), we obtain:

$$Z_{in} = Z_0 \frac{e^{\gamma l} + \rho_0 e^{-\gamma l}}{e^{\gamma l} - \rho_0 e^{-\gamma l}} \qquad (33a)$$

$$= Z_0 \frac{1 + \rho_0 e^{-\gamma l}}{1 - \rho_0 e^{-\gamma l}} = Z_0 \frac{1 + \rho(l)}{1 - \rho(l)} \qquad (33b)$$

where $\rho(l) = \rho_0 e^{-2\gamma l}$ . . . the reflection coefficient (ratio of the reflected voltage wave to the incident voltage wave) at l from the termination.

To obtain a general formula for Z_{in} in terms of Z_T, γ, Z_c we substitute $\rho_0 = (Z_T - Z_0)/(Z_T - Z_0)$ into (33a), then:

$$Z_{in} = Z_0 \frac{(Z_T + Z_0)e^{\gamma l} + (Z_T - Z_0)e^{-\gamma l}}{(Z_T + Z_0)e^{\gamma l} - (Z_T - Z_0)e^{-\gamma l}}$$

$$= Z_0 \frac{Z_T(e^{\gamma l} + e^{-\gamma l}) + Z_0(e^{\gamma l} - e^{-\gamma l})}{Z_T(e^{\gamma l} - e^{-\gamma l}) + Z_0(e^{\gamma l} + e^{-\gamma l})}$$

$$= Z_0 \frac{Z_T \cosh\gamma l + Z_0 \sinh\gamma l}{Z_T \sinh\gamma l + Z_c \cosh\gamma l} = Z_0 \frac{Z_T + Z_c \tanh\gamma l}{Z_0 + Z_T \tanh\gamma l} \qquad (34)$$

2.4 Normalised impedances

It is often common practice to specify impedances in transmission line problems as a fraction of Z_0, the characteristic impedance of the line. If this is done the impedances are known as normalised impedances and are written in small type. The normalised input impedance of a terminated line is then given by:

$$z_{in} = \frac{Z_{in}}{Z_0} = \frac{\frac{Z_T}{Z_0} + \tanh\gamma l}{1 + \frac{Z_T}{Z_0} + \tanh\gamma l} = \frac{z_T + \tanh\gamma l}{1 + z_T + \tanh\gamma l} \qquad .(35)$$

where $z_T = \frac{Z_T}{Z_0}$ is the normalised load impedance.

Normalisation of impedances and admittances has important practical advantages. For example it allows problems, regardless of different Z_c values, to be worked out graphically by use of the Smith chart.

2.5 Low-loss line approximation, half and quarter wave transformers

For loss-less and low-loss for which αl is sufficiently small to be neglected, $\gamma = j\beta$ and $\tanh\gamma l = \tanh j\beta l = j \tan \beta 1$. For these cases (34) reduces to:

$$= Z_{in} = Z_0 \frac{Z_T + jZ_0 \tan\beta l}{Z_0 + jZ_T \tan\beta l} \qquad .(36)$$

Now since $\tan \beta l$ is a periodic function, Z_{in} is also periodic. That is:

$Z_{in}(l) = Z_{in}(l + {}^1/_2 m\lambda)$, m = 1,2,3,4 . . .

In particular if $l = {}^1/_2 m\lambda$):

$$= Z_{in}(\tfrac{1}{2} m\lambda) = Z_0 \frac{Z_T + jZ_0 \tan m\pi}{Z_0 + jZ_T \tan m\pi} \qquad .(37)$$

Thus a single or multiple half wavelength of line acts as a 1 : 1 transformer. A single half wavelength of line is often referred to as a half wave transformer.

Also if l = (2m + 1)$^1/_4$, m = 0, 1, 2, 3 . . .

$$\tan\beta l = \tan\frac{\pi}{4}(2m-1) \rightarrow \infty$$

Thus:

$$Z_{in}\left\{(2m+1)\,\tfrac{1}{4}\lambda\right\} = Z_0\left[\frac{Z_T + jZ_0\tan\beta l}{Z_0 + jZ_T\tan\beta l}\right]_{l=(2m+1)(\lambda g/4)} = \frac{Z_0^2}{Z_T} \quad .(38)$$

Thus a quarter wavelength or an odd number of quarter wavelengths of line acts as an impedance ‘inverter’. A single $^1/_4\lambda$ section of line is often referred to as a quarter wave transformer. The $^1/_4\lambda$ transformer finds important applications as an impedance matching device.

Notes

1 INTRODUCTION

The Smith chart is extremely useful for solving transmission line problems at RF and microwave frequencies by graphical means.

The charts may be used to:

1 Determine normalised impedances and admittances at any position on a line by knowing or measuring:

(i) the load impedance z = r + jx
or load admittance y = g = jb
where r = normalised resistance
j = normalised reactance
g = normalised conductance
j = normalised susceptance

(ii) The VSWR on the line and a reference position such as a voltage minimum.

2 Plotting the variation with frequency of line impedances.

3 Designing impedance matching networks.

4 Designing RF and microwave circuits including amplifiers and oscillators.

A simplified form of the chart indication the essential features is shown in Figure C-1, the full form is given in Figure C-2.

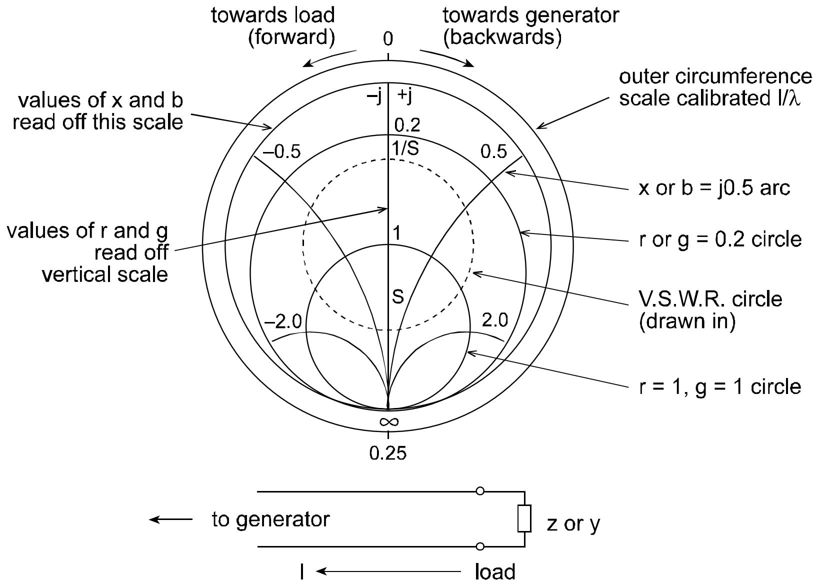
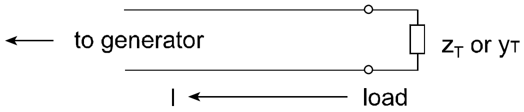


Figure C-1: Simplified form indicating essential features of the Smith Chart



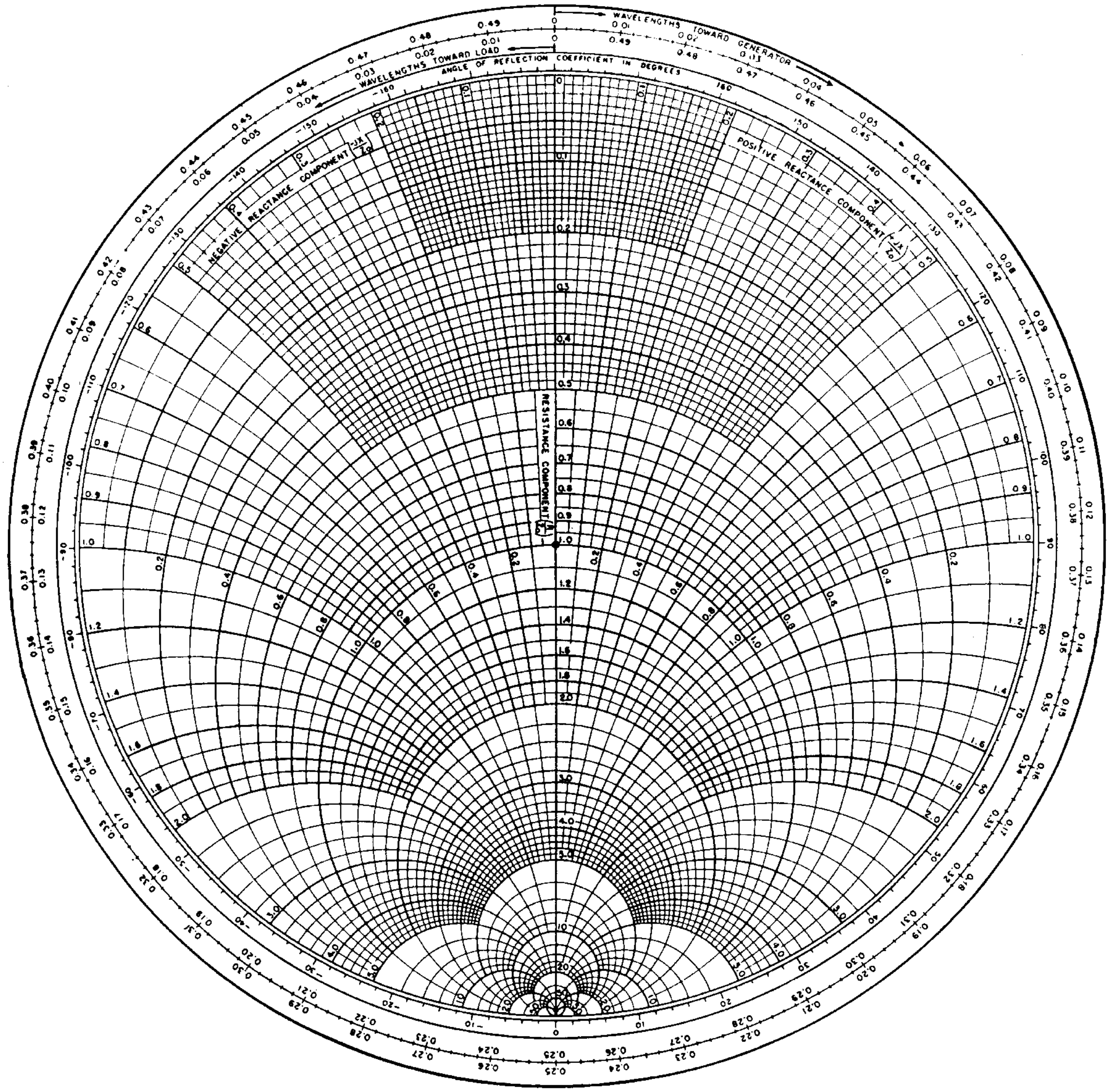


Figure C-2: Smith Impedance Chart

2
Use of Smith chart to determine the input impedance and admittance of a line given the normalised load impedance

The normalised load, $Z_T/Z_O = z_T = r_T + jx_T$ or $Y_T/Y_O = y_T = g_T + jb_T$) is first located on the chart. In Figure C-3 we take this point to be A. The VSWR. or S circle is then drawn by describing a circle centre (1, 0), i.e. point M and radius MA. If required, the value V.S.W.R. on the line may be read off where this circle cuts t he vertical r scale, i.e. S corresponds to point B.

The input impedance, $z_{in}(l) = r + jx$ may now be found as follows:

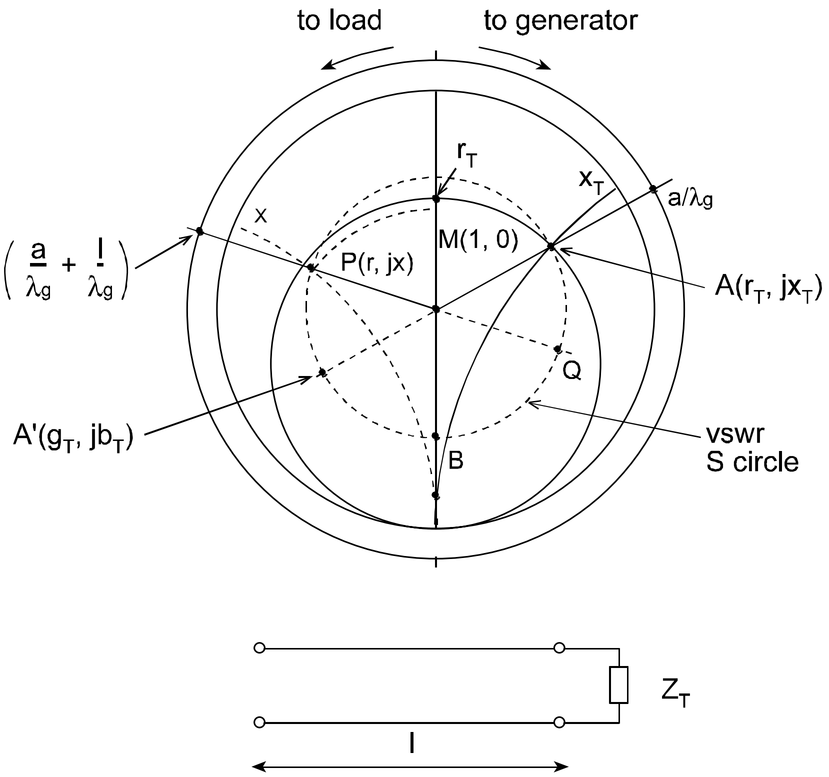


Figure C-3

(1) Draw a line through M passing through the load point A to outer l/λ_g scale. Read off the value a/λ_g say, where this line cuts the l/λ_g scale.

(2) Add l/λ_g to this value and locate the new value $(a/\lambda_g + l/\lambda_g)$ in the clockwise (towards generator) direction. Join this point to the centre M.

(3) The normalised input impedance corresponds to the coordinates of point P where the line drawn in (2) cuts the S circle.

(4) Note that y_T corresponds to point A' and $y_{in}(l) = g + jb$ corresponds to point Q. In general, normalised admittance and impedance points are always diametrically opposite points on the S circle ($^1/_4\lambda_g$ transformer principle).

Try this example

Given the load impedance $Z_T = 0.4 + j0.5$ and $\lambda_g = 34$ mm, determine the VSWR on the line and the normalised impedance and admittance for a line of length $l = 10.5$ mm.

Solution

(1) $a/\lambda g = 0.082$

(2) $(a + l)/\lambda g = 0.391$

(3) $Z_{in} = 0.47 - j0.7$

(4) $y_{in} = 0.66 + j1.0$

3

Determination of impedance and admittance knowing the VSWR and position of voltage minimum on the line.

In this case, the S circle can be drawn immediately by locating S on the vertical r axis and describing a circle centre M(1, 0) radius S as shown in Figure C-4. Then referring to this figure:

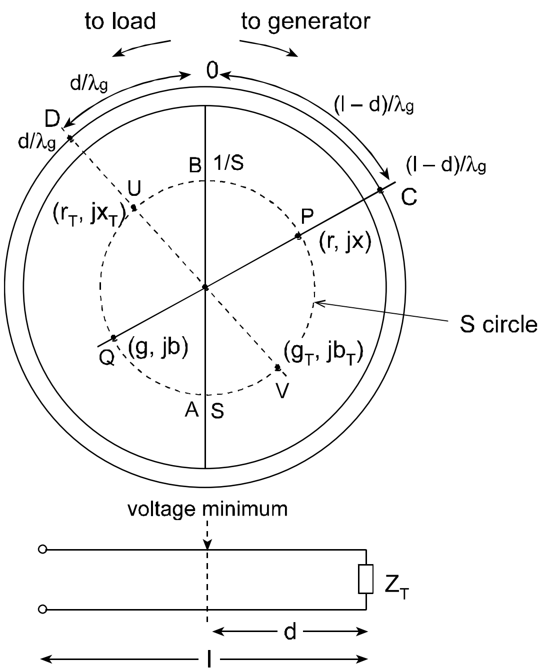


Figure C-4

(1) Point A (S, 0) refers to positions on the line of voltage maximum where z_{in} = S, point B (1/S, 0) refers to positions of voltage minimum where z_{in} = 1/S.

(2) To find zin(l) locate position (l –d)/ λ_g on outer wavelength scale (point C in fig A3-4). Join MC. Co-ordinates of point P give $z_{in}(l) = r + jx$, point Q gives $y_{in}(l) = g + jb$.

(3) To find normalised load impedance zT locate point Dd/λg towards load with respect to B. Join DM. Point U gives (r_T, jx_T), point V gives normalised load admittance $y_T = (g_T, jb_T)$.

Try this example

The VSWR and guide wavelength measured on a line are S = 3.2, λ_g = 34; the first voltage minimum occurs at a distance of d = 4 mm from the load. Determine the normalised load impedance and the normalised input impedance and admittance at a distance l = 30 mm from the load.

Solution

Load impedance z_T = 1.17 –j1.3

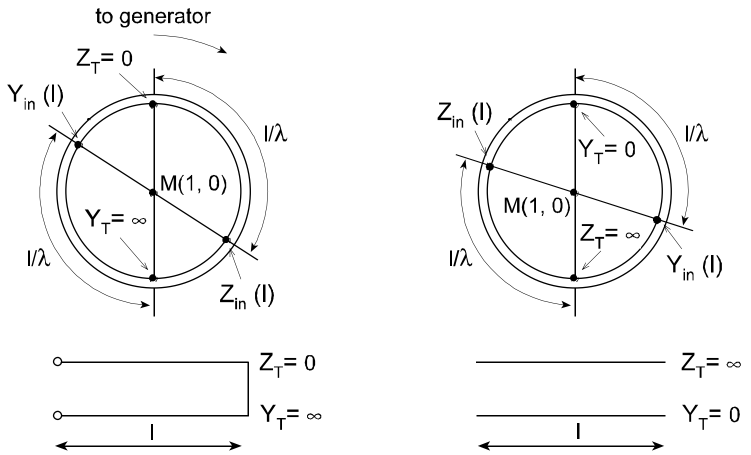
Input impedance and admittance at l = 30mm:

z_{in} = 1.9 + j1.45

y_{in} = 0.34 –j0.26

4
Impedance and admittance of short and open-circuited lines)

For these cases, S = ∞ and the S circle corresponds to the outer circle (r = ∞) on the Smith chart. The input admittances are purely imaginary and their evaluation is intimated in Figure C-5.



(a)
Short-circuited line
z_{in} = jtanβl, y_{in} = −jcot βl

(b)
Open-circuited line
z_{in} = −jcot βl, y_{in} = jtanβl

Figure C-5

Examples

Determine the input impedance and admittance of the following:

(1) A short-circuited line; length l = 15.6 mm, λg = 40 mm

Answers: z_{in} = −j0.83, y_{in} = +j1.21

(2) An open-circuited line; l = 36.2 mm, λg = 320 mm

Answers: z_{in} = −j1.16, y_{in} = +j0.86

(3) The minimum length of short-circuited line which produces a susceptance jb = −0.2; λg = 30 mm

Answer: l = 6.56 mm

5
Application to single stub matching

Short and open-circuit lines (stubs) are extensively used as reactive susceptive elements in the matching of transmission lines to a load termination.

Shunt-stub matching

(1) Plot normalised load admittance y_T, point A in Figure C-6, on Smith chart and draw in the S circle.

(2) Locate point (usually for practical considerations nearest the load to minimise the frequency sensitivity of the match) where y_{in} = 1 + jb, i.e. point B, where the S circle cuts the g = 1 circle, and note its distance u/λ_g from load.

(3) Find length of a short- (i.e. open-) circuit stub which has an admittance ystub = ∓jb and place this element in parallel with the line at u/lg from load. Then at this point the combined input admittance yin _ ystub = 1 and a match is achieved.

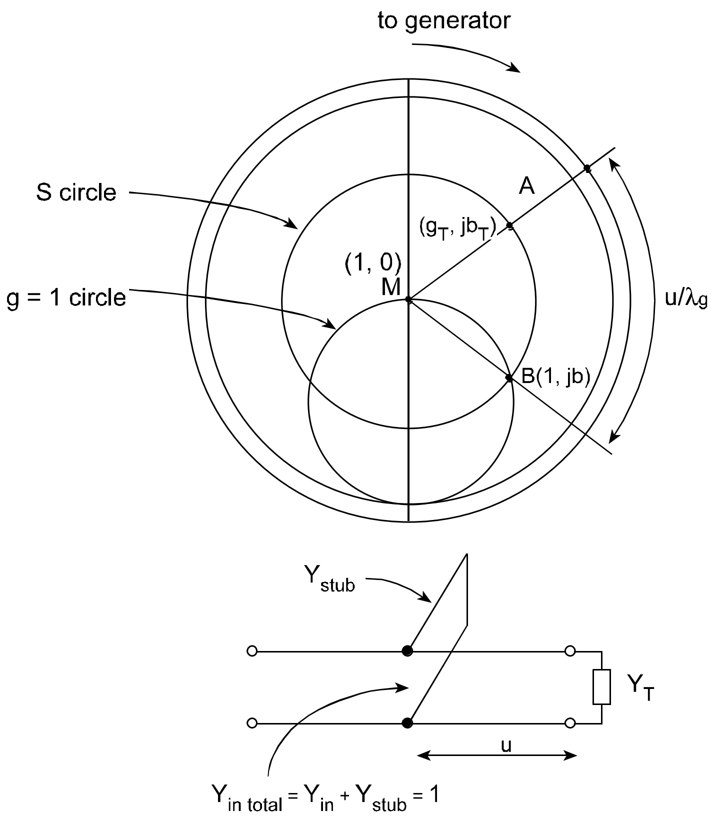


Figure C-6

Examples

(1) Determine the position of a shunt stub and the value of its susceptance which when placed closest to a load of normalise susceptance $y_T = 0.5 + jo$ will produce matching; $\lambda_g = 40$ mm.

Answer: $0.152\ \lambda_g = 6.08$ mm, $jb = -j0.7$

(2) Repeat the case where a VSWR S = 3 is measured on the line and the position of the first voltage minimum d = 5.0 mm from the load, $\lambda_g = 40$ mm

Answer: $0.293\ \lambda_g = 11.7$ mm, $jb = -j1.2$.

Notes