Ahsanullah University of Science and Technology

Department of Electrical and Electronic Engineering

LABORATORY MANUAL
FOR
ELECTRICAL AND ELECTRONIC SESSIONAL COURSES

Course no: - EEE 4174
Course title: - Microwave Engineering Lab
4th Edition • 2021

For the students of
Department of Electrical and Electronic Engineering
4th Year, 1st Semester

Student Name: ______________________________

Student ID: ______________________________

Lab section: ______________________________

Table of contents

Experiment no: **01**
Name of the Experiment: **Introductory Study and basic Measurement of Microwave System**

(a)Objective:

1. To understand briefly the basic principles of microwave system with special emphasis on antenna and waveguide.
2. To learn about various terms related to microwave system such as Radiation Pattern, Polarization, Reflection, Attenuation, Standing wave, Diffraction, Interference etc.

(b)Equipment list:

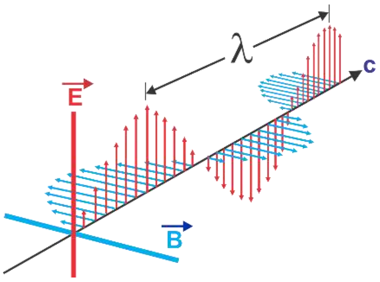
1. Power supply
2. Transmitter and Receiver
3. **Horn Antenna:** The transmitting and receiving both antennas are used for vertically polarized microwave. This wave is generated by FET oscillator mounted vertically. The waveguide takes the wave and open into a horn.
4. **Polarization Grille**: This rectangular metal grille with edges of insulating materials is used to test the plane of polarization. The distance between the metal bars of the grille must be less than half a wavelength of microwave-signal. If grille having bars in vertical position is placed in the path of vertically polarized (E field-vertical, H field- Horizontal) microwave signal, then the signal will be reflected and no wave will transmit through the grill. This occurs because magnetic field cuts across the metal bars, inducing a current in them and causing microwave to be reflected.
5. Baseboard
6. Waveguide
7. Metal/ Plastic/ Hardboard

(c) Theory

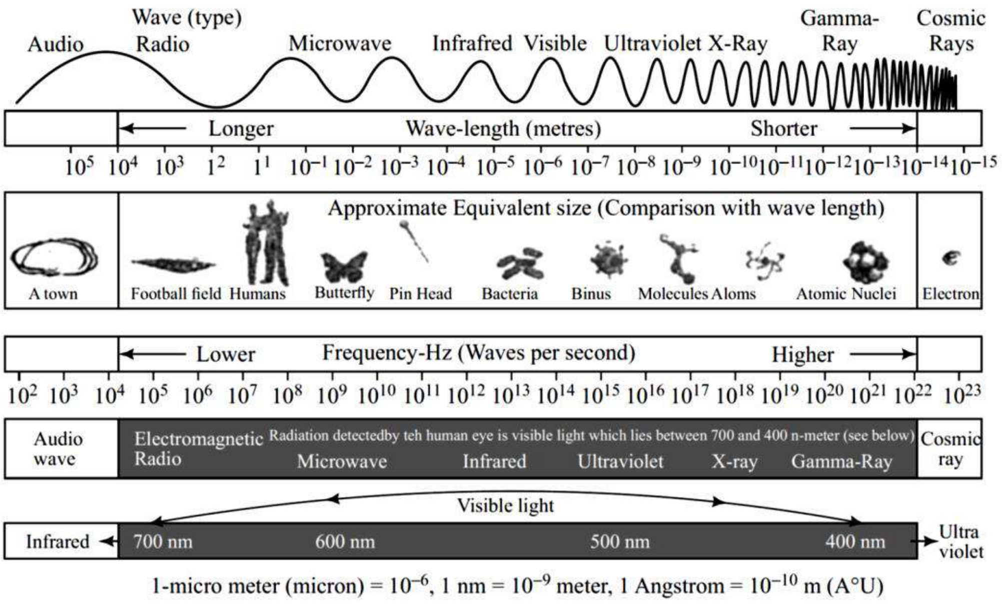
C1. Introduction to Microwave

Microwaves are electromagnetic waves (Figure 1a) that have very high frequency (300 MHz-300 GHz) and very short wavelength (1mm-1m). Satellite communication (for TV, weather forecasting, GPS, voice communication, military), mobile communication, RADARs (used for air traffic control, meteorological, military, guided missile etc.), land phone relay link, heating (household, industrial/ medical), millimeter wave imaging are the main areas where microwave wave are used. Typical microwave bands are L (1-2 GHz), S (2-4 GHz), C (4-8 GHz), K_u (12-18 GHz), K (18-26 GHz), K_a (26-40 GHz) etc.

Bangladesh's very own satellite 'Bangabandhu Satellite-1' has 40 transponders from which use 14 are C band and 26 k_u band. Recently developed 5G technology use LTE frequency range (600MHz to 6GHz) and also millimeter wave (24-86GHz). Verizon 5G UWB use 28 and 39GHz frequency which has 40x bandwidth compared to 700MHz 4G LTE. They also use 700MHz -2500MHz range to cover wide area.



(a)



(b)

Figure 1.1: (a) Electromagnetic wave (TEM mode) (b)Electromagnetic spectrum

C2. Sources of Microwave Signal in Laboratory

There are different types of microwaves signal sources, such as-

(i) Klystron Oscillator

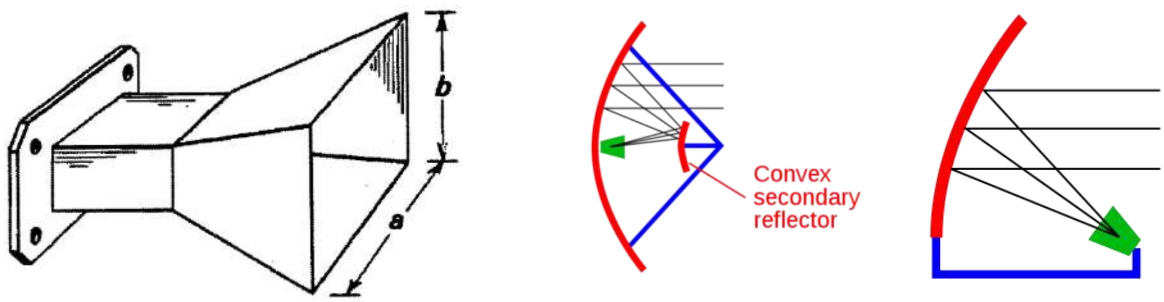
It is vacuum tube type microwave signal source employing velocity modulation and transit time effect. The details of its working principles will be discussed in Experiment 2.

(ii) Gunn Oscillator

It is a solid state type oscillator working on the basis Gunn effect (negatively differential conductivity effect in bulk semiconductor which has two conduction band minima separated by an energy gap. details of its working principles will be discussed in Experiment 2.

C3. Antenna as source of Microwave Radiation

Microwave signal can be transmitted into free space by the radiating structure called antenna. It is a kind of transducer that transforms generated electrical energy (power supplied by feeder) into radiating energy (in the form of electromagnetic waves) into free space. For Transverse Electromagnetic waves (TEM), electric and magnetic fields are perpendicular to each other and also in a plane perpendicular to the direction of wave propagation (Figure1a). In this experiment we will use Horn type antenna.



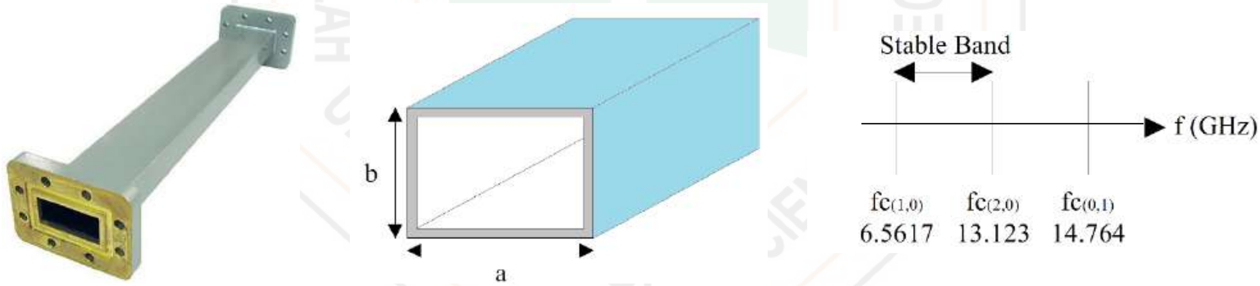
(a) (b) (c)

Figure 1.2: (a) Horn Antenna (b) Cassergrin feed antenna (c)Off axis or offset feed antenna

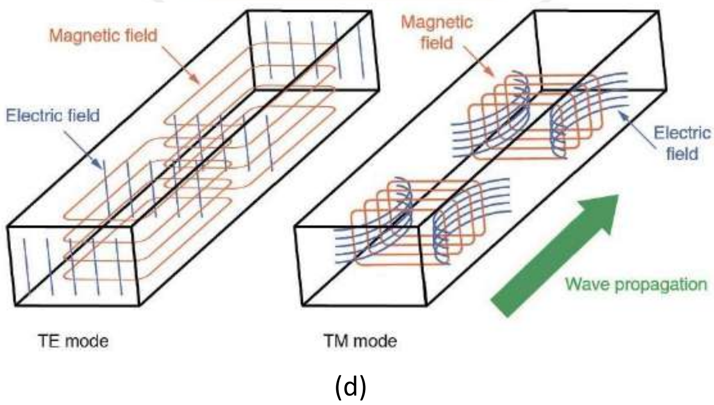
Horn Antenna: It is a kind of aperture antenna consisting of tapered waveguide transition into free space in the form of pyramidal horn Figure 1.2(a). If the impedance matching is perfect, then all the energy travelling from waveguide will be radiated. This antenna is used in satellite Communication and is very important for lab testing because its bore sight gain can be accurately calculated (within 0.1 dB). Most ground based small broadcast satellite receiver dishes use a small horn feed (of low gain) at focal point of dish between 5 to 1-meter diameter (such antennas are known as Cassergrin feed antenna (Figure1.2b). Often this feed is offset from bore sight direction of reflector dish (Figure1.2c). This offset feed arrangement directs the main beam away feed and this result in less and improved side lobe performance. The main beam hit the feed of sub-reflectors, it will be diffracted around the obstacle and radiation will be scattered or diffracted into side lobes. Parallel beams coming from target are reflected by parabola as a convergent beam and reelected by hyperbolic sub-reflector and converge the rays at feed.

C4. Importance of Waveguide in microwave System

Waveguide is specially constructed hollow dielectric inside metal tube for transfer/guidance of electromagnetic energy at very high frequency. It consists of metal conductor which surrounds dielectric, vacuum (Figure 1.3a, b) The metal plates (brass, copper, aluminum etc.) contain the microwave energy and prevent it from spreading outside. Metal plates must be aligned to the electric field to avoid disturbances to field. Metal structure dimensions depend on frequency of microwave signals, Microwave signals are reflected from inside of metal walls and reach destination with minimum loss. In case of microwave signal travels close to ground they are susceptible to reflection, phase shift. This problem is overcome by waveguides. Outside part of Metal plates also eliminates external interference. Waveguides are used in underwater microwave. telephone link systems. There are different modes of propagation of microwave signals through wave guides such as TM (transverse magnetic), TE (transverse electric) etc. (Fig 1.3d). But TEM (transverse electromagnetic wave) cannot exist in a hollow waveguide because it would be impossible for magnetic field to surround transverse electric field without having an axial component and vice versa. The frequency below which the microwave signals are attenuated are called cut-off frequency. Each of these forms of wave may be set up in the waveguide in several modes of propagation such as TE_{10}, TIM_{11} etc.



(a) (b) (c)



(d)

Figure 1.3: (a) Rectangular waveguide (b) Waveguide dimension (c) Stable Band for WR90 Waveguide (d)TE and TM mode of a rectangular waveguide

Cut off frequency: Below a certain frequency the microwaves will be attenuated as they pass along the waveguide. The Frequency bellow which the microwaves are attenuated is called cut off frequency.

$$f_{c(m,n)} = \frac{c}{2}\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2}$$

Here m, n are positive integer including zero and c is the speed of light in vacuum. For TE_{10} mode and WR 90 waveguide with dimensions a= 2.28 cm and b = 1.02 cm the cut off frequency is 6.56 GHz. So in WR 90 waveguide the lowest frequency that can travel down is 6.56 GHz.

For reliable microwave transmission along a waveguide, there should be only one mode of propagation. Waveguides are designed to operate in a single mode over a range of frequency. The frequency range is called **stable band** (Figure 1.3c).

C5. RADAR as detector of Microwave Signal

RADAR (Radio wave detection and ranging) is the most welt known detector of unknown objects. It is developed more or less simultaneously in UK, USA, Germany and France during World War II. In a RADAR, a transmitter send signal (usually radio waves in short rectangular pulses/ large UHF microwave power/particular type of pulse modulated sine wave) in the direction RADAR is pointing at that time. That signal is reflected back from the distant target and echo signal is detected by sensitive receiver Fig 1.4. The receiver receives, analyze and displays the corresponding signal. The received pulse is displayed in oscilloscope which is triggered to start by transmitted pulse. If a narrow beam antenna is used, the target's detection can be accurately given by the position of antenna. Antenna rotates and scans the desired area. In the received mode, the returned signal is amplified, and mixed with the local oscillator to produce the desired IF signal which is then amplified. Such RADARs often use a continuously rotating antenna for 3600 azimuthal coverages (display in polar plots)

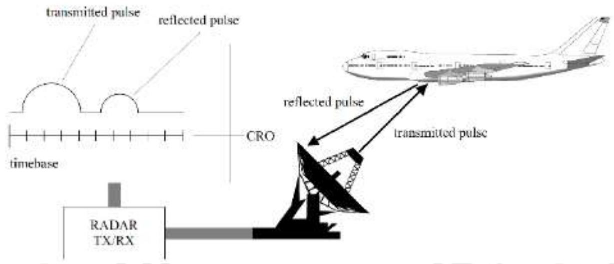


Figure 1.4: Signal reflection used to detect airplane in a RADAR system

C6. Microwave Connectors

There are different types of connectors. But, important one are coaxial connectors. Coaxial cable is a round conducting wire surrounded by insulator spacer surrounded by cylindrical conducting sheath usually surrounded by another insulating sheath. It is used as high frequency transmission fine to carry high frequency or broadband signals. It is not suffered by external EM filed because it EM field exist between inner and outer conductors (see exp. 2 Fig. 2.2)

For operation in microwave frequency region, the average circumference of coaxial cable must be limited to about one wavelength in order to reduce multimodal propagation and eliminate erratic reflection coefficients, power tosses and signal degradation. Microwave coaxial connectors can be classified into different types shown in appendix A2. They are- UHF connector or APC-3.5 (Amphenol Precision Connector 3.5 mm with low VSWR; male or female end of this 50 ohm connector can mate opposite SMA has threaded coupling and it prevents them from being removed accidentally), APC-7 (Amphenol Precision Connector 7 mm that provides coupling mechanism without male or female distinction, VSWR tow frequency range 1.02 to 18 GHz), BNC (Bayonet Navy Connector operates upto 4 GHz then radiate EM energy, accept flexible cable of diameter of 6.35 mm, presence constant impedance 50 to 75 ohm), SMA (Sub miniature A used below 24 GHz because of higher order modest provides RF connectivity between boards and components like fitters, attenuators, mixer, oscillators), SMB (Sub miniature B can be connected and disconnected swiftly, does not require nuts to be tightened when two connectors are mated rather they snap fit together), SMC (Sub miniature C is 50 ohm connector with smaller dimension-dia 3.17 mm than SMA used below 7 GHz) TNC (Threaded Navy Connector has screw fitting and is threaded BNC to stop radiation at higher frequency operates up to 12 GHz, during vibration resistance not changed like BNC and result is less noise), and type N (Navy, is high performance RF coaxial connector which has a threaded coupling interface to ensure that it mates correctly and is able to withstand high powers compared to BNC or TNC, good for 11 GHz).

A Male connector is the one where inner conductor protrudes and Female connector is the one where inner conductor forms a sleeve around its mate counterpart. Alternatively, plug is used for male and jack is for female.

Barrel or male to male adaptor refers to an adaptor with two male ends. Bullet or female to male adaptor refers to an adaptor with two female ends. Connector saver is an adaptor with one male and another female end.

C7. Important parameters related to Microwave Signals

i. Polarization: EM wave contains energy associated with electric and magnetic field (both can be in the direction perpendicular to the direction of wave propagation). Polarization of EM wave is defined as the direction of electric field of the wave. For transverse EM wave (TEM), polarization at any point between transmitting and receiving antenna is usually at right angle to the line of sight joining transmitting and receiving antenna and also perpendicular to direction of magnetic field. Plane/direction of polarization can change with time and distance -this is called rotation of plane of polarization. If electric field is directed vertical, horizontal, vary within one plane or plane of polarization rotates a complete cycle every wavelength then corresponding polarization are called vertical (e.g. Whip, Vertical dipole, Monopole, Discone, Broadside array etc.), horizontal (e.g. Dipole, Square loop, Quad, Vee, Yagi etc.), plane polarization or circular polarization (eg. Helix, Crossed Dipole). Electric field generated by antenna is actually runs parallel to antenna and magnetic field runs perpendicular.

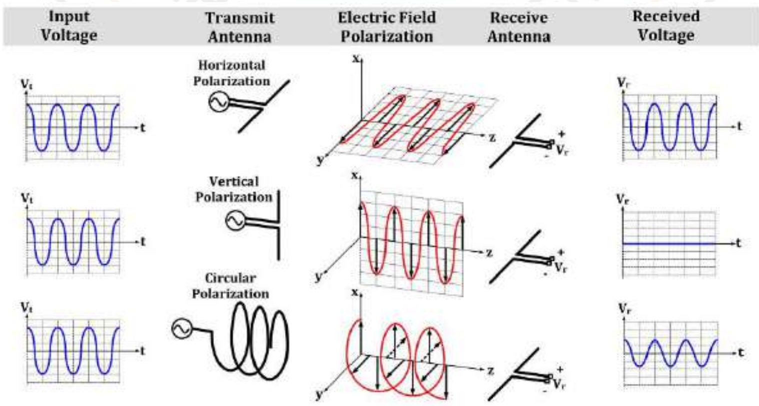
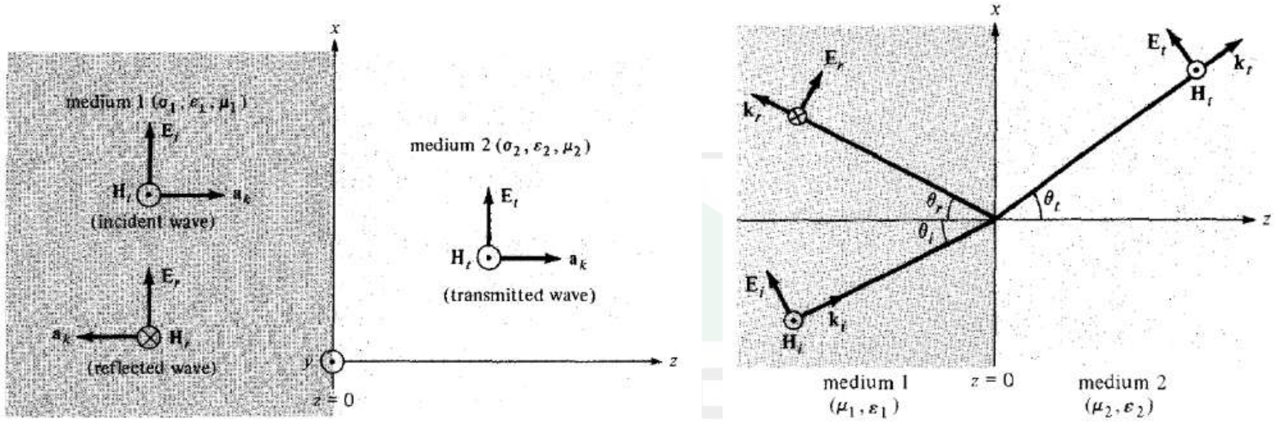


Figure 1.5: Effects of polarization mismatch on coupling transmit/receive antenna pairs.

ii. Reflection: During microwave signal transmission, energy is transmitted by wave and can be reflected off other surfaces of a particular material. Different material reflects in different amount, the degree of reflection by a particular material/load is given by,

$$Reflection\ Coefficient, \Gamma = \frac{Electric\ field\ of\ Reflected\ wave, E_{ro}}{Electric\ field\ of\ Incident\ wave, E_{io}} = \frac{Z_L + Z_0}{Z_L + Z_0}$$

Where, Z_L = Characteristics impedance of second material/load, Z_0=Characteristics impedance of first material



(a) (b)

Figure 1.6: (a)Normal Incident (b)Oblique incidnet of the Plane wave

There are different types of reflection such as: Reflection from normally incident (Fig 1.6a) and obliquely incident (Fig 1.6b) wave. The material can be good conductor or good dielectric.

Another term related to reflection is,

$$Return\ Loss, RL = -20\ log|\Gamma|$$

iii. Attenuation: During microwave signal transmission, if there are objects (of any material) the path of propagation, then signal will decay with distance- this is called Attenuation. For each material, there is a constant which describes the amount by which a microwave beam is attenuated in passing through a unit distance- this is called Attenuation constant. When microwave beam penetrates a material, the electric field intensity of the beam is attenuated exponentially inside the material (Fig 1.7)

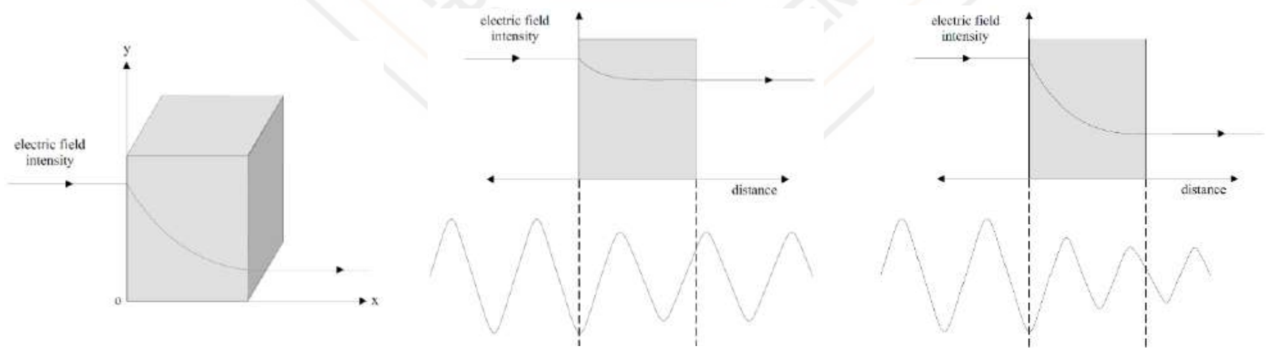


Figure 1.7: Attenuation of Electromagnetic wave due to different material

Attenuation Coefficient $\alpha = 10\log_{10}\frac{P_1}{P_2} = 10\log_{10}\frac{P_1/4\pi r_1^2}{P_2/4\pi r_2^2} = 20\log_{10}\frac{r_2}{r_1}$

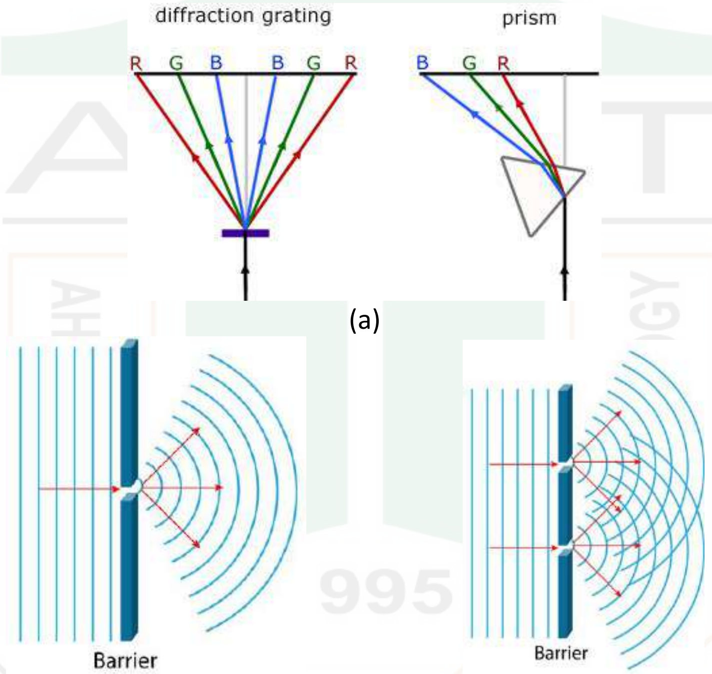
Here, P_t =transmitted power, P_1= signal power at entrance point r_1 of material, P_2= signal power at exit point r_2 of material

Skin depth is defined as the distance at which microwave signal is attenuated by 8.86 dB. Skin depth is a frequency dependent parameter. Typically, in conductors as the frequency is increased the signal penetration is less. So high frequency signal transmission in conductors becomes less efficient.

$$Skin\ Depth = \frac{1}{\alpha}$$

Attenuation problems can be overcome by placing antennas at high altitude/ using straight line of sight communication/ placing large number of repeaters etc.

iv: Diffraction: Diffraction is the spreading of waves when they pass through an opening or round obstacle into regions where we could not expect them. Fig. 1.8a is a light diffraction which is observed in a Prism. A diffraction grading has the same effect as a prism but it diffracts by light hitting the periodic gap.



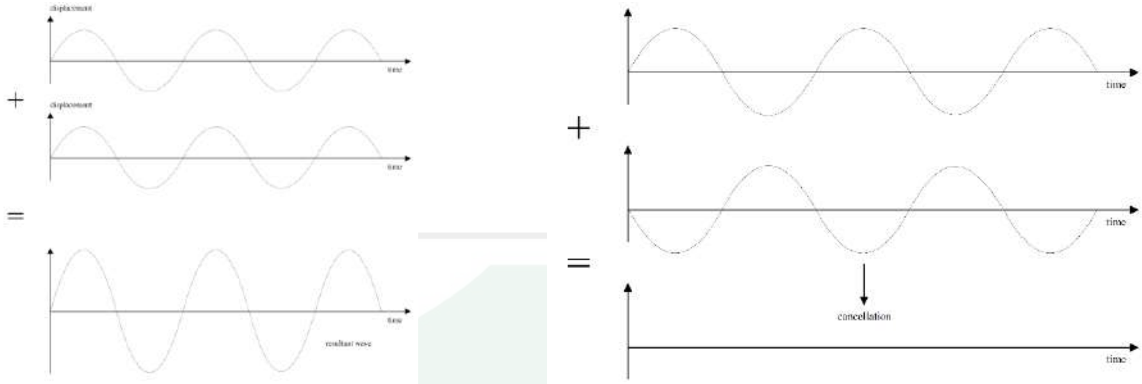
(a)

(b) Diffrection (c) Interference

Figure 1.8: (a)Diffraction in Prism and grating (b)Diffraction of plane wave (c)Interference of wave

Diffraction is based on Huygens's principle which states that, no matter how small a slit is made on an opaque plane, light on the side opposite to the source would spread out in all direction and no matter how small a light source is constructed a sharp shadow cannot be obtained at the edge of sharp opaque obstacle. It is possible to estimate the optimum size of the hole in order to minimize the loss. This kind of diffraction can occur in satellite antenna (shaped like a concave mirror and made of wire mesh to offer less resistance to wind and avoid damage). The hole size in the mesh should be a compromise between wind resistance and leaking of the signal by diffraction. Diffraction lowers the maximum strength of the received signal but also causes a spreading of the microwave signal.

v: Interference: Interference is a phenomenon which occurs when one electrical signal is affected by another unwanted signal. It occurs when two waves from the same source meet at a point while travelling through separate path (Fig 1.8d). High frequency sky waves and microwave propagation are the interesting examples of interference. It actually disturbs the actual strength of the signal. When microwave signals of same frequency superimpose at some point they either reinforce (constructive interference [Fig 1.9a]) or cancel each other (destructive interference [Fig1.9b]). The resulting effect is called *interference pattern*. This pattern was first investigated by Thomas Young (commonly known as Young's slit experiment) using light to fall on a pinhole punched in a screen (Fig. 1.8d). Interference is not quite significant up to VHF because of long wavelength. A RADAR will be completely blind at null points. To solve this, elevation of antenna should be high and made to point downwards.



(a) (b)

Figure 1.9:(a) Constructive interference (b)Destructive interference

D) Experiment procedure:

D1. Radiation Pattern

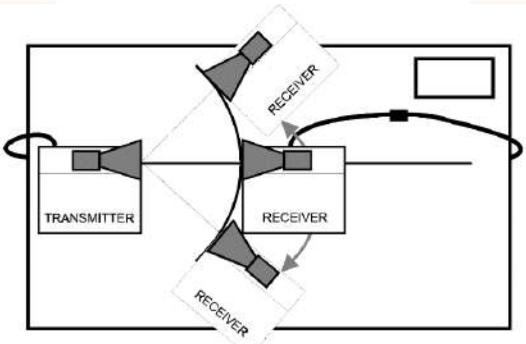


Figure 1.10: Setup for radiation pattern measurement

1. Place the transmitting and receiving antenna in a straight line. Measure the received signal strength.
2. Keeping the transmitter in a fixed position place the receiver in 4/5 position in an arc as shown in Fig. 1.10 and measure the data.
3. Plot the radiation pattern graph from your measured data.
4. Measure beam width from your measured data.
5. Alter the distance between the transmitting and receiving antenna and observe how the strength changes with the distance.

D2. Polarization

1. Set both the transmitting and receiving antenna for vertical polarization. Check the reading at receiving end.
2. Set the transmitting antenna for vertical polarization and receiving antenna horizontal polarization, check the reading at receiving end.
3. Rotate the receiver in the vertical plane while keeping receiving antenna in line with transmitting antenna (Fig 1.11). Locate the positions of maximum and minimum received signal

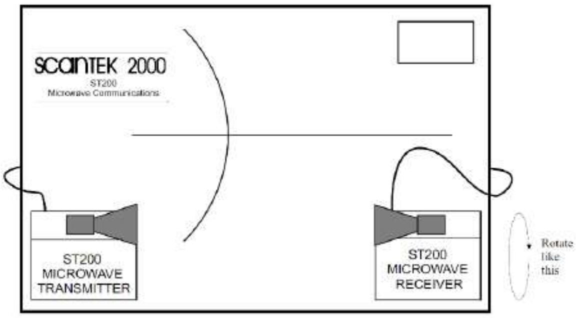


Figure 1.11: Setup for Polarization measurement

Set both the transmitting. and receiving antenna for vertical polarization.

1. Place the grille in the way of vertically polarized microwave signal with the grille bars in the vertical position. Check the reading.
2. Place the grille in the way of vertically polarized microwave signal with the grille bars in the horizontal position. Check the reading.
3. Place the grille in the way of vertically polarized microwave signal with the grille bars in the slanted position. Check the reading.

D3. Microwave Signal reflection

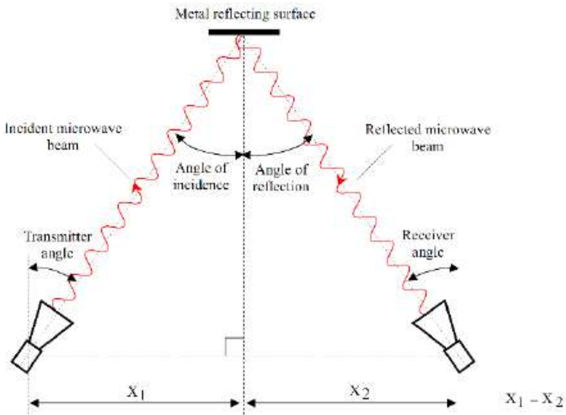


Figure 1.12: Signal reflection measurement

1. Set up the transmitter, receiver and metal reflecting plate according to Fig 1.12. Measure the receiver reading.
2. Change the position of receiver and take reading for the case of (i) angle of incident < angle of reflection, (ii) angle of incident > angle of reflection and (iii) angle of incident = angle of reflection. Check when the reading is maximum.
3. Replace metal by *plastic* and repeat steps 1-2.
4. Use *hardboard* and repeat steps 1-2.

D4. Microwave signal Attenuation

1. Set up the transmitter and receiver for a line of sight (LOS) configuration and place a metal plate in the path of microwave signal. Take reading.
2. Use plastic and take reading.
3. Use hardboard and take reading.
4. Use book, damp paper towel, dry paper towel and take reading.

5. Observe the effect of thickness of material, take reading. Calculate attenuation constant and skin depth.

D5. Effect of Waveguide

1. Set up the transmitter and receiver
2. Observe received signal strength at receiver without placing waveguide between them
3. Observe received signal strength at receiver placing waveguide between them.
4. Measure the length of a and b of the waveguide and calculate cut off frequency for TE_{10} mode.

D6. Demonstration of Diffraction

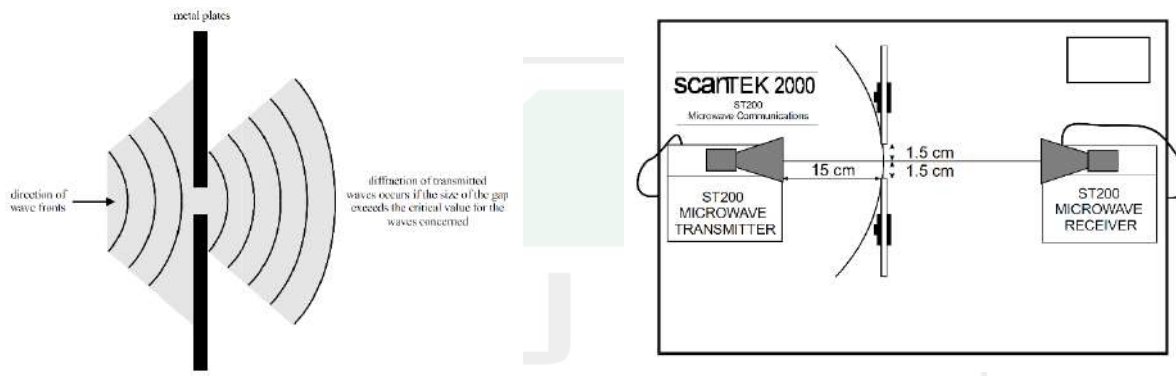


Figure 1.13: Setup for diffraction measurement

1. Set the transmitter, receiver and an obstacle and hole/opening arrangement suitable for diffraction between them as shown as Fig 1.13.
2. Vary the size of the opening and measure the strength of received signal.

D7. Demonstration of interference

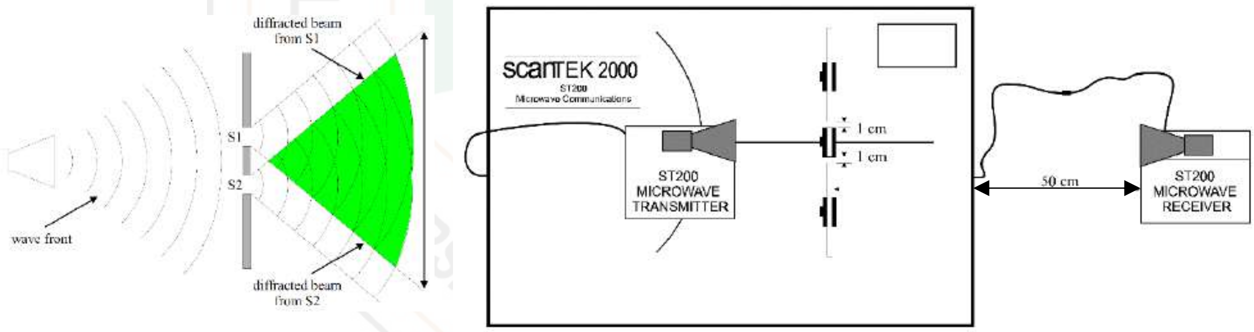


Figure 1.14: Setup for interference measurement

1. Set the transmitter, receiver and an obstacle and hole/opening arrangement suitable for interference between them as shown Fig 1.14.
2. Vary the position of receiver in a line parallel to obstacle plane arrangement and determine the position of maximum and minimum signal strength.

e) Report:

1. Comment on the results observed in the experiment.
2. Discuss about different microwave signal related terms in a very brief manner.
3. Give practical example where polarization, attenuation, diffraction, interference, reflection becomes important in case of a wireless communication system.
4. What is the function of horn antenna and polarization grille in the experiment?

Experiment no: 2

Name of the Experiment: Characterization of microwave Oscillator, Introduction to Spectrum Analyzer and Power Measurement techniques

2.1 Objective:

1. Understand the theory, operation and characteristics of microwave oscillators (Klystron and Gunn oscillator).
2. To learn about different ways of measuring microwave power.
3. To learn about different microwave power sensor.
4. Introduction to Spectrum analyzer, spectrum monitoring and ACPR measurement.

2.2 Equipment:

1. **Klystron and Gunn oscillator power supply**
2. **Microwave Oscillator**

(a) **(b)** **(c)**

Fig. 2.1: (a)Klystron Oscillator (b)Gunn Oscillator (c)PIN Modulator

3. **PIN Modulator:**

   PIN diodes are mounted as shunt elements between the RF transmission path and ground. The transmission path has a characteristic impedance of 50 ohms. When the PIN diodes are forward-biased the equivalent diode resistance is about 30 ohms and most of the RF energy is absorbed by the diodes instead of propagating down the 50-ohm transmission path. However, when the diodes are reverse- or back-biased the equivalent diode resistance is in the order of thousands of ohms and the microwave currents will flow down the transmission path because diode resistance compared to the 50-ohm path impedance is negligible. Thus by placing the PIN diode in the path of the transmission line and modulating the diode with a modulating frequency the RF energy can be amplitude or pulse modulated. Leaving the diode unbiased could be destructive to the diode when there is a signal flow in the system.

4. **Variable attenuator/Fixed attenuator**

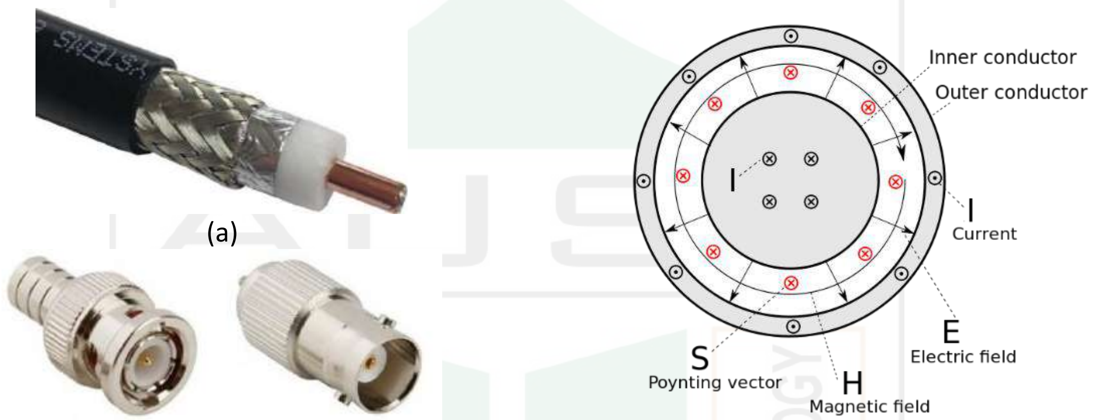
(a) (b)

Fig. 2.2: Variable attenuators (a) Flap attenuator (b) Cavity attenuator

A variable attenuator provides attenuation of microwave signal by varying the degree of insertion of matched resistive strip into waveguide. In our experiment we will use cavity type and flap type attenuator.

5. Coax cable with BNC (Bayonet Neill–Concelman) connector

It provides a match between a waveguide and a 50 ohm coaxial. Coaxial cable is used as high frequency transmission line to carry high frequency or broadband signal. Sometimes DC bias is added to the signal to supply the equipment at other end, as in direct broadcast satellite receivers. Because electromagnetic field exist only in the space between inner and outer conductors, it cannot interfere or suffer interference from external electromagnetic field. The BNC (Bayonet Neil Connector) operates very well at frequencies up to about 4 GHz, beyond that it tends to radiate electromagnetic energy. The BNC can accept flexible cables with diameters of up to 6.35 mm and characteristics impedance of 50 or 75 ohms (Fig. 2.3). Waveguide to Coax adapter was also used for the same purpose.



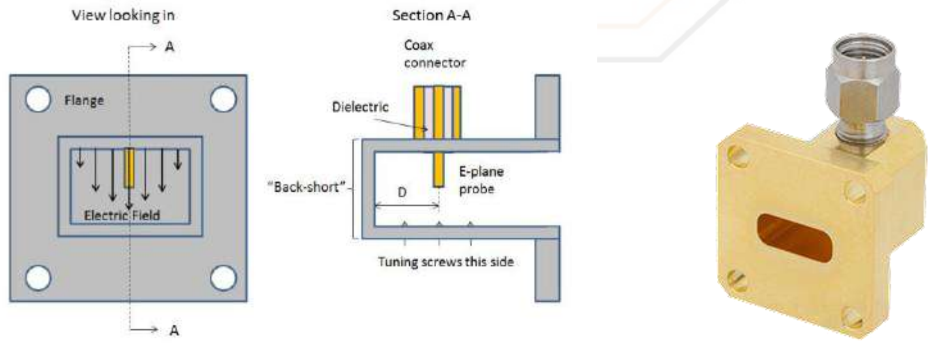
(a)

(c) (d) (b)

Fig. 2.3: (a) Coaxial cable (b) cross sectional view of a coax showing the electric and magnetic field (c) BNC Male connector (d) BNC Female connector

6. Waveguide to Coax adapter

Waveguide-to-coaxial adapters are composed of a waveguide component that fits the waveguide tubing and ends with a flange, and a coaxial probe assembly with a coaxial adapter and connection hardware. The coaxial cable adapter is typically tapped through one wall of the waveguide adapter housing. It provides a match between a waveguide (which might be 200 ohm or any other value) and a 50 ohm coaxial line. The power flow can be in either direction. However, SWR in the adapter should be kept less than 1.2.



(a) (b)

Fig. 2.4: Right angle Waveguide to Coaxial adapter

7. Microwave Power sensor

a) Very low power (0.01–1 mW): Thermocouple/diode sensor technique. (used in Gunn diode)
b) Low power (0.1–10 mW): Bolometer technique. (used in Klystron).
c) Medium power (10 mW–10 W): Bolometer with directional coupler.
d) High-power (10 W–1 kW): calorimetric watt meter

8. Spectrum Analyzer

Spectrum analyzer is a device that measures the amplitude of an input signal versus frequency. By analyzing the spectra of electrical signals, dominant frequency, power, distortion, harmonics, bandwidth, and other spectral components of a signal can be observed that are not easily detectable in time domain waveforms.

The RSA306B Spectrum Analyzer from Tektronix is a modern USB based Real time spectrum analyzer which use USB 3.0 port (due to very high data accusation by the spectrum analyzer) to connect to a powerful PC to analyze data captured by the hardware. The software called SingnalVU-PC can be seen in figure 2.5. As the Spectrum analyzer is fully digital it uses various DSP techniques and software algorithms to show various complex analysis of the captured signal, which cannot be done by any traditional heterodyne spectrum analyzer.

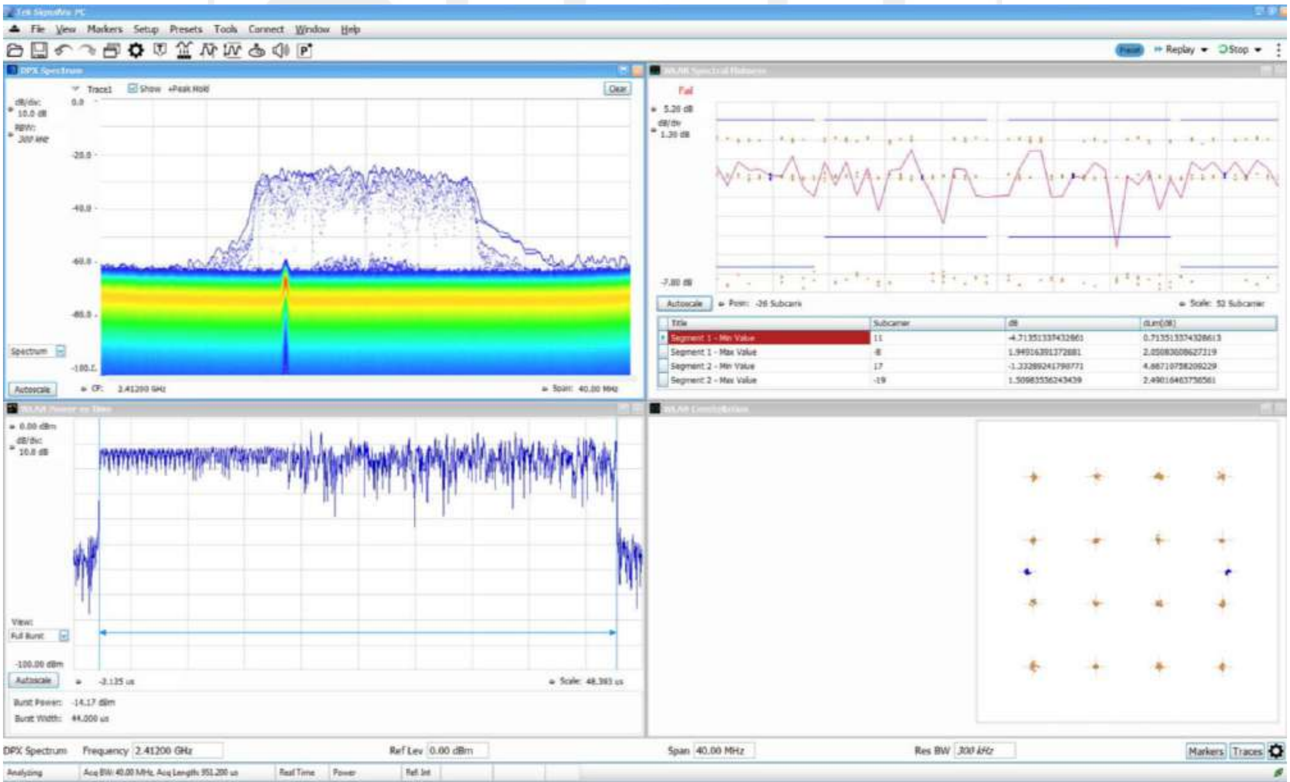


Fig 2.5: SignalVu software showing (Left top): Real time spectrum (Left bottom): Power vs time (Right top) :Spectral flatness (Right bottom): Constellation diagram of a Wireless LAN signal

2.3 Theory:

2.3.1 Microwave oscillator

There are two types of microwave signal source which generates microwave signal - (i) tube sources (such as klystron, magnetron, TWT etc.) and (ii) solid state sources (such as special diodes and transistors). Typically, Solid state sources have higher efficiency but after a certain power low efficiency cavity based oscillators are the main method to generate microwave (Fig 2.6).

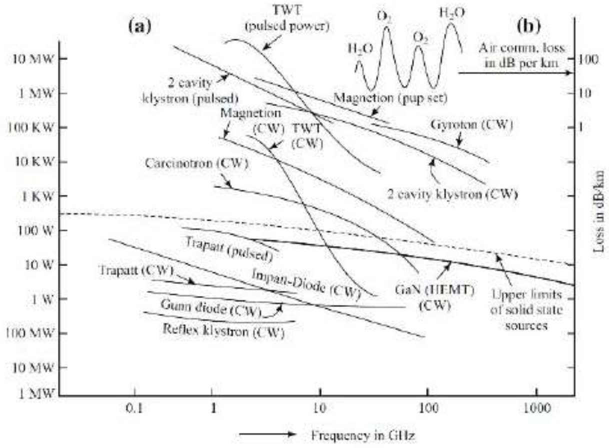
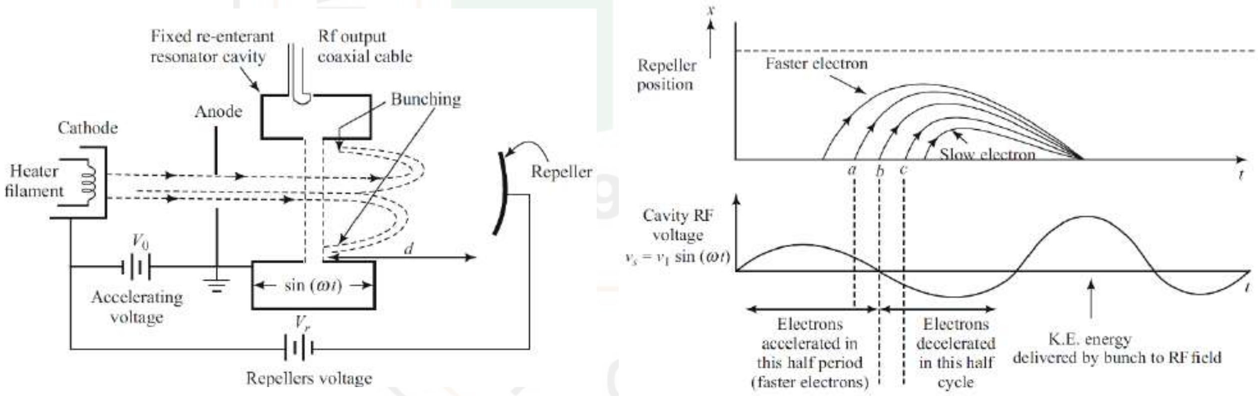


Fig 2.6: (a) Power versus frequency for various microwave sources (b) Air communication loss in dB/km

In communication, there is certain frequency range which is avoided due to high absorption loss due to molecular/atomic natural frequency of oscillation; e.g., oxygen has these frequencies as 69 and 122 GHz, while for water it is 23 and 160 GHz (Fig. 2.6). Rain attenuation limits the range to 5 km, while oxygen limits to 1 km at these frequencies

2.3.1 (a) Klystron Oscillator

Typical single cavity reflex klystron oscillator is low power (less than 2 W) only in the range 1–25 GHz. The most useful characteristic is its tunability over a wide frequency range.



(a) (b)

Fig 2.7: (a) Schematic diagram of a reflex klystron (b) Applegate diagram showing the bunching process

In Fig 2.7(a) it can be seen that from the electron gun, electron beam is accelerated towards anode plate due to its high +ve DC potential (~250-1000V) and crosses through its gap to the cavity resonator. It then crosses the cavity gap mesh grid also (due to momentum) and moves towards the repelled, but never reaches it. It gets repelled back to the cavity due to its high −ve potential (~100-1000V). By this time, the modulation (bunching) becomes still more sharper, i.e. denser. If this bunch of electron returns back at the first wall of the cavity at the moment when it has +ve RF voltage, then the bunch of electrons loses energy by transferring to the RF field, which then gets amplified (positive feedback process). This leads to sustained RF oscillations. Then the amplified signal can be tapped out by probes or loops.

The bunching process: The electron 'a' (see Fig 2.7b) coming out of the gap is accelerated during its +ve cycle, the electron 'c' retarded during −ve cycle, while the 'b' electron comes out with no change in speed when $V_S =$

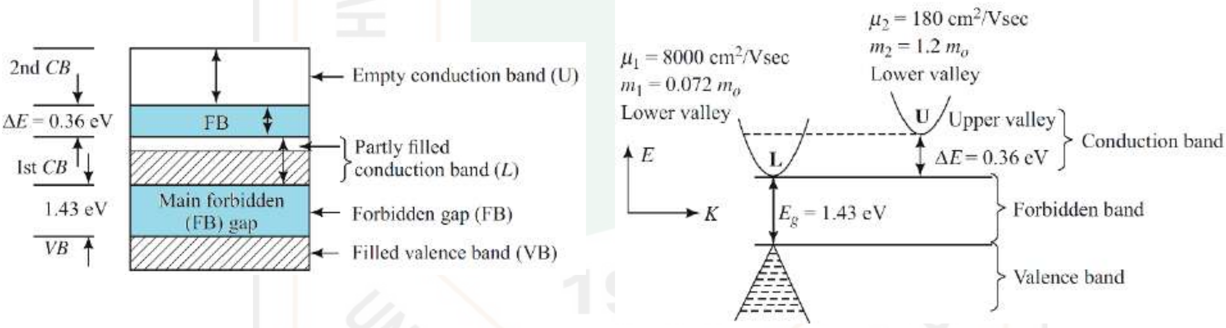
$V_1 \sin \omega_0 t = 0$ .The faster electron 'a' travels deeper in the *repeller space* ('d' in fig 2.7a) as compared to slower electron 'c' and forms bunches. As the bunch is to be formed exactly at the resonator cavity point, we have to adjust the repeller space and repelled voltage (V_r). This way the returning (b) and retarded (c) electron bunch losses its kinetic energy and gives to the RF field in the cavity. When the energy delivered by a certain bunch of electron to the cavity is greater than the energy it had collected (while it was crossing forward), then the oscillation of that frequency signal is sustained.

As in any electronics system noise is always present with all frequencies 0 to ∞ and if there is a +ve feedback system which keeps on amplifying only that frequency signal, which corresponds to the cavity resonant frequency, then that frequency signal gets generated (oscillation is sustained according Barkhausen stability criterion). Here, if the repelled electron bunch reaches back to the cavity at some phase point where it left it, then it is a +ve feedback.

So for oscillation to sustain (i) The reflected back electrons should form bunch just at the time they reach the cavity. (ii) At this moment, the signal at the cavity is to be in +ve phase. Thus, the start of the central part of the bunch is when $V_s = V_1 \sin \omega_0 t = 0$ and reaching back time is when $V_s = V_1 \sin \omega_0 t = V_1$ (Peak value) for optimum efficiency.

2.3.1(b) Gunn Oscillator

Gunn oscillator is used in medium power and low to high frequency application such as 1.0-100GHz microwave receiver, air traffic control transponder (pulsed mode), CW RADAR source, pump source in parametric amplifier etc.



(a) (b)

Figure 2.8: (a) Band diagram of n-GaAs (b) The two-valley model of the n-GaAs conduction band at no bias condition

Gunn (1963) discovered microwave oscillators in GaAs, InP, and CdTe. All these semiconductors have closely spaced two or three energy valleys in the conduction band (Fig 2.8a, b). At DC voltage and hence low electric field (E_f) in the material, most of the electron will be located in the *lower valley* (Fig. 2.8b). At higher E_f beyond E_{th} most of the electrons will be transferred to the high-energy *upper valley*, where the effective electron mass, m_2 is much larger and hence the mobility (μ_2) and velocity (v) are much low than that at lowest valley (μ_1). As the conductivity is proportional to mobility, the conductivity and hence current decreases with higher Electric field, E_f ($E_f > E_{th}$) or voltage ($V > V_{th}$). This is called *transferred electron effect*, and the device is also called Transferred Electron Device (TED) or Gunn diode. The bulk material (without any junction) behaves as a –ve resistance device over a range of applied voltage and therefore used as microwave oscillators.

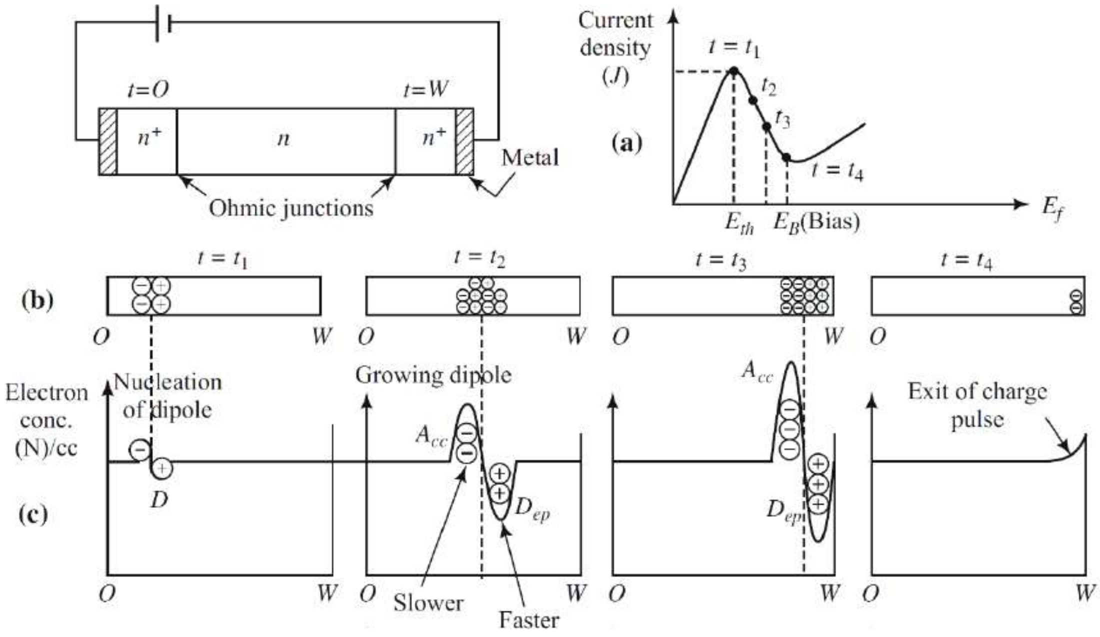


Figure 2.9: (a) Current density (b) Moving e-h dipole in the device (c) Electron concentration

Initially at t=t_1 (see Fig 2.9) when a bias voltage is applied such that the V>V_{th} to the Gunn diode due to high field domain is created near cathode. Therefore, increase of E_f in one region will lead to decrease in rest of the region.

At t=t_2 electrons from *lower valley* gets transferred to *Upper valley,* so their velocity drops. The electrons to the right of the domain move out to the anode faster, causing deficiency (Region D_{ep}) of electron (i.e. makes it +ve charged). To the left of the domain, slow-moving *upper valley* electrons get accumulated (region A_{cc}) and thus form the '*dipole charge region'* around the high-field domain.

This space charge dipole as well as the high field of the domain keeps growing (at t=t_3) while moving right (towards anode) and exits out (at t=t_4) of the anode as a current–voltage pulse. Immediately after this, the electric field again grows to a uniform value and the domain formation restarts at the cathode end, i.e. left end

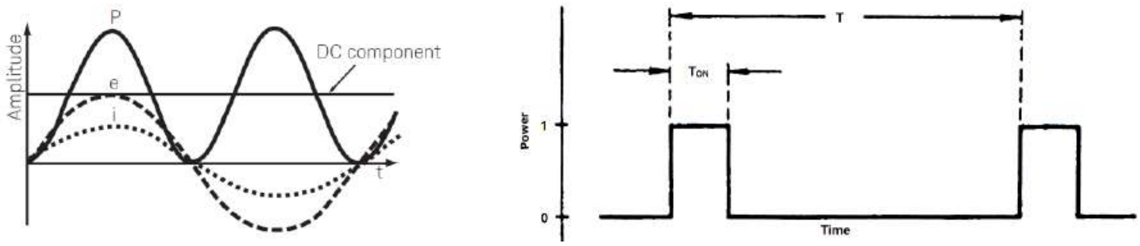
2.3.2 Power in RF:

The power is defined as the time rate of transforming energy. In case of microwave, this energy is used in many different forms: exchange of information over long distance, heating a microwave oven or acceleration of particles in nuclear engineering etc.

For low frequency signals, power measurements are done from the voltage, current or lumped values of circuit parameters. For microwave frequencies, the difficulty arises due to

1. Distributed nature of the circuit elements
2. Reflection of the signal, wherever there is an impedance mismatch.

Average power in an alternating current circuit: $P_{avg} = \frac{1}{T}\int_0^T E\ I\ dt$



(a) (b)

Figure 2.10: (a) Average Power (b) Pulsed power

Pulsed power: Peak power of the pulse is related to the average power of the pulse by the duty cycle of the pulse. Thus:

Peak RF Power, $P_{Peak} = \frac{P_{avg}}{D}$;

Here,

Duty cycle, $D = \frac{T_{ON}}{T}$; Pulse Repetitive Time period (PRT)= $T = t_{on} + t_{off}$

Usually, average power is involved in the microwave circuit which has a continuous signal source. But, where pulsed signals serve as a signal source, peak power is more meaningful way of expressing power.

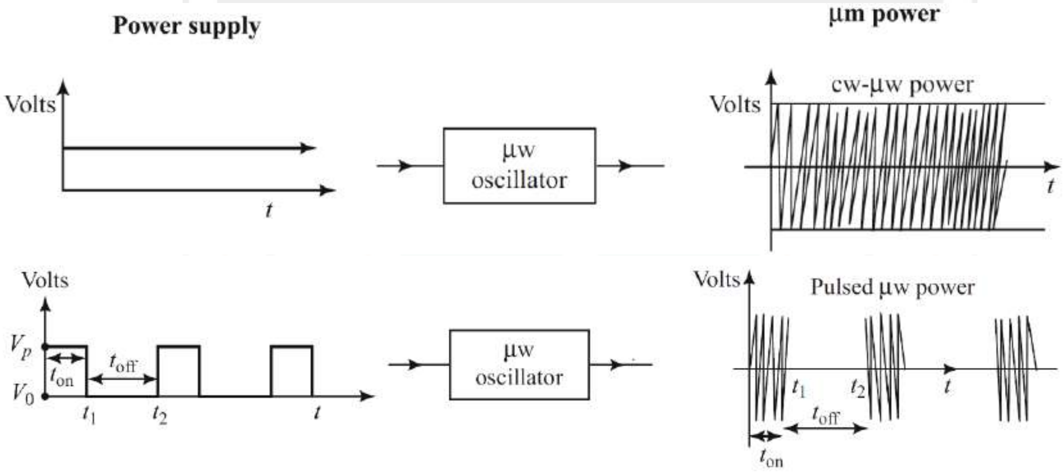


Fig 2.11: CW and pulsed power supply and the corresponding microwave power output

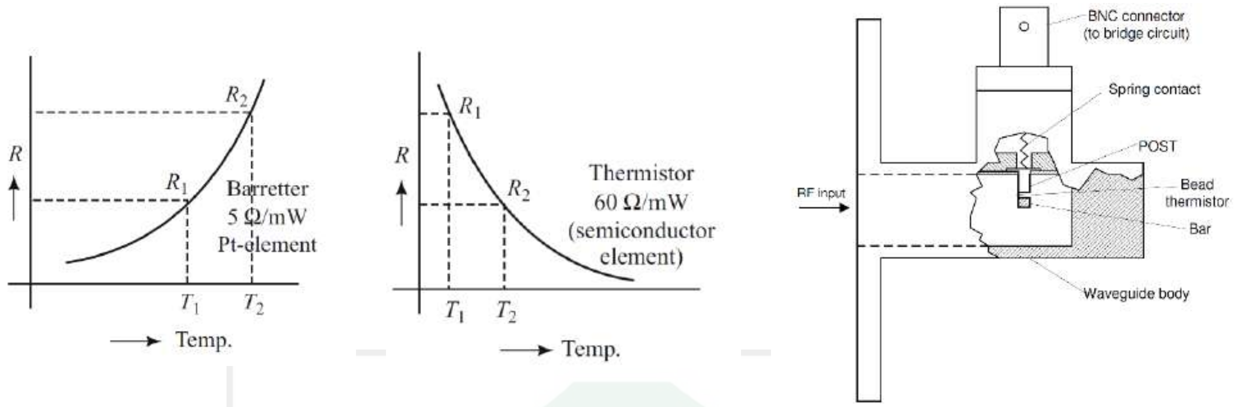
2.3.3 Microwave Power Sensor

(i) Bolometer

Bolometer is a device which is used for measuring the power of incident electromagnetic radiation via the heating of a material with a temperature-dependent electrical resistance (thermistor in our case). Bolometers can be thermistor mount or waveguide mount. Fig 2.12(c) shows a waveguide mount bolometer. Typically, a thermistor bead has a diameter of around 0.05–0.5 mm with a small-size (diameter of 15–100 µm) metal wire embedded inside. The bolometer mount must be designed to satisfy the following requirements:

1. Present a good impedance match to the transmission line over frequency of interest.
2. Keep i^2R and dielectric losses within the structure minimized so that power is not dissipated in electrical contacts, waveguide walls or insulators.
3. Provide isolation from thermal and physical shocks

4. Keep leakage small so that microwave power does not escape from the mount in a shunt path around the bolometer.



(a) (b) (c)

Fig 2.12: (a) Barretter (b) Thermistor (c) Bolometer in a waveguide

This technique is based on devices such as detectors, bolometers and thermocouples (whose resistance changes with applied power). A bolometer is a square law device that produces current that is proportional to applied power. Bolometer can measure power in the range of .1 to 10mW and are very sensitive.

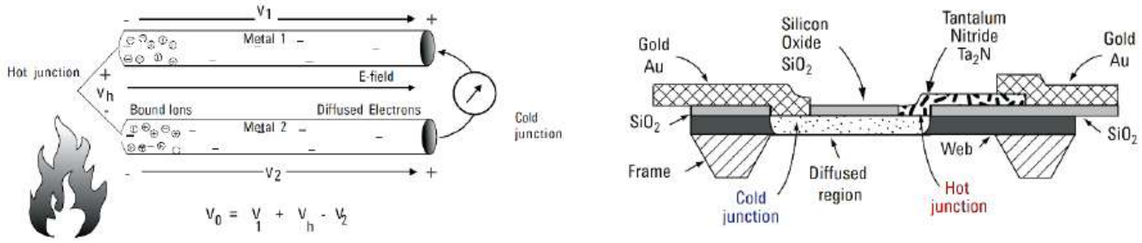
There are two different types of bolometers (Fig 2.12a, b): Barretter (a wire mounted in cartridge like ordinary fuse & its resistance increases with temperature) and thermistor (a small semiconductor bead with connecting wire & its resistance decreases with temperature).

(ii) Thermocouple

Thermocouples are based on the fact that dissimilar metals generate a voltage due to temperature differences at a hot and a cold junction of the two metals. A high-quality thermocouple (high frequency material for thermo-junctions with low error rate) can convert the microwave energy to a readily measurable DC voltage which is generated as the wires conduct the high frequency current and heat is generated across the two different metal contacts. The meter indication is calibrated to represent the power level in the waveguide.

A thermocouple is usually a loop or circuit of two different materials as shown in Fig 2.13(b). One junction of the metals is exposed to heat, the other is not (Fig 2.13a). If the loop remains closed, current will flow in the loop as long as the two junctions remain at different temperatures. If the loop is broken to insert a sensitive volt-meter, it will measure the net e.m.f. In contrast, a power meter works when voltage is applied to a meter and meter is calibrated to indicate the power.

Modern RF thermocouple use Silicon Oxide and Tantalum Nitride. From the Fig 2.13(b) it can be seen that the gold-diffused region create the cold junction and tantalum nitride- diffused region create the hot junction of the thermocouple. This diffused region is heavily doped by diffusing impurities into it.



(a) (b)

Figure 2.13: (a) Hot and cold junction (b) Cross section of a thermocouple

Since thermocouples are heat-based sensors, they are true "averaging detectors." So they are used in (CW) to complex digital phase modulations. They are more rugged than thermistors, exhibit higher sensitivity and also the DC voltage out is proportional to the input RF power. They are typically used to measure very low power in the range of .01-1mW.

2.3.4. Power measurement by bolometric (thermistor) using Wheatstone bridge [set- up1]:

One of the simplest methods for bolometric power measurement is to place a thermistor bolometer in one leg of Wheatstone bridge (Fig. 2.14). The bridge is excited by a regulated DC supply whose amplitude may be adjusted with R_1. Since R_4 is a thermistor, its resistance may be controlled by the amount of current allowed to pass through it.

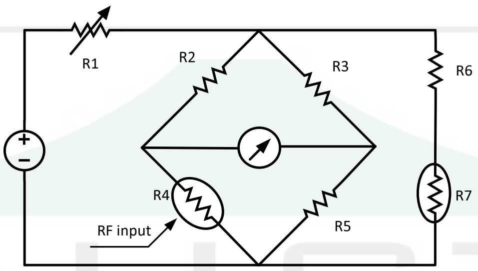


Fig 2.14: Bridge assembly using thermistor

In operation, R_1 is adjusted until just enough current passes through the bridge to make the thermistor resistance equal to R_5, bringing the bridge into balance and causing the meter to read zero.

In this case, half of the total current I_T will pass through R_4.

$$I_4 = \frac{1}{2} I_T$$

$$\text{Power, } P_T = I_4^2\, R_4 = \frac{1}{4} I_T^2\, R_4$$

Microwave power is then applied to the thermistor and heating effect causes the thermistor resistance to decrease, unbalancing the bridge in proportion to the power applied. The unbalance current is indicated on the meter, which is calibrated directly in mW. If we adjust R1 to balance the bridge again, the total current changes. This dc power change is equal to the RF power applied.

DC power, $P_{DC} = \frac{1}{2} I_{DC}^2\, R_4$

RF power, $P_{RF} = P_T - P_{DC} = \frac{1}{4}(I_T^2 - I_{DC}^2)\, R_4$

$$= \frac{1}{4}(I_T - I_{DC})(I_T + I_{DC})\, R_4$$

$$= \frac{1}{4}(\Delta I)(I_T + I_{DC})\, R_4$$

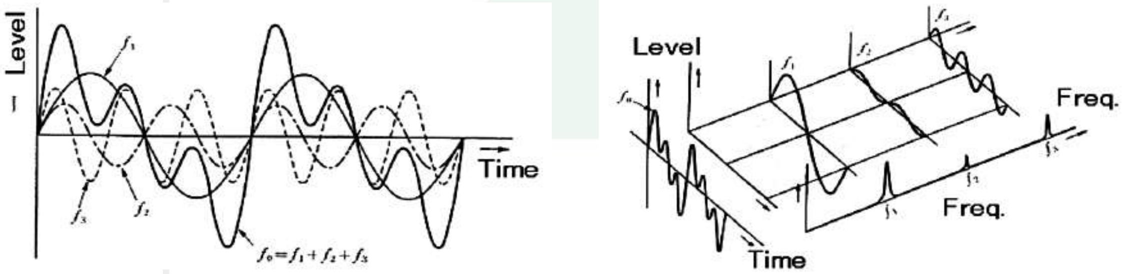
$$\propto \Delta I$$

This method is acceptable for power measurements when RF power is 1 mW (0 dBm) or above. Otherwise, ΔI becomes difficult to deal with.

2.3.5.1 Spectrum Analyzer

We use an oscilloscope to view the instantaneous value of a particular electrical event (or some other event converted to volts through an appropriate transducer) as a function of time. In other words, we use the oscilloscope to view the waveform of a signal in the time domain. Fourier theory tells us any time-domain electrical phenomenon is made up of one or more sine waves of appropriate frequency, amplitude, and phase. In other words, we can transform a time-domain signal into its frequency-domain equivalent. Measurements in the frequency domain tell us how much energy is present at each particular frequency. To transformation from the time domain to the frequency domain, the signal must be evaluated over all time, that is, over infinity. However, in practice, we always use a finite time period when making a measurement

For example the f_0 wave in Fig 2.15(a) looks complicated in time domain but in frequency domain (fig 2.15b) it can be easily seen it consists of 3 different waveform with frequency f_1, f_2, f_3.



(a) (b)

Fig. 2.15: (a) Time Domain (display in oscilloscope) (b) Frequency Domain

2.3.5.2 Spectrum Analyzer (SA) architecture

In Fig. 2.16 block diagram, we see that an input signal passes through an attenuator (ensure the signal enters the mixer at the optimum level to prevent overload, gain compression and distortion), then through a low-pass filter (This filtering prevents out-of-band signals from mixing with the local oscillator and creating unwanted responses on the display) to a mixer, where it mixes with a signal from the local oscillator (LO).

Because the mixer is a non-linear device, its output includes not only the two original signals, but also their harmonics and the sums and differences of the original frequencies and their harmonics. If any of the mixed signals falls within the pass band of the intermediate-frequency (IF) filter, it is further processed (amplified and perhaps compressed on a logarithmic scale). It is essentially rectified by the envelope detector, filtered through the low-pass filter and displayed. A sweep generator creates the horizontal movement across the display from left to right. The ramp also tunes the LO so its frequency change is in proportion to the ramp voltage. The output of a spectrum analyzer is an X-Y trace on a display.

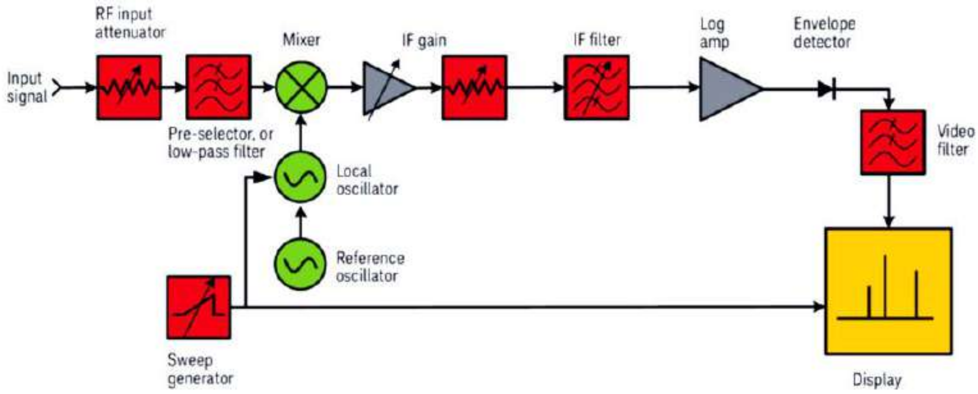


Fig. 2.16: Block diagram of a traditional spectrum analyzer that uses the heterodyne principle

The Modern spectrum analyzers (Fig 2.17) digitize incoming signals much earlier in the signal path compared to spectrum analyzer designs of just a few years ago. The change has been most dramatic in the IF section of the spectrum analyzer. Digital IFs have had a great impact on spectrum analyzer performance, with significant improvements in speed, accuracy and the ability to measure complex signals using advanced DSP techniques.

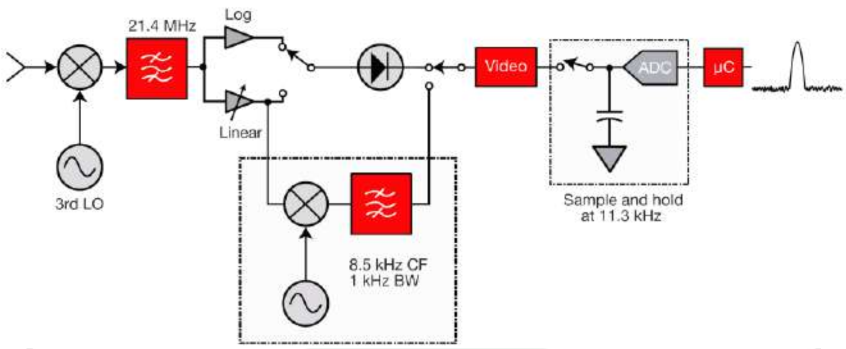


Fig. 2.17: Block diagram of a Modern DSP based spectrum analyzer

The benefits of using a digital IF based analyzer include the use of high measurement speeds at low RBWs and the ability to record the signal in the time domain with all of the phase information. This makes it possible to analyze complex modulations which the heterodyne based SA cannot do. Modern Spectrum analyzer use both heterodyne and digital IF based architecture to create the SA. It is therefore sometimes known as signal analyzer rather then Spectrum analyzer.

2.3.5.3 Some Important parameter related to SA

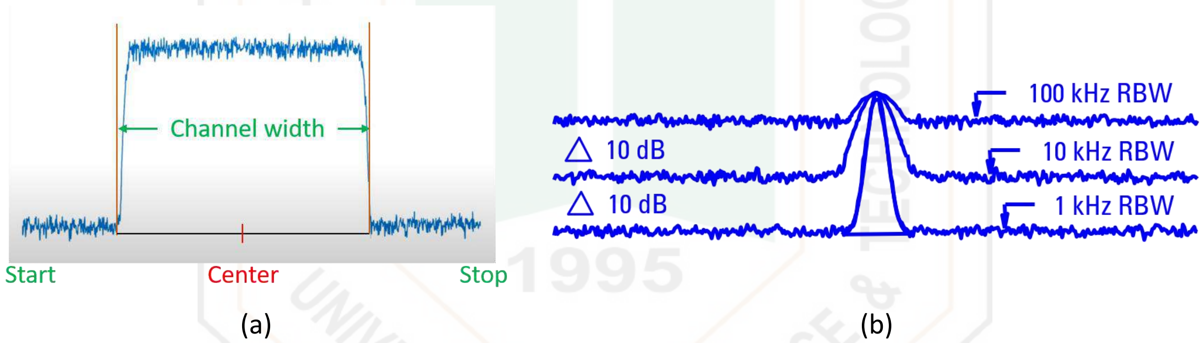


Fig. 2.18:(a) Typical spectrum parameter (b) RBW and DANL

A typical spectrum in spectrum analyzer can be represented by inputting the start and stop frequency of concern. This will make the Ramp generator in the SA to sweep from the 'start' frequency to the 'stop' frequency. The same can be done by entering the 'center frequency' and 'span' in the SA interface.

The *Resolution Bandwidth Filter (RBW)* filter is the bandpass filter in the IF path. Adjusting the bandwidth of this filter allows for the discrimination of signals with closely spaced frequency components while also changing the measured noise floor. Reducing the RBW will lower the noise floor of the analyzer and make the signal clearer but it will also reduce the frequency sweep speed. Fig 2.18 Show the signal in three different RBW level.

The *video bandwidth filter (VBW)* filter is the low-pass filter directly after the envelope detector. VBW is used to average or smooth the display trace. VBW effect the display but not the way the signal is measured

The Displayed Average Noise Level (DANL) is the average noise level displayed on the analyzer. Changing the RBW from 100 kHz (RBW_{old}) to 10 kHz (RBW_{new}) results in a change of noise level = - 10 dB

2.4 Experiment procedure:

2.4.1 GUNN OSCILLATOR - THERMOCOUPLE SETUP

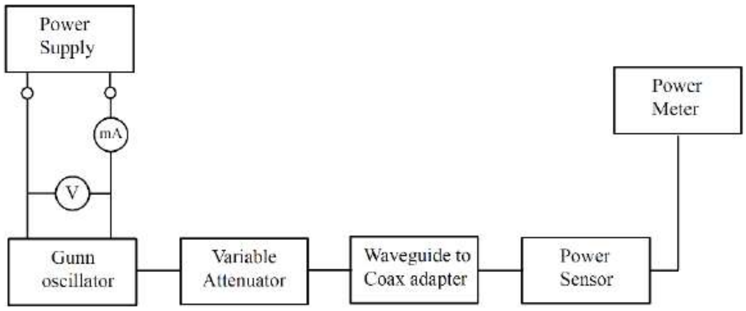


Figure 2.19: Setup for measuring the current vs. voltage characteristic of the Gunn diode

(i) I-V characteristics of Gunn oscillator

1. Set up the equipment as shown in Fig. 2.19
2. Set up the voltage to 4 V. Set the variable attenuator to 10 db. This will ensure proper isolation to Gunn oscillator.
3. Raise the voltage in 0.5 V increment. Record the current output to the table.

Voltage (V)	4.5	5	5.5	6	6.5	7	7.5	8	8.5	9	9.5	10	4	3.5	3	2.5	2	1.5
Current (mA)																		

4. Reduce the voltage to 0 V.
5. Plot the V-I characteristics curve from the measured data

(ii)Oscillator output power vs. supply voltage

1. Tune the Gunn oscillator so the oscillator output frequency is in X band (8-12GHz).
2. Set the power meter so sense X-band and the unit is “absolute” and in “mW”.
3. Turn the power meter on and calibrate the power meter to zero.
4. Raise the gun diode voltage in 0.5 V increment and record the power indication on the power meter and the attenuator setting.
5. Set gunn diode voltage to 4V.
6. Convert the obtained power reading in milliwatts to dBm. Then add the attenuation (in dB) to the dBm. Then reconvert this Gunn diode output power back to mw.

   Example:

   Assume supply voltage = 8.5 V, power reading = 6.3 mW

   Converted power reading in dBm = $10\log\left(\frac{P_{(mW)}}{1mw}\right)$ =$10\log\left(\frac{6.3}{1}\right)$ = 7.99 dBm

   Add an attenuation (let 3dB)

   Measure the power after attenuation in dBm (Let’s say 3.99dBm)

   Converted power reading in mW =$1mW * 10^{\left(\frac{P_{(dBm)}}{10}\right)} = 1 * 10^{\left(\frac{3.99}{10}\right)}$= 2.51 mW
7. Repeat the step 5 for Gunn diode voltage of 4.5 and 5 and complete the table.

Gunn oscillator supply Voltage	Power meter reading (mW)	Converted power (dBm)	Attenuator setting (dB)	Gunn diode output (dBm)	Gunn diode output (mW)
4					
4.5					
5					

8. Draw a graph showing the relationships between the supply voltage and the output power.

2.4.2. WIRELESS CHANNEL POWER MEASUREMENT, SPECTRUM HUNTING AND MONITORING USING SPECTRUM ANALYZER

Channel is the range of frequency in which most of the power is concentrated. Channel is Defined by Start and Stop frequency (a and b in Fig. 2.20a).

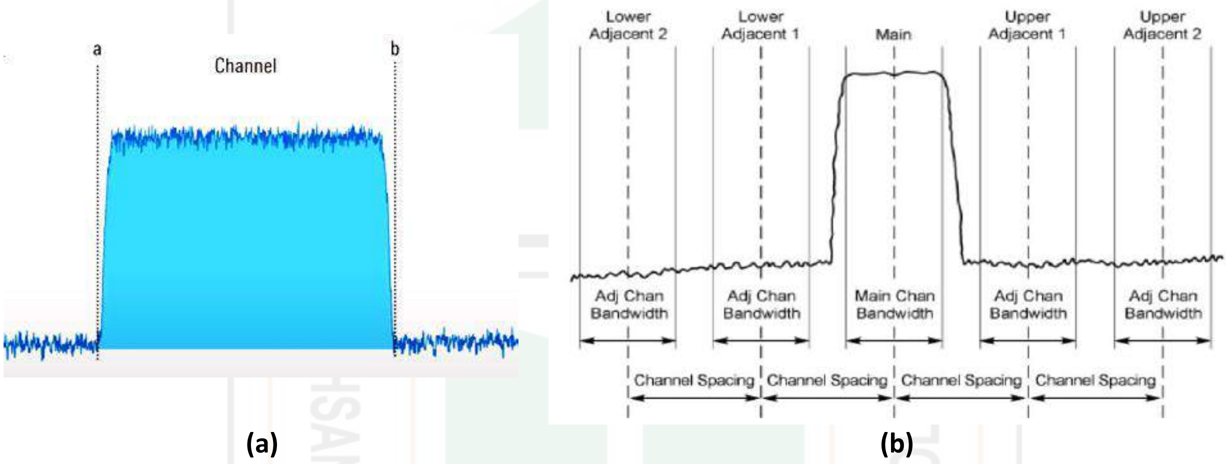


Fig. 2.20: (a)RF channel (b) Adjacent Channel, channel spacing and Bandwidth

Average power is the square root of the sum of the squares of the voltage samples over the measurement time.

Channel Power is defined as the total average RF power in the selected channel (frequency band). Spectrum analyzer can measure power by integrating all the power in between a range of frequency using the following formula:

$$Power\ of\ the\ channel = \int_a^b Power$$

Adjacent Channel Power Ratio (ACPR) is ratio between the total power of adjacent channel (intermodulation signal) to the main channel's power (useful signal). ACPR measure of the signal power leaking from the main channel into adjacent channels. So

$$ACPR_{dBc} = 10 \log \frac{P_{adj}}{P_{ch}}$$

Peak/Avg. Ratio is the peak power in the transmitted signal to the average power in the transmitted signal (located in the CCDF display).

2.4.3 Experimental procedure in SignalVu software: Measuring Channel power and Adjacent Channel Power Ratio (ACPR) of FM radio channel

1. Connect the Monopole antenna in the N type connector of the RSA 306B SA.
2. After connecting the Spectrum analyzer in USB 3.0 port Open SignalVU software.

3. Click **Preset** to set the instrument to its default state.
4. Click the **Displays** button.
5. Select **RF Measurements** from the **Measurements** box (Fig 2.21).
6. Double-click **Chan Pwr and ACPR** in the 'Available displays box' to add the option. Click 'OK' to complete selection.

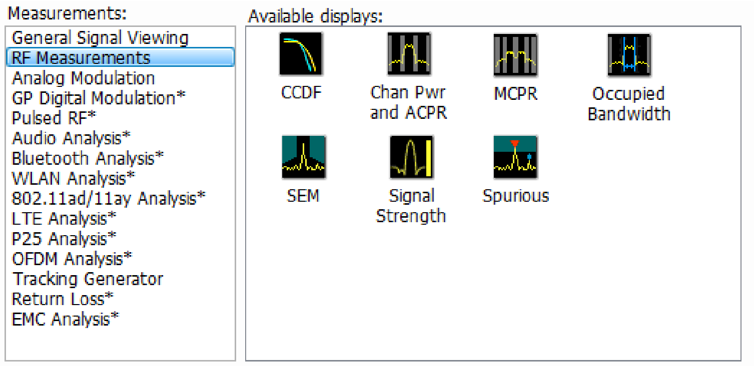


Figure 2.21: RF measurement options

7. Press the front-panel **Freq** button and adjust the frequency to that of your main channel.
8. Press the **Settings** button. This displays the control panel for Chan Power and ACPR (the tab displayed will be the tab displayed the last time the Settings panel was opened).

Figure 2.22: Channels Tab

9. To set the number of adjacent channels, select the **Channels** tab (Fig 2.22). Enter the number of channels in the **Number of adjacent pairs** value box. If zero is entered for the number of adjacent pairs, the resultant measurement will be **channel power only**.
10. To set the spacing between channel centers, enter the required value in the **Channel Spacing** value box. (*see Appendix B1 and Fig 2.23 for different parameter*)

1. FM Radio Broadcasting:

The license of FM radio broadcasting is given from the Ministry of Information. According to NFAP, frequency is assigned from BTRC to the licensee from 87.5 to 108 MHz.

Till 2013, licenses have been given to twelve organizations including state owned 'Bangladesh Betar' for FM radio broadcasting. Each organization has been assigned 200 kHz of spectrum.

Figure 2.23: FM Broadcasting standard (from BTRC.gov.bd)

11. To set the channel bandwidth, enter the required value in the **Channel Bandwidth** value box.
12. After you have configured the channel settings, click the close button in the Settings panel or press the **Settings** button again to remove the settings panel.
13. Press **Replay** to take measurements on the recalled acquisition data.

(e) Report:

1. Draw respective block diagrams of experimental set-ups and briefly explain the function of each part.
2. Discuss the results.
3. Briefly explain different microwave signal generation techniques.
4. Why spectrum monitoring is used?

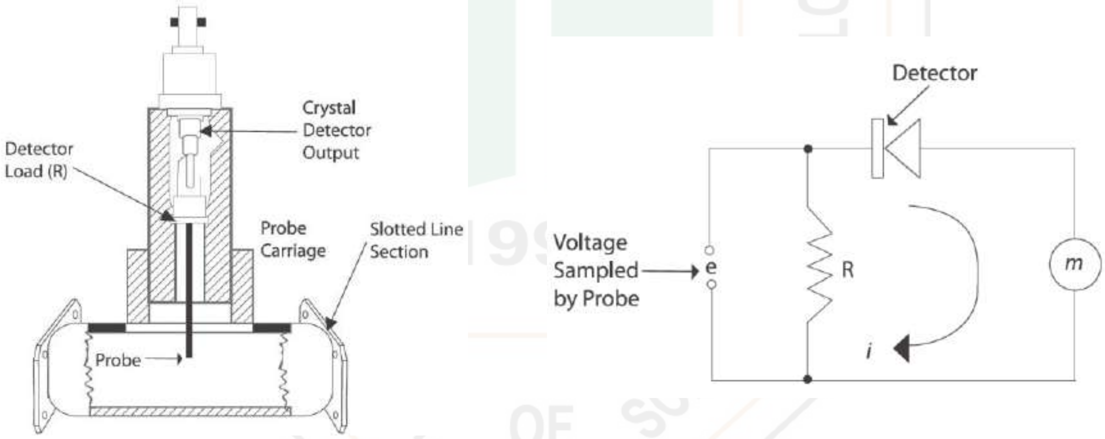
Experiment no: 3
Name of the Experiment: Measurement of Standing Wave Ratio, frequency & Wavelength

(a) Objective:

1. To learn how to determine SWR using slotted line or SWR indicator.
2. To learn how measure frequency and wavelength of microwave signal.

(b) Equipment & Waveguide component list:

1. Gunn and Klystron power supply
2. Gunn and Klystron oscillator
3. **PIN-diode modulator:** In old setup external module and in new setup embedded with Gunn oscillator module.
4. Variable attenuator
5. **Crystal Detector:** The crystal detector is basically a diode assembly which responds to the electromagnetic field inside the waveguide. The diode assembly consists of a small thin piece of silicon, a thin tungsten wire and a case. One side of the silicon is directly connected to the case and the other side is connected to the tip of tungsten wire (Fig 3.1a).
   The diode action is due to the different properties of silicon and tungsten. Silicon has few surplus electrons but there are many free electrons in tungsten. Therefore, when a voltage is applied across diode in such a direction to force electrons to leave silicon and enter tungsten, a very small current result in. When direction of the voltage is reversed, a large current flows from tungsten into silicon. This is how the diode can be used for detection of microwave energy. For such diodes, output voltage/ current is proportional to square of input voltage (square law characteristics)



(a) (b)

Figure 3.1: Slotted Transmission Line and Probe (a) Cutaway Sketch of a Slotted Section with Probe, Carriage, and Detector (b) Equivalent Detector Circuit

6. **Slotted line:** In measuring the standing waves inside a waveguide, a slotted line is used to probe the amplitude and the phase of the standing wave pattern. As the name implicates, a slotted line has slot along the center line along the long side. An assembly consisting of a probe and a crystal detector, is designed to slide along the slot and as it does, the probe samples the field in the waveguide, while the crystal detector provides a rectified signal.

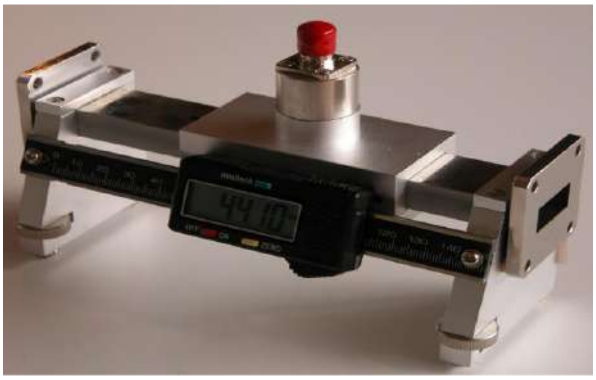


Figure 3.2: Slotted Line and Crystal detector

7. **Slide screw tuner:** The primary use of the slide screw tuner is to match loads, detectors, or antennas to the characteristic impedance of the waveguide. It consists of a probe mounted on a carriage which slides along a narrow and long on the feeding waveguide. When the adjusting micrometer is turned, depth of the probe varies. The depth and the position of the probe causes reflection in the waveguide at a specific amplitude and phase.

Micrometer's Scale (mm)	3	5	7	9
Probe's Depth (mm)	7	5	3	1

8. **Matched termination:** The matched terminator is essentially a matched to the microwave transmission line. As the standing waves occur due to impedance mismatches in the system, the matched termination is used to minimize the SWR in a system.

(a) (b)

Figure 3.3: (a)Slide screw tuner (b) Matched termination

9. **SWR Meter:** The SWR meter has a tuned 1 KHz input BNC to read the SWR. It is basically a calibrated voltmeter (1000mv in our case) with additional amplifier which can be used to increase the input signal. The meter has a voltage, dB and SWR scale (Fig. 3.4). In some SWR meters there is also an expanded SWR scale which is used to measure very low SWR.

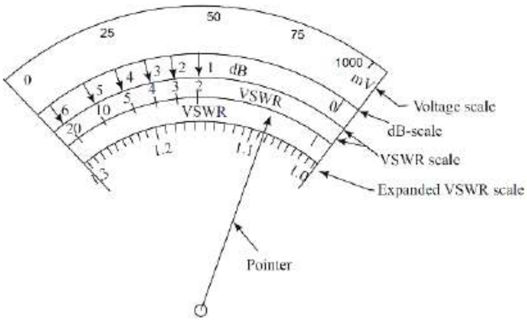


Figure 3.4: VSWR meter display

c) Theory:

1. Standing wave ratio:

At any point along a transmission line, electromagnetic field is the sum of two waveforms: one travelling towards the load (transmitted wave) and another towards generator (reflected wave). It occurs because of impedance mismatch between left and right side of the observation point of transmission line. Any open spot on the line is responsible for another impedance mismatch and cause reflection. The amplitude and phase of the reflected wave depend on the load mismatch. The degree of attenuation of the line affects amplitude of the reflected wave also. The only way reflection can be eliminated is either the line is infinitely long or there is impedance match.

A standing wave results from two travelling waves in opposite direction. The vector sum of two waves creates minimum and maximum points on standing wave pattern in a loss less transmission line. (See *Appendix C2* for standing wave pattern for various case)

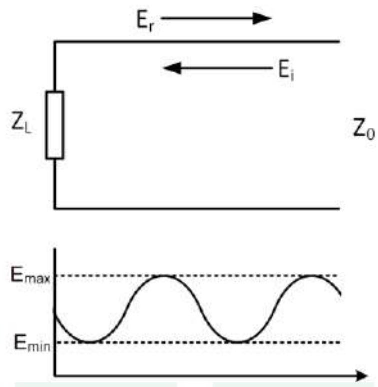


Figure 3.5: Incident and Reflected wave in a circuit

According to transmission line of Fig. 3.5, Reflection coefficient,

$$\Gamma = \frac{E_r}{E_i} = \frac{Z - Z_0}{Z + Z_0}$$

For reflection coefficient evaluated at the complex load (Z_L),

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0}$$

$$VSWR = \frac{|E_{max}|}{|E_{min}|} = \frac{1 + |\Gamma|}{1 - |\Gamma|} \quad (3.1)$$

When RF signal is transmitted down a line into a load, even one with a good match, some of the signal reflects back toward the source. The amount of signal sent vs. the amount of reflected back is compared and referred to as the standing wave ratio (SWR). When the pattern deals with measurements of voltage, it is called the voltage standing wave ration (VSWR)-Fig. 3.4

2. Current or voltage measurement method:

This diagram shows us a sample length of transmission line, in our case X-band wave- guide. In Figure 3.6, we see an expanded view of the RF sine wave rising and falling along the length of the line. Placing a slotted line into the transmission line from point A to point B allows us to inspect a probe into the slot in the line. We can then move the probe along the line reading peaks and dips, e_{max} and e_{min} which are shown as X and Y. Because RF voltages are so difficult to measure, we will make the probe like the one in Fig. 3.6.

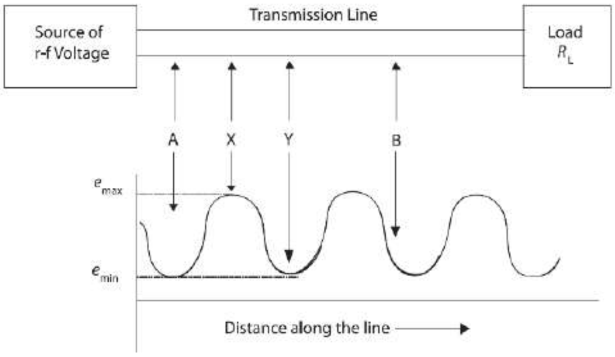


Figure 3.6: Voltage Distribution along a Transmission Line

The crystal detector which change the RF voltage into a DC current and will be read by a current meter. Here we used a "square law" detector. These detectors have an output proportional to the square of the input. So if we place the probe at Y on the slotted line (Fig. 3.6), we will get

$$i_{min} = ke_{min}^2$$

$$\Rightarrow e_{min}^2 = \sqrt{i_{min}/k}$$

Then if we move the probe to point X (Fig. 3.6), we get

$$i_{max} = ke_{max}^2$$

$$\Rightarrow e_{max}^2 = \sqrt{i_{max}/k}$$

Together we get:

$$VSWR = \frac{e_{max}}{e_{min}} = \sqrt{\frac{i_{max}/k}{i_{min}/k}}$$

$$VSWR = \sqrt{\frac{i_{max}}{i_{min}}} \tag{3.2}$$

Here, e_{max} and e_{min} are the input of square law detector and i_{max} and i_{min} are the corresponding outputs

This way the measurement of VSWR is by far the easiest, but if minimum reading cannot be measured with reliable degree of accuracy, the inherent flaws in the detector may create distortion increasing the chance of error.

3. Double Minimum Method

When standing wave ratios are larger than ten to one, the 'double minimum' measurement system can be used to increase the accuracy and reduce the percentage error (Fig. 3.7).

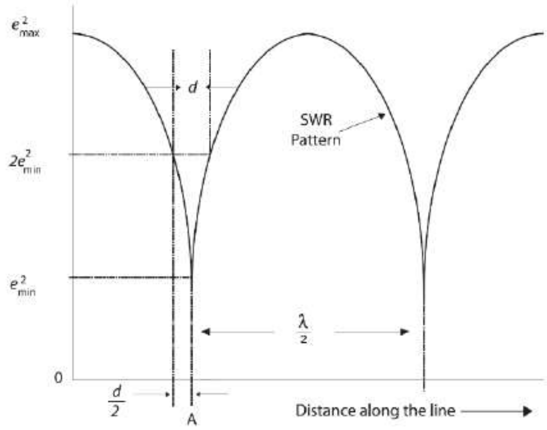


Figure 3.7: Double Minimum Method

The mathematical expression of the curve is,

$$e = e_{min}^2 + \left(e_{max}^2 - e_{min}^2\right)\sin^2\left(\frac{2\pi l}{\lambda}\right)$$

Starting at point 'A', we move to the right ($l = \frac{d}{2}$) until the reading is twice ($2e_{min}^2$) what it was at point 'A'.

$$2e_{min}^2 = e_{min}^2 + e_{max}^2 \sin^2\left(\frac{\pi d}{\lambda}\right) - e_{min}^2 \sin^2\left(\frac{\pi d}{\lambda}\right)$$

$$\Rightarrow e_{min}^2\left[1 + \sin^2\left(\frac{\pi d}{\lambda}\right)\right] = e_{max}^2 \sin^2\left(\frac{\pi d}{\lambda}\right)$$

$$\Rightarrow \frac{e_{max}^2}{e_{min}^2} = \frac{1 + \sin^2\left(\frac{\pi d}{\lambda}\right)}{\sin^2\left(\frac{\pi d}{\lambda}\right)}$$

$$\Rightarrow \frac{e_{max}^2}{e_{min}^2} = \frac{2 - \cos^2\left(\frac{\pi d}{\lambda}\right)}{\sin^2\left(\frac{\pi d}{\lambda}\right)}$$

$$VSWR = \frac{e_{max}}{e_{min}} = \frac{\sqrt{2 - \cos^2\left(\frac{\pi d}{\lambda}\right)}}{\sin\left(\frac{\pi d}{\lambda}\right)}$$

VSWR's with ten or greater angle $\frac{\pi d}{\lambda}$ will be small, So $1 + \sin^2\left(\frac{\pi d}{\lambda}\right) \approx 1$

$$VSWR \approx \frac{1}{\sin\left(\frac{\pi d}{\lambda}\right)}$$

Because of the small angle, $\sin\left(\frac{\pi d}{\lambda}\right) \approx \frac{\pi d}{\lambda}$

$$VSWR \approx \frac{\lambda}{\pi d} \qquad (3.3)$$

4. Frequency measurement using slotted line:

The relationship of a RF wave could be seen as an equation between frequency, wavelength and velocity:

$$V = f\lambda$$

Frequency remains fixed, so that wavelength is directly affected by the velocity. The velocity of our wave is determined by its physical path of travel if the wave is traveling through air, velocity is equal to:

$V = V_0 = 3 \times 10^{10} cm/sec$

When it travels along paths other than air, the velocity is affected by other factors, as expressed:

$V = \frac{V_0}{\sqrt{\varepsilon_r \mu_r}}$; Here μ_r and ε_r are the permeability and permittivity of the path.

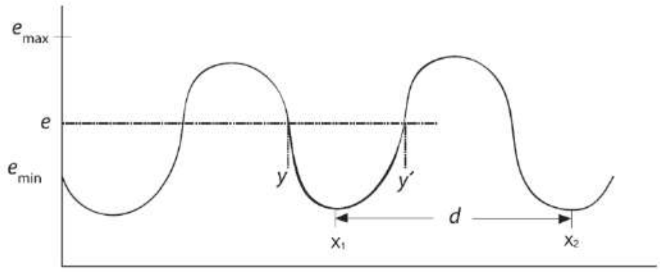


Figure 3.8: Frequency and wavelength measurement

The distance d between X_1 and X_2 (Fig. 3.8) is represented by the equation $\lambda_L = 2d$. λ_L is equal to the wavelength as measured along the slotted line. When using a coaxial line, velocity V_L will be about equal to free space velocity V_0, so we see that $\lambda_L = \lambda_0$.

When we use waveguide, velocity is determined by the larger dimension 'a' of waveguide. We can compare the wavelength in waveguide to the wavelength of "air" by

$$\frac{1}{\lambda_0^2} = \frac{1}{\lambda_L^2} + \frac{1}{(2a)^2}$$ (For dominant TE_{10} mode cutoff wavelength, $\lambda_C = 2a$)

$$\Rightarrow \frac{1}{\lambda_0} = \sqrt{\frac{\lambda_L^2 + 4a^2}{4a^2\lambda_L^2}} \tag{3.4}$$

$V_0 = f\lambda_0$, we calculate

$$f = V_0 \frac{\sqrt{\lambda_L^2 + 4a^2}}{2a\lambda_L} \tag{3.5}$$

d) Experiment procedure

d1) Gunn oscillator setup

1. Set up equipment as shown in Fig. 3.9.

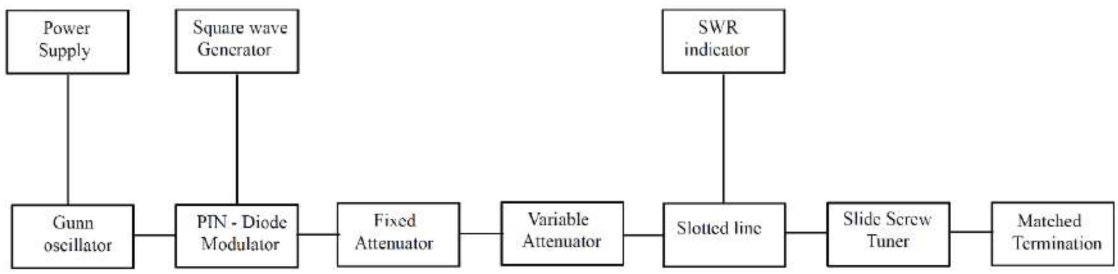


Figure 3.9: Setup diagram for SWR measurement

2. Turn on the Gunn oscillator. Push the modulation button.
3. Set the variable attenuator to 10dB.

d 1.1) Measuring low and medium range SWR (direct method)

1. Completely disengage the probe of slide screw tuner (VSWR reading should be less than 1.3)
2. Set the gain in SWR meter to 0dB and fine dial to fully CW direction
3. Move the probe of the slotted line and observe SWR indicator meter deflection. If there is no deflection increase the gain of the meter.
4. Increase the gain dial until the there is a deflection and the deflection can be adjusted to SWR =1 using the fine dial.
5. Move the probe in slotted line until a maximum deflection is observed in SWR scale. Adjust the fine dial to set the maximum deflection as SWR=1.
6. Move the probe, when there is a minimum deflection take the SWR reading and put it on the table (this is the VSWR).
7. Repeat the procedure for three different probe depths.

Probe depth (mm)			
VSWR			

d 1.2) Measuring high SWR (double Minimum method)

1. Maximize the depth of the probe of the slide screw tuner. Large depth of the probe is required for high SWR measurements.
2. Move the probe from left to right along the slotted line until a maximum deflection is observed on the SWR scale.
3. Adjust the gain of the meter until 0dB is shown on the dB scale. If required, reduce the attenuation. Record the position reading as d_0. (see Figure 3.10)

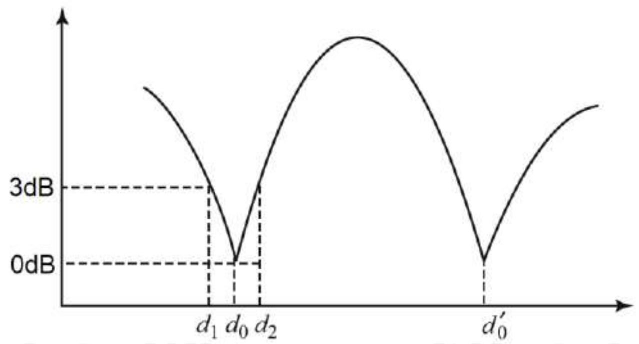


Fig 3.10: Wave shape for Experimental data

4. Move the probe to the left along the slotted line until 3 dB is obtained on dB scale.
5. Record the position of the probe under the d_1 column in table.
6. Move the probe to the right along the slotted line until 3 dB is obtained on dB scale.
7. Record the position of the probe under the d_2 column in table.
8. Move the probe to the right until a 0dB value is observed again. Record the length as d'_0.
9. Repeat the measurement at three different probe depths.

Probe Depth (mm)	d_0 (mm)	d_1 (mm)	d_2 (mm)	d'_0 (mm)	$\lambda = 2(d_0' - d_0)$ (mm)	$SWR = \dfrac{\lambda}{\pi(d_2 - d_1)}$

d2) Klystron oscillator setup (Using current Measurement/square law detector)

1. Set up test fixture in Figure 3.11
2. Connect one of the BNC cables from the detector on the slotted line to the input marked VSWR on the power supply.
3. Turn on power supply and apply RF power.
4. Move probe on slotted line and adjust for maximum deflection (try to keep this at the center of the slotted line if possible).
5. With the shorted attenuator set at maximum, adjust the RF feed attenuator for a close to full scale reading.
6. Move the probe for a minimum reading.

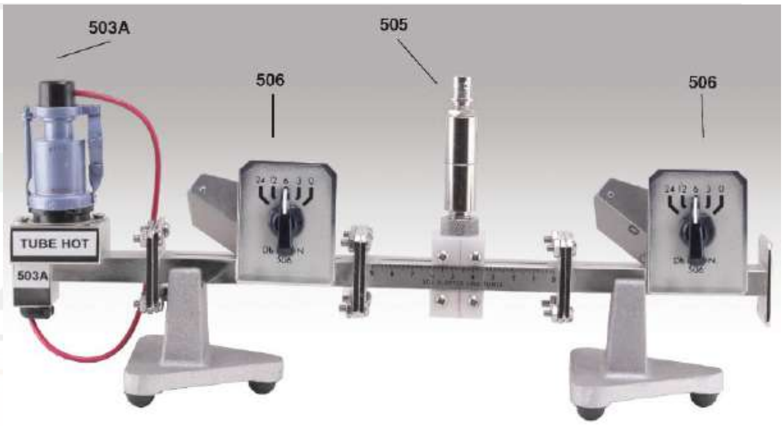


Figure 3.11: Klystron Oscillator setup

7. Adjust the short feed attenuator for a reading of about 1/4 scale.
8. Measure and record the readings at the points indicated on Table.

Position (cm)									
Current (μA)									

9. Using equation 3.2 Determine the VSWR for the system

(e)Report:

1. Draw respective block diagrams of experimental set-ups and briefly explain the function of each part.
2. Briefly explain mathematical background of VSWR, frequency and wavelength measurement.
3. Comment on the results.
4. Explain (a) Reflection coefficient, (b) Transmission coefficient, (c) Standing wave Ratio with respect to a microwave signal transmission line.

Experiment no: 4

Name of the Experiment: **Measurement of unknown Impedance using SMITH chart**

a) Prelab:

- Have a clear understanding about Smith chart
- Understand how to measure VSWR of a system

b) Objective:

- To learn how to use smith chart to measure unknown impedance with slotted line using the minima shift method.

c) Equipment list:

1. Gunn power supply
2. Gunn oscillator
3. PIN-diode modulator
4. **Isolator:** The isolator is a two port device with small insertion loss in forward direction and a large in reverse attenuation. It thus allows power flow in one direction only. It can thus absorb reflected power from a mismatched load and isolate the Gunn source.
5. Variable attenuator
6. Slotted line
7. Slide screw tuner
8. Matched termination
9. SWR indicator

d) Theory:

Smith Chart:

Smith chart is a graphical indication of the impedance of a transmission line and of the corresponding reflection coefficient as one moves along the line. For a load impedance of Z_L the reflection coefficient is:

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0} \tag{4.1}$$

And

$$\Gamma = |\Gamma|\angle\theta_\Gamma = \Gamma_r + j\Gamma_i \tag{4.2}$$

Where $Z_L = R + jX$ =load impedance and Zo is the characteristics impedance. Normalized impedance z_L is

$$z_L = \frac{Z_L}{Z_0} = r + jx \tag{4.3}$$

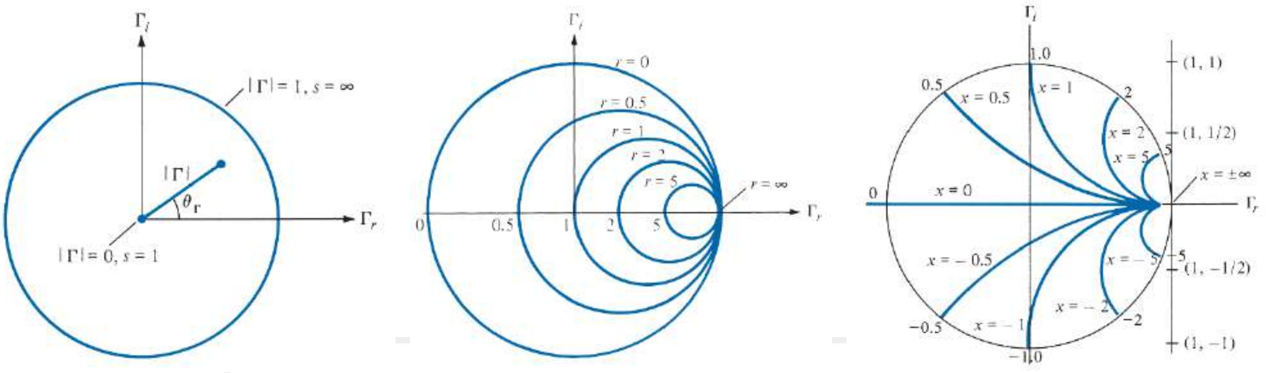
From equations 4.1, 4.2 and 4.3:

$$\Gamma = \Gamma_r + j\Gamma_i = \frac{z_L - 1}{z_L + 1} \tag{4.4}$$

The Smith chart is constructed within a circle of unit radius ($|\Gamma| \leq 1$) as shown is Fig 4.1(a). Equation 4.4 can be rearranged as:

$$z_L = r + jx = \frac{(1 + \Gamma_r) + j\Gamma_i}{(1 - \Gamma_r) - j\Gamma_i} \tag{4.5}$$

Using equation 4.5 we can get equation of circle as a function of r and x. By plotting all the r circle Fig 4.1(b) and x circle Fig 4.1(c) and combining them we get the Smith chart (see appendix for the full smith chart.



(a) (b) (c)

Fig 4.1: (a) $|\Gamma| \leq 1$ Unit circle. (b) r Circle (c) x Circle

Impedance measurement:

The VSWR, S can be expressed as a function of reflection coefficient as:

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

$$\Rightarrow |\Gamma_L| = \frac{S - 1}{S + 1} \tag{4.6}$$

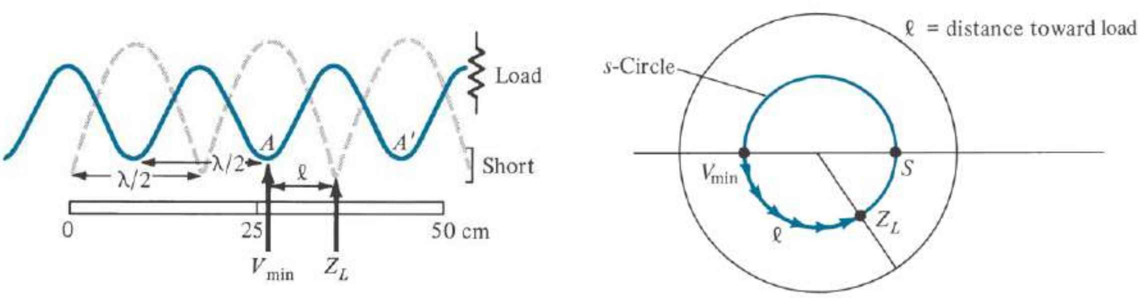
The angle of rotation of the phase of 'θ_Γ' at a distance 'd' from the load is determined by

$$\theta_\Gamma = \frac{2\pi d}{\lambda_g} \tag{4.7}$$

Once we know the Γ_L by combining equation 4.6 ,4.7 and using equation 4.4 we can easily determine the normalized impedance using equation 4.3.

Impedance measurement using minima shift method with Smith Chart:

1. With the load connected, read VSWR, (s) using the VSWR meter. Draw the s-circle on the Smith chart.
2. With the load replaced by a short circuit, locate a reference position for Z_L at a voltage minimum point.
3. With the load on the line, note the position of V_{min} and determine ℓ as shown in Fig 4.2(a).
4. On the Smith chart, move toward the load a distance ℓ from the location of V_{min}. Find Z_L at that point as shown in Fig 4.2(b).
5. If the shift is to the left, then the load is inductive + resistive and if the shift is to the right, then the load is capacitive + resistive. If there is no shift or exactly $\lambda_g/4$ shift, then the load is purely resistive.



(a) (b)

Fig 4.2: (a) Minima shift due to short and Load (b) Impendace measurement using Smith chart

The method described here is based on the fact that the waveguide is assumed to be lossless, otherwise traces of smith chart will be spiral rather than a circle (We will see this effect in the Network Analyzer experiment). In a lossy line, the SWR increases when the point of observation moves towards the load and decreases towards the generator.

e) Experiment procedure:

e1. Minima shift method with Smith Chart

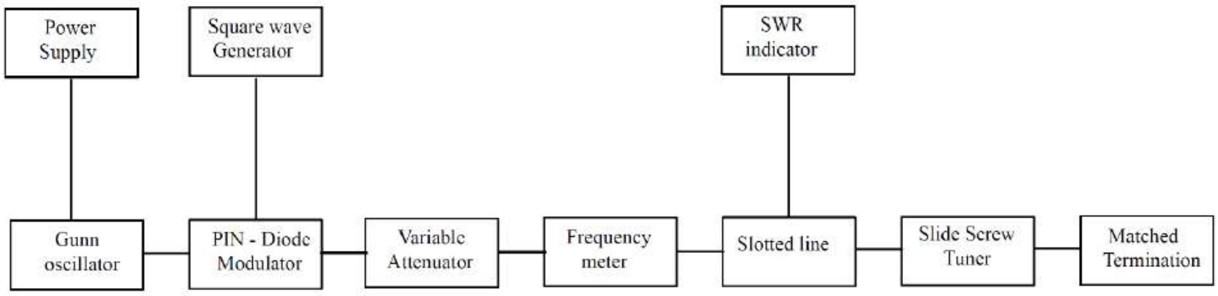


Fig 4.3: Setup for the experiment

1. Set up the equipment as shown in the Fig 4.3. The Load is matched in the setup.
2. Tune the probe of the *slide screw tuner* such that the depth of the probe is approx. 5 mm. Keep the position of the tuner in a fixed position. This will make the Load mismatched (inductive or capacitive)
3. Set the SWR meter such that maximum deflection is observed.
4. Measure VSWR using any of the techniques mentioned in the previous experiment.

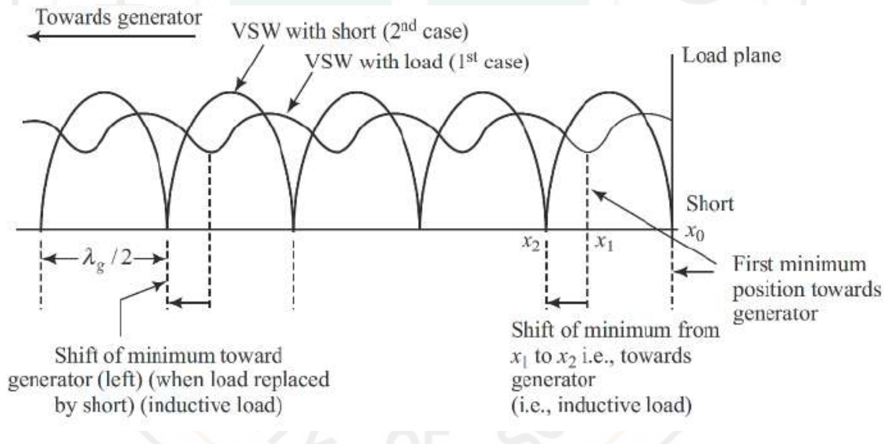


Fig 4.4: Minima shift due to change in load.

5. Starting from the load side move the probe along the slotted line until a maximum deflection is observed on the SWR indicator.
6. Adjust the SWR indicator until the meter indicates 1.0.
7. Move the probe along the slotted line until min deflection is observed. This is the X1 point in Fig 4.4.
8. Remove the *slide screw tuner* and the *matched termination* from the setup. Place a *shorting plate* to the slotted line.
9. Measure how much the minimum distance has shifted by locating the X_2 point.
10. Obtain the distance, d of two consecutive minimum point and determine the guided wavelength, $\lambda_g = 2d$
11. Repeat step 3-10 with a different probe depth and position.

Probe Depth (mm)	**Load VSWR** s	**Load Minima** x_1 (mm)	**Short Minima** x_2 (mm)	**Minima shift** $\ell = x_1 - x_2$ (mm)	**Guided Wavelength** $\lambda_g = 2d$	**Impedance of the load connected** $Z_L = R + jX$

Sample Calculation:

ℓ in term of wavelength $= \frac{\ell}{\lambda_g}\lambda =$

As the $\lambda = 720^{\circ}$ in smith chart, This corresponds to an angular movement in s circle in degree =

So the normalized complex impedance (from the Smith chart) is, z_L=

Actual impedance of the load, $Z_L = z_0 z_L =$

e2. Minima shift method without using Smith chart:

Fill up the following table using the data from previous table and hence calculate the load impedance.

$\lvert\Gamma_L\rvert = \frac{S-1}{S+1}$	x_1	x_2	λ_g	$\phi = 2\beta(x_1 - x_2) - \pi = \frac{4\pi}{\lambda_g}(x_1 - x_2) - \pi$	$\Gamma_L = \lvert\Gamma_L\rvert e^{j\phi} = \lvert\Gamma_L\rvert(\cos\phi + j\sin\phi)$	$Z_L = \frac{1-\Gamma_L}{1+\Gamma_L}$

f) Report:

1. Draw respective block diagrams of experimental set-ups and briefly explain the function of each part.
2. How the characteristics impedance (z_0) of the system should be actually calculated
3. Discuss the results.
4. Explain (a) Smith Chart (b) Impedance measuring technique with respect to a microwave signal transmission line.

Experiment no: 5

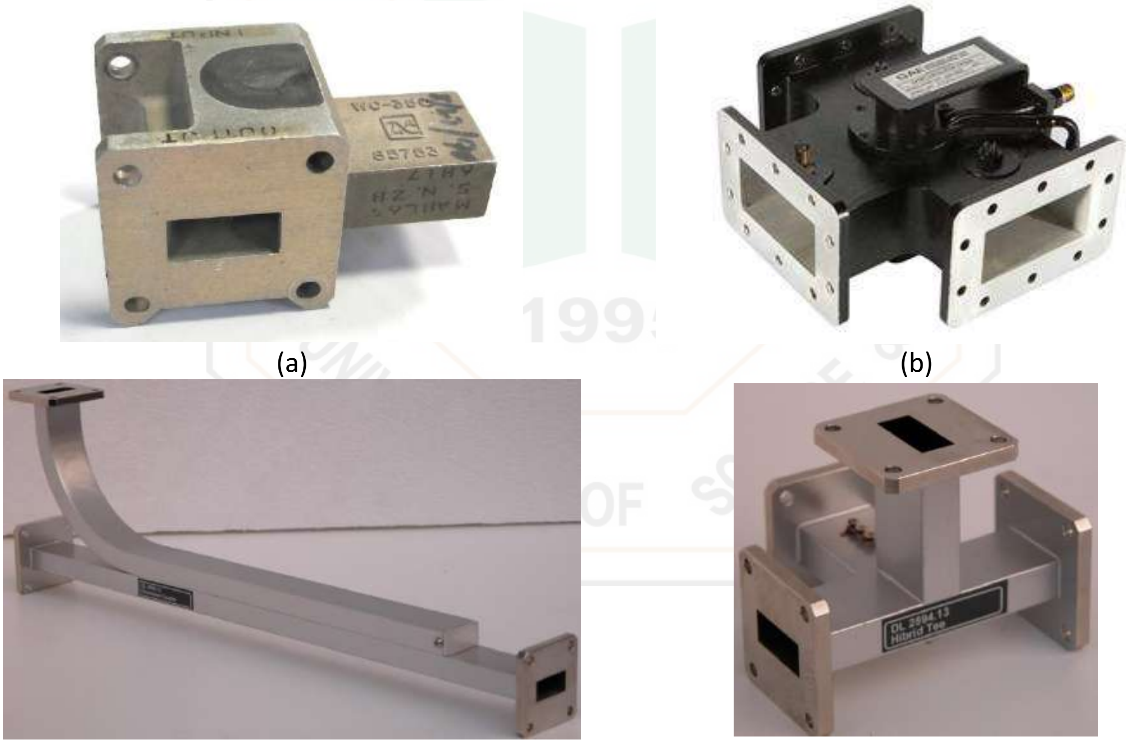
Name of the Experiment: Study of Waveguide components

a) Objective:

1. To learn the basic properties of a directional coupler.
2. To understand the basic principle of hybrid tee.
3. To study isolators and circulators.

b) Equipment List:

1. Gunn power supply and Oscillator
2. PIN-diode modulator
3. Variable attenuator
4. Crystal Detector
5. SWR indicator
6. Matched termination
7. Waveguide to Coax adaptor
8. Thermocouple mount
9. Power Meter
10. Frequency meter
11. Waveguide section
12. Isolator, Circulator, Directional coupler, Hybrid Tee.



(a) (b) (c) (d)

Figure 5.1: (a)Isolator (b)Circulator (c)Directional Coupler (d) Hybrid Tee

c)Theory:

- A N-Port network is by described by $N \times N$ size square matrix.
- For a *Symmetric network* if $[S] = [S]^T$ then the network is said to be ***reciprocal***.
- For a *Unitary Network* if

$$\sum_{k=1}^{N} S_{ki}S_{ki}^{*} = 1 \, and \sum_{k=1}^{N} S_{ki}S_{kj}^{*} = 0 \;\; Here \; i \neq j$$

  then the network is said to be ***lossless. (see Appendix E)***
- If $S_{ii} = 0$ then the i[th] port is said to be ***matched***.

C 1) Two Port Network

C 1.1) Isolator:

Isolator is a two-port device having unidirectional transmission characteristics. transmission occurs only in the direction from port 1 to port 2 but not the other way around.

S parameter for a two port network is:

$$S = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix}$$

As both the port are matched in an isolator $S_{11} = S_{22} = 0$. From the above discussion power can only transmit from port 1 to port 2 ($S_{21} = 1$) and no power should be reflected from 2 to port 1 ($S_{12} = 0$).

$$S = \begin{bmatrix} 0 & 0 \\ 1 & 0 \end{bmatrix}$$

Since the scattering matrix is not unitary, the isolator must be lossy. And also as the [S] is not symmetric isolator is a nonreciprocal component.

C2) Three Port Network:

C 2.1) Circulator

The circulator is a three port junction that permits wave transmission in only direction. Clockwise or anti clockwise.

For a three port device for all port to be *matched* and network to be *non-reciprocal,* the S-Matrix is:

$$[S] = \begin{bmatrix} 0 & S_{12} & S_{13} \\ S_{21} & 0 & S_{23} \\ S_{31} & S_{32} & 0 \end{bmatrix}$$

For the network to be *lossless*:

$$|S_{12}|^2 + |S_{13}|^2 = 1$$

$$|S_{21}|^2 + |S_{23}|^2 = 1$$

$$|S_{31}|^2 + |S_{32}|^2 = 1$$

$$S_{31}^{*}S_{32} = 0$$

$$S_{21}^{*}S_{23} = 0$$

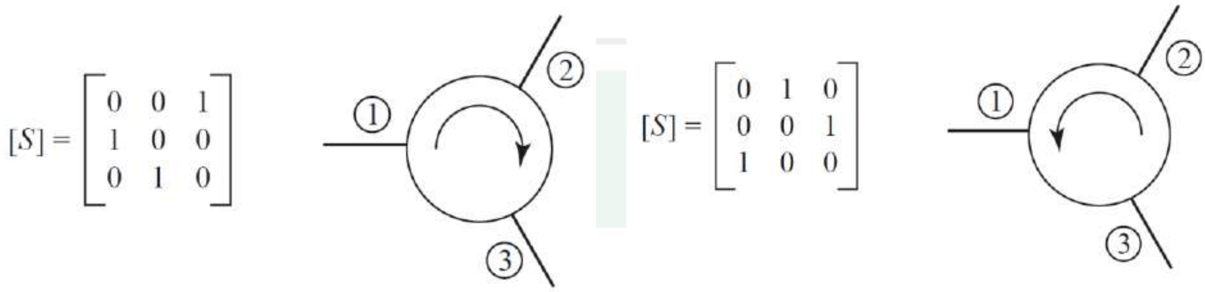
$$S_{12}^{*}S_{13} = 0$$

Here we have six parameters ($S_{12}, S_{13}, S_{21}, S_{23}, S_{31}, S_{32}$). To satisfy the 1st three equations, it can be seen $S_{12}\ and\ S_{13}$, $S_{21} and\ S_{23}, S_{31}\ and\ S_{32}$ cannot be zero at the same time. Also to satisfy *non-reciprocal* condition we have to maintain $S_{ij} \neq S_{ji}$ for $i \neq j$ ($i.e.\ S_{21} \neq S_{12}$). So to satisfy all six equations we can have only two combinations:

Combination 1: If $S_{12} = S_{23} = S_{31} = 0$ then $|S_{21}| = |S_{32}| = |S_{13}| = 1$

Combination 2: If $S_{21} = S_{32} = S_{13} = 0$ then $|S_{12}| = |S_{23}| = |S_{31}| = 1$

Putting the value from combination 1 and 2 into the S-Matrix we have two type of circulator, clockwise and counterclockwise as seen in the Fig



(a) (b)

Figure 5.2: Two types of circulators and their scattering matrices. (a) Clockwise circulation (b) Counterclockwise circulation

Just to explain $S_{12} = 0$ means that no power can flow from port 2 to post one also $S_{21} = 1$ means 100% power will flow from port 1 to port 2 and vice versa.

If the three-port network is allowed to be lossy, it can be reciprocal and matched at all ports which is the resistive power divider.

<u>Parameter associated with circulator:</u>

Insertion loss: The ratio of the power supplied by a source to the input port to the power detected by the detector in the coupling arm (i.e. output arm with other port terminated to the matched load) is defined as insertion loss or forward loss. For a CW circulator S_{21}, S_{32}, S_{13} are insertion loss. (ideally 1)

Isolation: It is the ratio of power fed in the input arm and the power detected at not coupled port with other port not terminated in the matched load. For CW circulator S_{12} , S_{23}, S_{31} are Isolation (ideally 0)

Input VSWR: The input VSWR of an isolator or circulator is the ratio of the voltage max to the voltage min of the standing wave existing on the line and the others have matched termination. For circulator VSWR can be determined from $S_{11},\ S_{22},\ S_{33}$ (Ideally 0 for perfectly matched port i.e. VSWR=1)

<u>C3) Four Port Network</u>

C 3.1) Directional coupler

It is a four port/three port sampling device that does not introduce reflections to the main systems. Power supplied to port 1 is coupled to port 3 (the coupled port) with the coupling factor$|S_{13}|^2 = \beta^2$, while the remainder of the input power is delivered to port 2 (the through port) with the coefficient $|S_{12}|^2 = \alpha^2 = 1 - \beta^2$. In an ideal directional coupler, no power is delivered to port 4 (the isolated port).

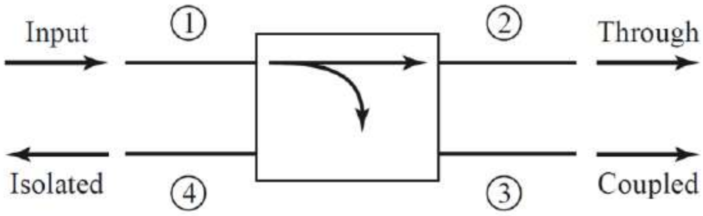


Figure 5.3: Directional coupler with power flow direction

For a four port device, the scattering matrix:

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{43} & S_{44} \end{bmatrix}$$

When all four ports are perfectly matched to the junction

$$S_{11} = S_{22} = S_{33} = S_{44} = 0$$

symmetric property of scattering matrix gives

$$(S_{12} = S_{21}, S_{23} = S_{32}, S_{13} = S_{31}, S_{24} = S_{42}, S_{34} = S_{43}, S_{41} = S_{14},)$$

As there is no coupling between ports 1 and 4, and also no coupling between 2 and 3, therefore

$$S_{14} = S_{41} = 0 \; and \; S_{23} = S_{32} = 0$$

All the above leads to scattering matrix with four unknown parameters:

$$[S] = \begin{bmatrix} 0 & S_{12} & S_{13} & 0 \\ S_{12} & 0 & 0 & S_{24} \\ S_{13} & 0 & 0 & S_{34} \\ 0 & S_{24} & S_{34} & 0 \end{bmatrix}$$

Applying unitary property on the above matrix it and further simplification can be made by choosing the phase references on three of the four ports shown that

$$S_{12} = S_{34} = \alpha, S_{13} = \beta e^{j\theta} \; and \; S_{24} = \beta e^{j\emptyset}$$

Here α and β are real and $\alpha^2 + \beta^2 = 1$. θ and $\emptyset$ are Phase constant where $\theta + \emptyset = \pi \pm 2n\pi$. So the Scattering parameter for a directional coupler is:

$$[S] = \begin{bmatrix} 0 & \alpha & \beta e^{j\theta} & 0 \\ \alpha & 0 & 0 & \beta e^{j\emptyset} \\ \beta e^{j\theta} & 0 & 0 & \alpha \\ 0 & \beta e^{j\emptyset} & \alpha & 0 \end{bmatrix}$$

Based on the value of Phase β (i.e. $\theta \; and \; \emptyset$)directional coupler can be of two types:

$$[S] = \begin{bmatrix} 0 & \alpha & j\beta & 0 \\ \alpha & 0 & 0 & j\beta \\ j\beta & 0 & 0 & \alpha \\ 0 & j\beta & \alpha & 0 \end{bmatrix}$$

Symmetric Coupler ($\theta = \emptyset = \pi/2$).

$$[S] = \begin{bmatrix} 0 & \alpha & \beta & 0 \\ \alpha & 0 & 0 & -\beta \\ \beta & 0 & 0 & \alpha \\ 0 & -\beta & \alpha & 0 \end{bmatrix}$$

An anti-symmetric Coupler($\theta = 0, \emptyset = \pi$).

So it can be said that any reciprocal, lossless, matched four-port network is a directional coupler.

Parameter associated with Directional Coupler:

The directional coupler that we have has 3-ports as the 4th isolated port is permanently terminated with a matched load.

Figure 5.4: Sampling direction of a directional coupler (a) forward wave (b) reflected wave.

$$Coupling\ coefficient = C = 10\log\frac{P_1}{P_3} = -20\log\beta = 10\log\frac{P_1}{P_{3F}}\,(dB)$$

$$Directivity = D = 10\log\frac{P_3}{P_4} = 20\log\frac{\beta}{|S_{14}|} = 10\log\frac{P_{3F}}{P_{3R}}\ (dB)$$

$$Isolation = I = 10\log\frac{P_1}{P_4} = -20\log|S_{14}| = 10\log\frac{P_1}{P_{3R}}\,(dB)$$

$$Insertion\ Loss = L = 10\log\frac{P_1}{P_2} = -20\log|S_{12}|\ (dB)$$

The coupling factor indicates the fraction of the input power that is coupled to the out- put port. The directivity is a measure of the coupler's ability to isolate forward and back- ward waves (or the coupled and uncoupled ports). The isolation is a measure of the power delivered to the uncoupled port. These quantities are related as I = D + C (dB). The insertion loss accounts for the input power delivered to the through port, diminished by power delivered to the coupled and isolated ports

Return loss measurement:

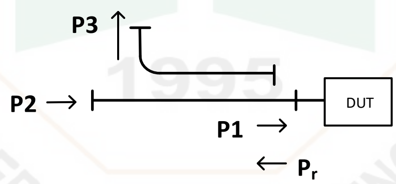


Fig 5.5: Return loss measurement

To measure return loss, the input signal is applied at the port 2 and the device under test (DUT) is connected to port 1 then the return loss signal is picked up at port 3. The power at port 3 when the coupling coefficient is C is:

$$(P_3)_{DUT} = \frac{P_r}{C}$$

Now if the DUT is replaced by a short all the power is reflected back and therefore P1 should appear at the port 3. The actual power at port 3:

$$(P_3)_{Short} = \frac{P_1}{C}$$

Since the voltage reflection coefficient of the DUT is given by, $|\Gamma| = \sqrt{\frac{P_r}{P_i}}$. The ratio of the two signals detected at port 3 is

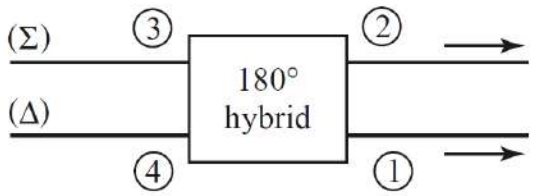
$$\frac{(P_3)_{Short}}{(P_3)_{DUT}} = \frac{P_1}{C} \times \frac{C}{P_r} = \frac{P_1}{P_r} = \frac{1}{|\Gamma|^2}$$

$$\therefore Return\ Loss, |\Gamma| = \sqrt{\frac{(P_3)_{DUT}}{(P_3)_{Short}}}$$

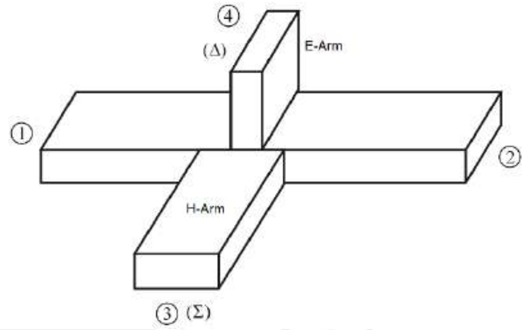
The accuracy of return loss measurement is dependent on directivity of the coupler, which describes how much of input power at port 2 leaks to port 3.

C 3.2) Waveguide Hybrid Junction / Magic-T

A magic tee (or magic T or hybrid tee) is a hybrid or 3 dB coupler used in microwave systems. The magic tee is a combination of E and H plane tees. Arm 3 forms an H-plane tee with arms 1 and 2. Arm 4 forms an E-plane tee with arms 1 and 2. Arms 1 and 2 are sometimes called the side or collinear arms. Port 3 is called the H-plane port, and is also called the Σ port, sum port. Port 4 is the E-plane port, and is also called the Δ port, difference port.



(a)



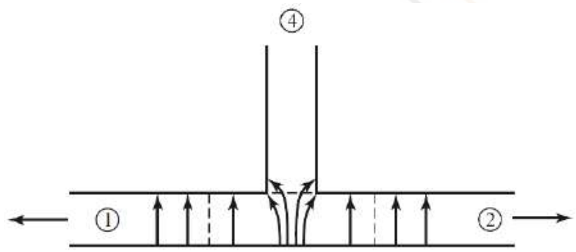
(b)

Figure 5.6:(a) Symbol for a 180° hybrid junction (b) A waveguide hybrid junction or magic-T.

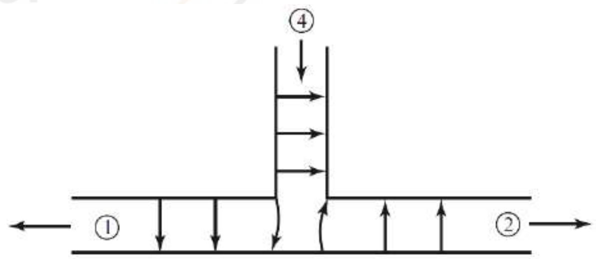
When a signal applied to port 3 it will be evenly split into two in-phase components at ports 1 and 2 ($S_{23} = S_{13}$)and port 4 will be isolated ($S_{43} = 0$).

If the input is applied to port 4, it will be equally split into two components with a 180◦ phase difference at ports 1 and 2 ($S_{24} = -S_{14}$)and port 3 will be isolated ($S_{14} = 0$)

When operated as a combiner, with input signals applied at ports 1 and 2, the sum of the inputs will be formed at port 3, while the difference will be formed at port 4. Hence, ports 3 and 4 are referred to as the sum and difference ports respectively.



(a)



(b)

Figure 5.7: Electric field lines for a waveguide hybrid junction. (a) Incident wave at port 3. (b) Incident wave at port 4.

First consider a TE_{10} mode incident at port 3. The resulting E_y field lines are illustrated in Figure 7.50a, where it is seen that there is an odd symmetry about guide 4. Because the field lines of a TE10 mode in guide 4 would have even symmetry, there is no coupling between ports 3 and 4. There is identical coupling to ports 1 and 2, however resulting in an in-phase and equal-split power division.

For a TE_{10} mode incident at port 4 (Figure 7.50b). Again ports 3 and 4 are decoupled, due to symmetry (or reciprocity). Ports 1 and 2 are excited equally by the incident wave, but with a 180◦ phase difference.

For a four port device, the scattering matrix:

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{43} & S_{44} \end{bmatrix}$$

It can be said from previous discussion, the H-plane tee section symmetry around port 1 $(S_{23} = S_{13})$ and port 4 $(S_{24} = -S_{14})$. Also input at port 3 cannot come out of port 4 as they are isolated $(S_{34} = S_{43} = 0)$.

Due to *symmetry* property $(S_{12} = S_{21}, S_{13} = S_{31}, S_{34} = S_{43}, S_{24} = S_{42}, S_{41} = S_{14})$. We also assume port 3 and 4 are perfectly matched $(S_{33} = S_{44} = 0)$.

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{12} & S_{22} & S_{13} & -S_{14} \\ S_{13} & S_{13} & 0 & 0 \\ S_{14} & -S_{14} & 0 & 0 \end{bmatrix}$$

Applying the *unitary* property we can find out $S_{11} = S_{22} = 0$ and also $S_{13} = S_{14} = \frac{1}{\sqrt{2}}$. This also shows $S_{12} = 0$.

$$[S] = \begin{bmatrix} 0 & 0 & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \\ 0 & 0 & \frac{1}{\sqrt{2}} & -\frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & 0 & 0 \\ \frac{1}{\sqrt{2}} & -\frac{1}{\sqrt{2}} & 0 & 0 \end{bmatrix}$$

$$= \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & 0 & 1 & 1 \\ 0 & 0 & 1 & -1 \\ 1 & 1 & 0 & 0 \\ 1 & -1 & 0 & 0 \end{bmatrix}$$

From the S-matrix it can be said that the magic of magic Tee are (i). If any two ports are perfectly matched to the junction, then the remaining two ports get, automatically matched to the junction and (ii). The signal input at port 1 does not come out of port 2, in spite of being collinear; i.e. ports 1 and 2 are isolated. Similarly ports 3 and 4 are isolated.

(d) Experiment procedure:

D1. Directional coupler measurement

D 1.1: Coupling factor measurement

1. Set up the equipment as shown in Figure 5.8(a). Set the variable attenuator at 20dB. Apply 1000Hz modulation signal to pin diode modulator and turn on to the Gunn oscillator. Read the SWR indicator. Use this value as reference.

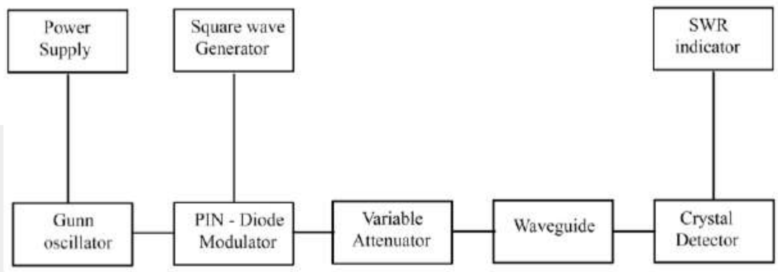


Figure 5.8 (a): Initial setup for coupling factor measurement

2. Replace the waveguide with the directional coupler Figure 5.8(b). Move the crystal detector to the auxiliary arm of the coupler.

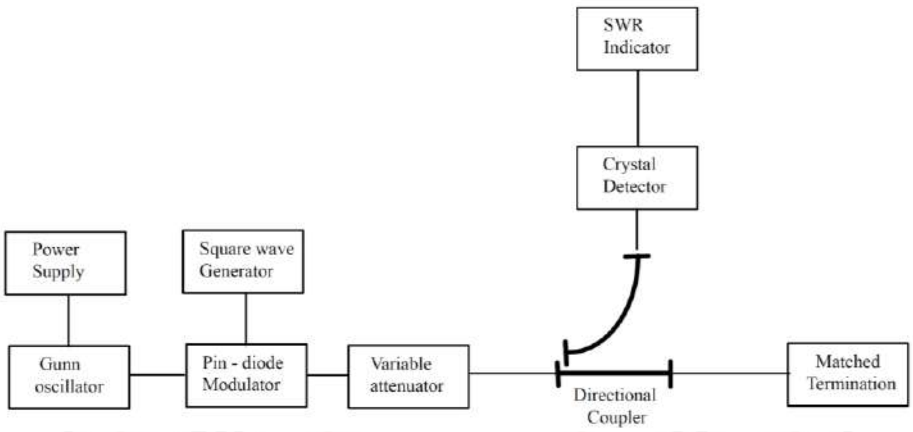


Figure 5.8(b): Setup diagram for coupling factor measurement

3. Adjust the variable attenuator until the same reference reading as in (1) is obtained.

A_1 (dB)	A_2 (dB)	A_3 (dB)	$A_1 - A_2$ (dB)	A_4 (dB)	$A_3 - A_4 + (n \times 10)$ (dB)

4. Fill in the table above with attenuation of the attenuator. The coupling factor of the directional coupler is A1-A2.

D 1.2: Directivity measurement

1. Set the attenuator to 20 dB.
2. Read the SWR. Use this value as reference. Record the attenuator setting.

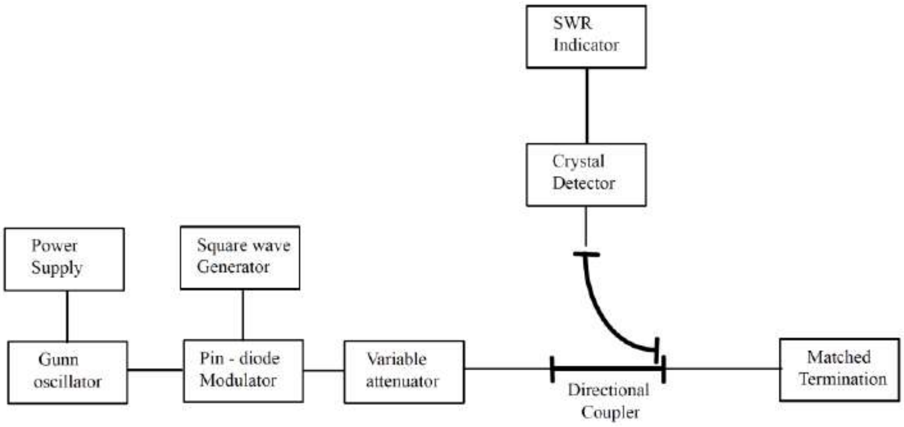


Figure 5.8(c): Directivity measurement

3. Change the coupler orientation as shown in Figure 5.8(c).
4. Reduce the attenuation and increase the SWR indicator gain by 10 dB steps until the same value in (2) is obtained. The directivity is $A_3 - A_4 + (n \times 10)$ dB.

AUST

D 1.3: Return loss measurement

1. Set up the equipment as shown in Figure 5.8(d).

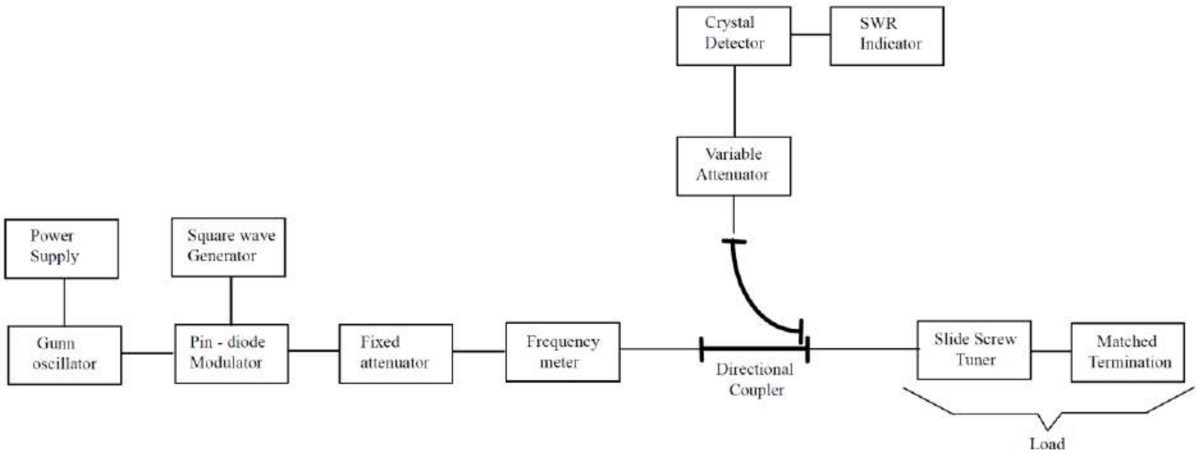


Figure 5.8(d): Return loss measurement

2. Set the probe depth of the slide screw tuner to 5mm.
3. Set the attenuator to 0dB (A5). Read the SWR indicator. Use this as reference.
4. Change the attenuator to maximum attenuation. Replace the load with a short.
5. Decrease the attenuator until the reference level in (3) is obtained. Record the attenuator position (A_5) in case it is necessary to change the range on the SWR indicator, add increase value to the position of the attenuator to get A_6.
6. The return loss = $A_5 - A_6 + (n \times 10)$

A_5 (dB)	A_6 (dB)	$A_5 - A_6 + (n \times 10)$ (dB)	Return Loss $\lvert\Gamma\rvert$	SWR

D2. Hybrid- T measurement

D 2.1: Initial adjustments

1. Set the equipment according to Figure 5.9(a).

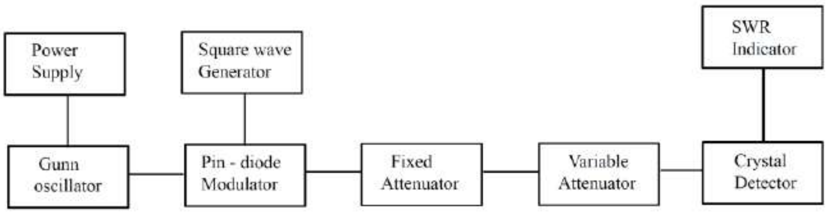


Figure 5.9(a): Initial Setup Hybrid T

2. Adjust the SWR indicator gain for obtaining any convenient deflection.
3. Apply 9 volt to the Gunn oscillator.
4. Apply modulation voltage to the pin diode modulator.
5. Adjust the offset voltage and the pulse freq of the square wave generator to obtain max deflection on the SWR indicator.

D 2.2: Measurement of decoupling between H-arm and E-arm

1. Set the equipment according to Figure 5.9(b).

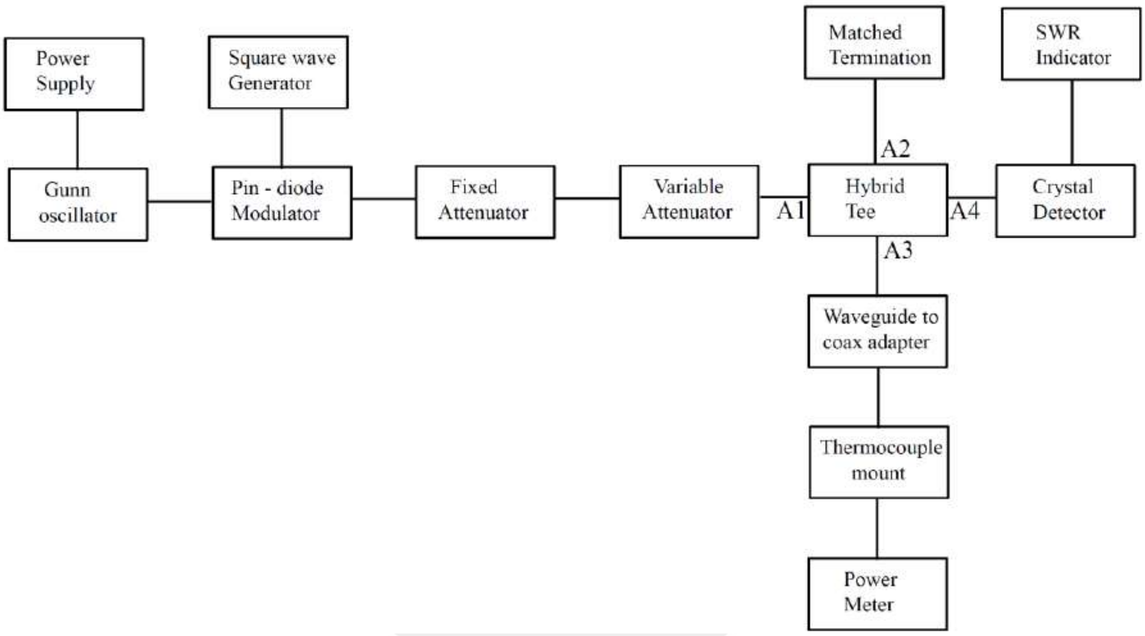


Figure 5.9(b): Setup diagram for measurement of decoupling between H arm and E arm.

2. Set the attenuator to 20 dB (A1)
3. Select the range on SWR indicator which gives a reasonable deflection on the indicator. Adjust the gain control to a reference reading on the dB scale of the indicator.
4. Remove the detector and connect the variable attenuator to arm1.
5. Connect matched termination and power meters to arm2 and arm3 and connect the detector to arm4. Keep the power meter at off position.
6. Increase the sensitivity of the SWR indicator in 10 dB increment until the same reference level as in (3) is obtained. The attenuation (Az) of the attenuator may be reduced if necessary.
7. Record the result as in table.

Attenuation of the variable attenuator		Variations of the SWR Meter Gain (in 10dB steps) n steps	Decoupling $A_1 - A_2 + (n \times 10)$ (dB)
A_1 (dB)	A_2 (dB)		

D 2.3: Measurement of insertion loss

1. In the Connect the detector to the attenuator which is set to be 20dB.
2. Select a range on the SWR indicator which gives a reasonable deflection on the indicator. Adjust the gain control to a reference reading on the dB scale of the indicator.
3. Remove the detector and connect the arm (1) of the hybrid T to the attenuator.
4. Connect the matched termination and the power meters to arm3 and arm4. Also connect the detector to arm2.
5. Decrease the attenuator (A4) until the same reference level as in (2) is obtained. The insertion loss between arm 1 and arm 2 is A3-A4.
6. For insertion loss between arm 1 and arm 3 repeat (4) and (5).
7. For insertion loss between arm 4 and arm 2 repeat (3), (4) and (5)
8. Record the results.

Insertion loss	Attenuation		Absolute Value
	A3 (dB)	A4 (dB)	
1-2	**20**		
1-3	**20**		
4-2	**20**		

D 2.4: Return loss Measurement of H-arm

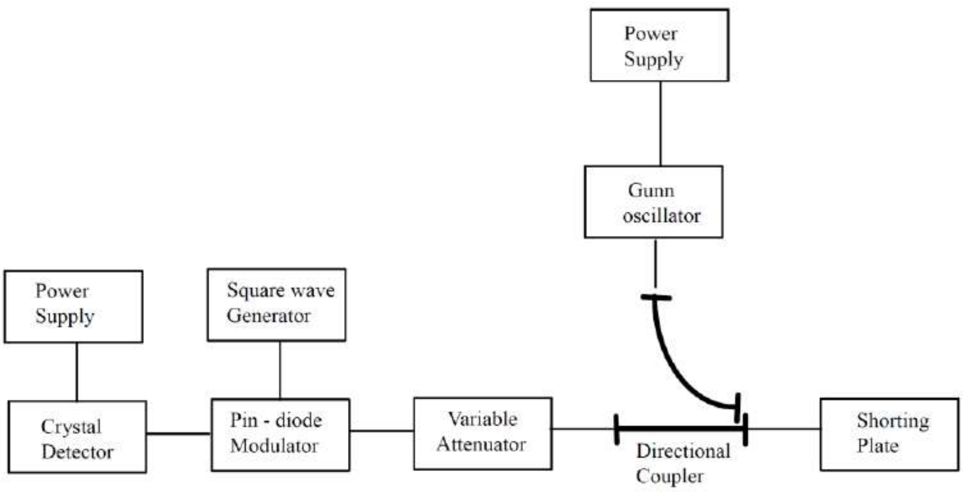


Figure 5.9(c): Initial setup for return loss measurement.

1. Set up the equipment as in Figure 5.9(c).
2. Turn on the oscillator. Do not exceed 9 V. Adjust the square wave output for the max modulation.
3. Set the variable attenuator to 20 dB (A5)
4. Set up the reference point on the SWR indicator.

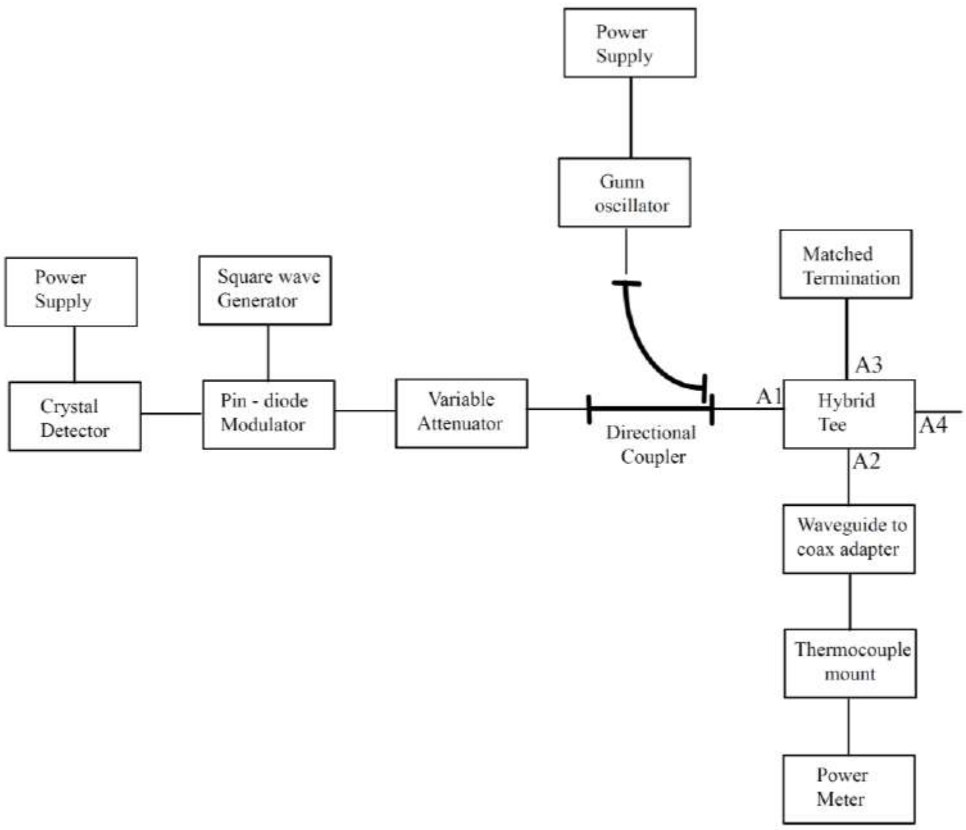


Figure 5.9(d): Return loss measurement setup.

5. Remove the shorting plate. Connect arm1 to the directional coupler as shown in Figure 5.9(d). Connect a matched termination to arm3. Leave arm4 open. (arm4 is the E-plane T. Since the decoupling to arm4 is almost 30-40 dB leaving it open should not affect the accuracy).
6. Increase the gain of the SWR indicator in 10 dB increments. Decrease the attenuation (A6) until the same level as in (4) is obtained. Record the results.
7. Repeat (5) and (6) using E-plane T (arm4) instead of H-plane T (arm1)

Object	**Attenuation**		**Gain increase of the SWR meter in 10dB steps**	**Return Loss**	
	A_5 (dB)	**A_6 (dB)**	$\boldsymbol{A_6 - A_5 + (n \times 10)}$	**dB**	**Absolute value**
Arm 1					
Arm 4					

D3. Circulator and Isolator measurement

D 3.1: Input VSWR measurement

1. Set up the equipment as shown in Figure 5.10.
2. Connect the isolator or circulator in the direction of flow of energy. The input port is towards the gunn and the output port is connected to matched termination. 3. Energize the microwave source for 10GHz of frequency.
3. With the help of the slotted line, probe and VSWR meter find out SWR of the isolator or circulator for low and medium SWR measurements.
4. Repeat the procedure for other frequencies.

D 3.2: Measurement of insertion loss and isolation

1. Remove the isolator and circulator from slotted line. Connect the detector at end of slotted line. The output of the detector mount should be connected to the VSWR meter.
2. Energize the microwave source for maximum output for a particular frequency of operation (eg. 10GHz). Tune the detector mount for maximum output in the VSWR meter.
3. Set any reference level of a power in VSWR meter with the help of a variable attenuator and gain control knob of the VSWR meter. (P1)
4. Remove the detector mount from slotted line without disturbing others. Insert isolator/circulator between slotted line and detector mount. Keeping input port to the slotted line and detector mount.
5. Record the reading in the VSWR meter. If necessary, change range dB switch to high or lower position and taking 10 dB change for one step change for one step change of switch position (P2).
6. Compute insertion loss on P1-P2 in dB.
7. For measurement of isolation, the isolation or circulation has to be connected in reverse, the output port to the slotted line and detector to input port with other port terminated by matched termination after setting a reference level without isolator or circulator in the set up as described in insertion loss measurement. Let same P1 level is set.
8. Record the reading of VSWR meter inserting the isolator or circulator as given in step 7. Let it is P3.
9. Do the same for other ports of circulator. Always terminate the unused port of the circulator.
10. Repeat the same for other frequencies.

E) Report:

1. Give example and block diagram of the system where Isolators, circulator, directional couplers and Hybrid tee are being used.
2. What is a reciprocal network? Give example with mathematical basis.
3. Define Insertion loss, return loss, Isolation, coupling factor with necessary diagram.

Experiment no: 6
Name of the Experiment: Inverse square law of Electromagnetic Wave Propagation.

6.1 Objective:

1. To understand the function of antenna as a radiating structure.
2. To be familiarized with antenna transmitter, receiver and stepper motor controller
3. To be familiarized with various antenna types.
4. To study the variation of field strength of radiated wave with distance from transmitting antenna (verification of inverse square law).
5. To understand the Friis radiation formula.

6.2 Equipment list:

1. Antenna Transmitter

It delivers necessary electrical signal to the antenna so that it can be transmitted in the form of electromagnetic wave. Fig. 6.1(a) & (b) shows the front panel layout of AMITEC & FALCON antenna transmitter respectively. The following controls are similar for both the cases-

Fig. 6.1 (a) Front panel layout of transmitter (AMITEC)

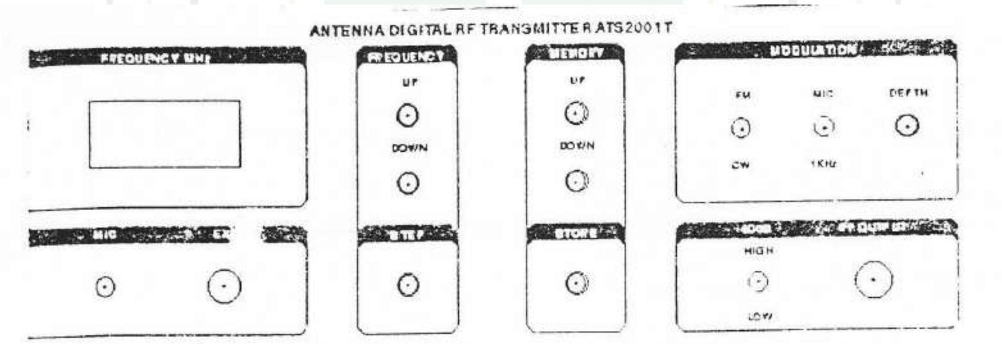


Fig. 6.1 (b) Front panel layout of transmitter (FALCON)

LCD (is used to display the frequency of the signal generated. The range is 86-860 MHz. The frequency displayed on Power ON is the frequency stored in the memory before power was switched off. For AMITEC, various step size for scrolling the frequency upward/downward are available from 50, 100, 250, 500 KHz, 1, 10, 100 MHz. For ATS, the resolution is 100 KHz with accuracy better than 10 KHz.)

UP/DOWN (increases/decreases generated frequency by selected steps),

MIC (connects the condenser microphone to frequency modulate the voice signal in the carrier signal displayed).

FM/CW (This toggle switch is used to select the modulation; CW is used for taking antenna measurements as the level remains stable; FM is used to frequency modulate voice etc. for communication).

EXT (This BNC input is used to connect any external signal to frequency modulate the generated carrier).

RF OUT (this is where the transmitted signal is present; for AMITEC & ATS, output impedance are 50 & 75 ohms respectively; the transmitting antenna can be connected to it with BNC lead provided).

HIGH LOW (the toggle switch is used to adjust the output level of the transmitted signal; high level is 110 dBµV (3 dBm for 50 ohm line) & low level is 70dbµV. In case of AMITEC, it is a pin diode attenuator.

<u>AMITEC- trainer</u>

ENTER for AMITEC (Purpose is to store a particular frequency in the current location of memory and also to select & store a particular step size and initiate serial dump. Frequency and level both are stored at any desired memory location on pressing this button. This display will blink to indicate that frequency has been changed.

MENU (is used to select the operation modes like frequency step size from 50 KHz to 100 MHz. Also to change from Manual to auto mode).

ESCAPE (is used to cancel any command and revert to default position)

FM deviation (is used to vary the frequency deviation of the FM signal. Rotating CW will increase the deviation and vice versa)

<u>FALCON trainer</u>

Memory UP & DOWN (is used to increment/decrement memory locations. There are 63 locations. On pressing the switch, the location number will be displayed in place of frequency for few secs after which the display revert back to frequency for that location.). Pressing it long will start the scroll mode & locations will start rolling,

STORE (is used to memorize a particular frequency to a specific location. This will blink to indicate that frequency has been stored)

STEP (is used to to select the frequency step size from default of 1 MHz to 100 KHz).

<u>Antenna Receiver</u>

Electromagnetic wave received from antenna is converted to electrical signal. Fig. 6.2(a) & 6.2 (b) shows the front panel layout of AMITEC & FALCON antenna receivers respectively. The following controls are similar for both the cases-

Fig. 6.2 (a) Front panel layout of receiver (AMITEC)

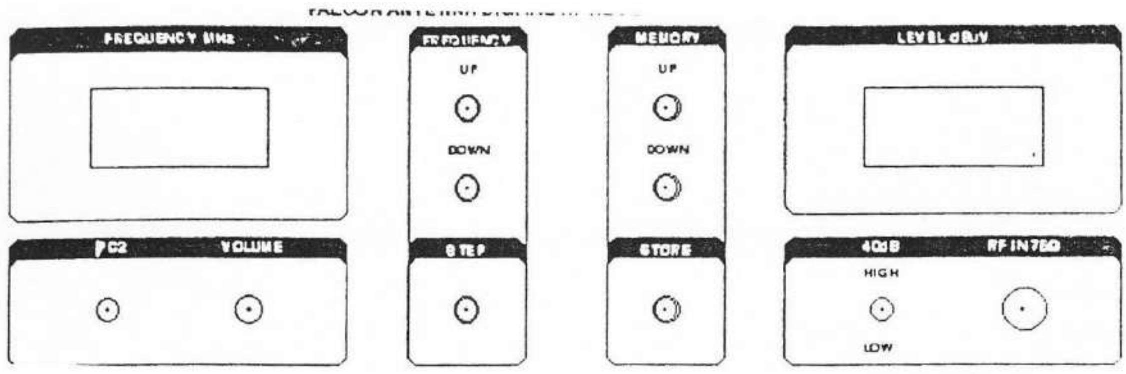


Fig. 6.2 (b) Front panel layout of receiver (FALCON)

LCD (is used to display the frequency of the signal received. The range is 86-860 MHz. The rest is similar to the transmitter LCD)

UP/ DOWN (increases/ decreases received frequency by selected steps)

RF IN (this is where the received signal from antenna is present for measurement. For AMITEC & ATS, input impedance are 50 & 75 ohms respectively; the receiving antenna can be connected to it with BNC lead provided).

HIGH LOW (the toggle switch is used to adjust the input level of the received signal; in low level the sensitivity is 40 dB down).

<u>AMITEC only</u>

ENTER for AMITEC (purpose is same as transmitter)

MENU (purpose is same as transmitter).

ESCAPE (purpose is same as transmitter)

FM demodulation (it gives demodulated output at CRO)

RS 232 (used to dump serial data with the help of RS232 cable into PC. If MENU key is pressed a number of times till serial mode dump appears. If it is No press up key and it will toggle to Yes. Now press Enter key and LCD will display uploading. It will actually upload whatever data has been stored in instrument to PC. Before uploading data into PC, open the software GUI and select the comport where RS232 lead from instrument is to be connected to PC. Suppose lead has been connected to COM1 to PC. Select Com1 from software GUI, then dump data from instrument.)

Down converter (It is a 39 MHz output which can be connected to any spectrum analyzer for viewing FM modulation and received RF level of receiver. When received signal decreases, it also decreases.)

RSSI (stands for received signal strength indicator. It is a DC output corresponding to received RF level. It can be viewed on CRO in DC coupled mode or a multimeter)

Stepper in (A stepper in cable with a switch is connected at the end of cable can be connected at this input. Put receiver in auto mode using menu and press the switch at different angular positions by rotating the antenna to be plotted in step size of 5 degrees etc. It will advance the memory location and store the received signal at different angular position.)

FALCON only

Memory UP & DOWN (purpose is same as transmitter.)

STORE (purpose is same as transmitter)

PC2 (this earphone socket is used to connect lead to the „line in‟ of computer. An internally generated 1 KHz signal whose level varies in proportion to the received signal strength is sent to the sound card of computer. The software in computer then plots the pattern of the antenna)

STEP (purpose is same as transmitter).

Stepper motor controller

Fig. 6.3(a) & 6.3 (b) shows the front panel layout of AMITEC & FALCON stepper motor controller respectively. The following controls are similar for both the cases-

Fig. 6.3 (a) Front panel layout of stepper motor controller (AMITEC)

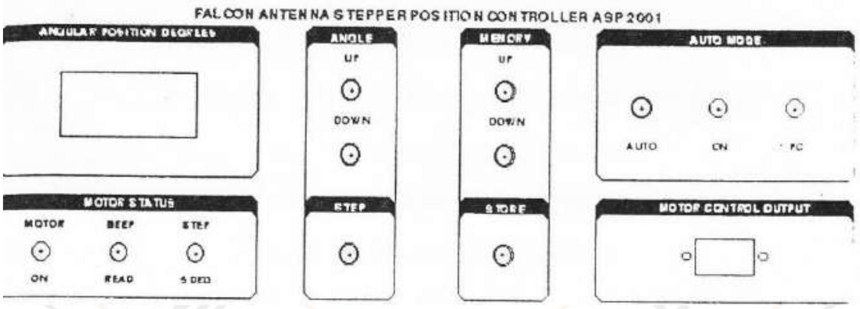


Fig. 6.3 (b) Front panel layout of stepper motor controller (FALCON)

Motor on (LED lights up to indicate that stepper motor is running)

Beep (This buzzer is used to indicate that the motor has reached its desired/specified position and readings can be taken)

UP (increases angular position of motor tripod by selected steps, pressing it longer starts the scroll mode and position will start to roll slowly and then faster. There is delay of few sec after which the motor starts to rotate).

DOWN (decreases angular position of motor tripod by selected steps & rest is same as UP).

Auto (is used to initiate the auto rotation mode. All other switches are disabled in auto mode. Pressing again will switch off the auto mode)

Motor control output (This 9 pin socket is used to connect the poles of the stepper motor to the controller).

LCD (to display the angular position of the motor tripod along with memory locations step size etc. Display is 0 degree on power reset)

<u>AMITEC only</u>

POS (this LED is used to indicate that the motor has reached its specified location and readings can be taken)

MENU (used to select the operation modes like angular step size from 1,5,10, 45 degree. Also to change from Manual to auto mode).).

ESCAPE (purpose is same as transmitter)

Enter is used to store a particular angular position in the current location of memory and also to select and store a particular step size)

Pulse (this phone socket is used to connect lead to the PULSE input of receiver. A 10ms pulse is generated internally and supplied to the receiver every time the motor reaches its specified location. This triggers the receiver to take reading of RF signal level)

Trigger out (it's a BNC o/p which is used to connect additional stepper motor controller. Instead of pressing pulse key every time a manual reading is to be recorded, a stepper motor controller will send pulse every time stepper motor rotates by 1or 5 degree and with receiver and stepper motor controller in auto mode the pulse from motor controller will advance and log readings in memory locations.)

<u>FALCON only</u>

Memory UP & DOWN (is used to increment memory locations. There are 99 locations. On pressing the switch, the location number will be displayed in place of angular position for few secs after which the display revert back to frequency for that location.)

STORE (purpose is same as transmitter)

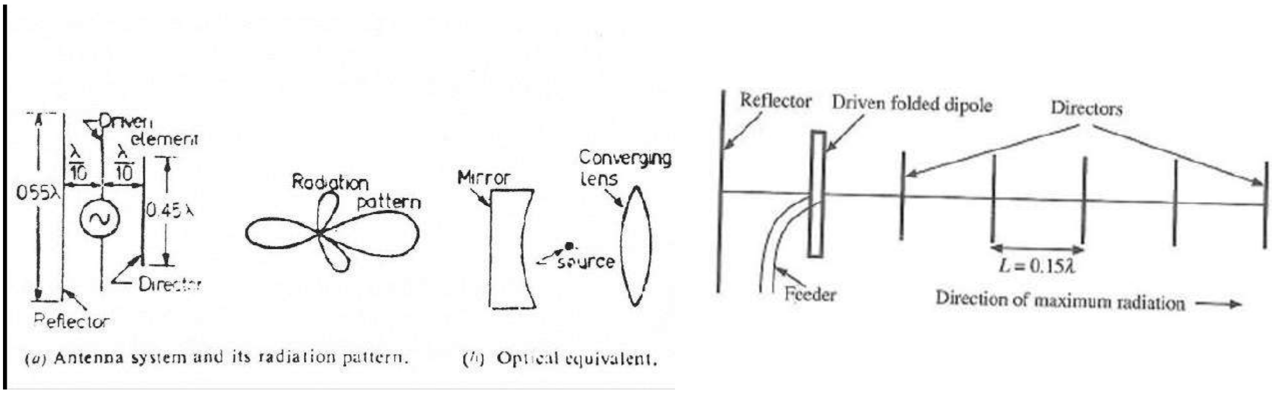
PC1 (this earphone socket is used to connect lead to the 'line in' of computer. A 10ms pulse is generated internally & send every time the motor reaches its specified location. This triggers the software to take reading of received signal level.

STEP (purpose is same as transmitter).

4. Antenna Tripod

5. Yagi-Uda and Yagi antenna

These are passive antenna arrays with a single driven element and other elements driven parasitically. The elements are strung out along the direction of propagation. The phase of the currents in each passive elements are such that when the phase delay is added for the wave to get from element to the next, the individual element currents all add contribution to the radiated field which are in phase with each other at the front of the antenna (Fig. 6.4).



(a) (b)

Fig. 6.4 (b) Yagi-Uda antenna

6.3 Theory:

6.3.1 Function of an antenna

Antennas are basically radiating structure which radiates electromagnetic energy into free space (Fig. 6.5). It is a transducer that converts electrical energy into radiating energy and vice versa. It transfers electric signal (it receives from its feeder circuit) into electromagnetic energy at its output. The power for such radiation is supplied by a feeder which is often a length of transmission line or a waveguide having well defined characteristics impedance. The process of electromagnetic wave propagation through an antenna is effected by an impedance transformation between space and source. The transmission line/ waveguide act as an antenna when its ends are flared or tapered in such a way that fields of the propagating EM wave expand in an ordinary manner thereby reduce the mismatch between guide/line and source. Antennas are actually tank circuit which resonates at a particular frequency (the radiating signal frequency).

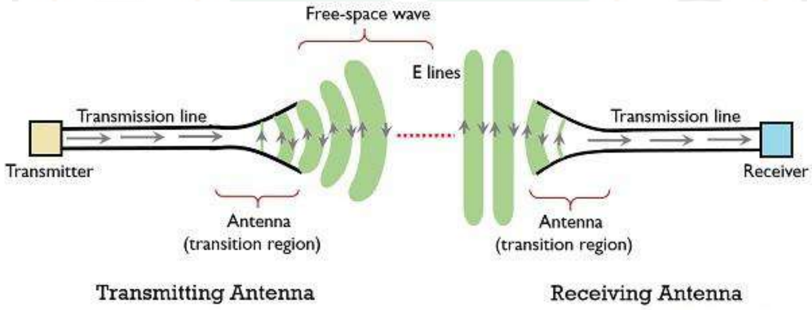


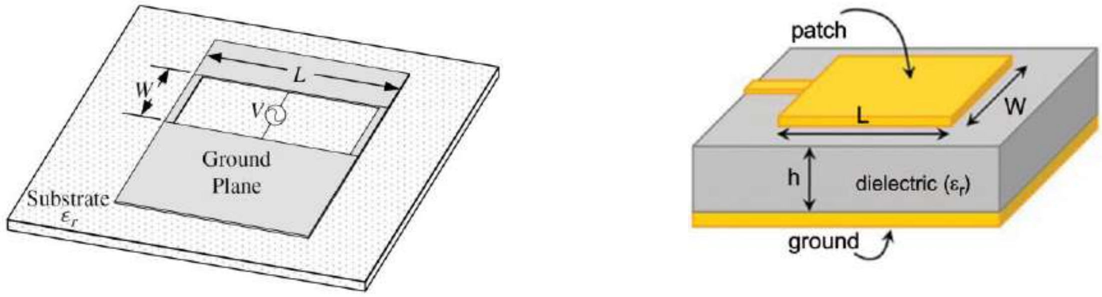
Fig. 6.5 Building up of energy radiation from antenna

6.3.2 Basic antenna types:

There are various types of antennas used for various purposes- dipole type, wire type, slot type, aperture type, parabolic type etc.

Slot type: Used in the frequency range 200 MHz to 10 GHz. Series of half wave slots are used in rectangular waveguide to form a leaky RF field across a narrow slot in a conducting plane.

Micro-strip antenna: Recently very popular and can easily be fabricated. It consists of an area (almost any shape) of conductor which is excited on the surface of substrate dielectric having back plane so there is more energy stored in reactive RF region.



(a) (b)

Fig 6.6:(a)Slot antenna b) Micro-strip antenna

Wire type: Simplest form of all radiators. Here wire need to be straight.

Loop antenna: It's a magnetic dipole type antenna (dimension << wavelength). Here a coil loop usually carry RF current. The loop need not be circular. There can be more than one turn.

Dipole type: A straight conductor broken at some point so that most electromagnetic wave propagate into space & not remained confined within the system. There are different types of dipole antenna such as halfwave dipole, infinitesimal dipole, folded dipole etc. (Fig. 6).

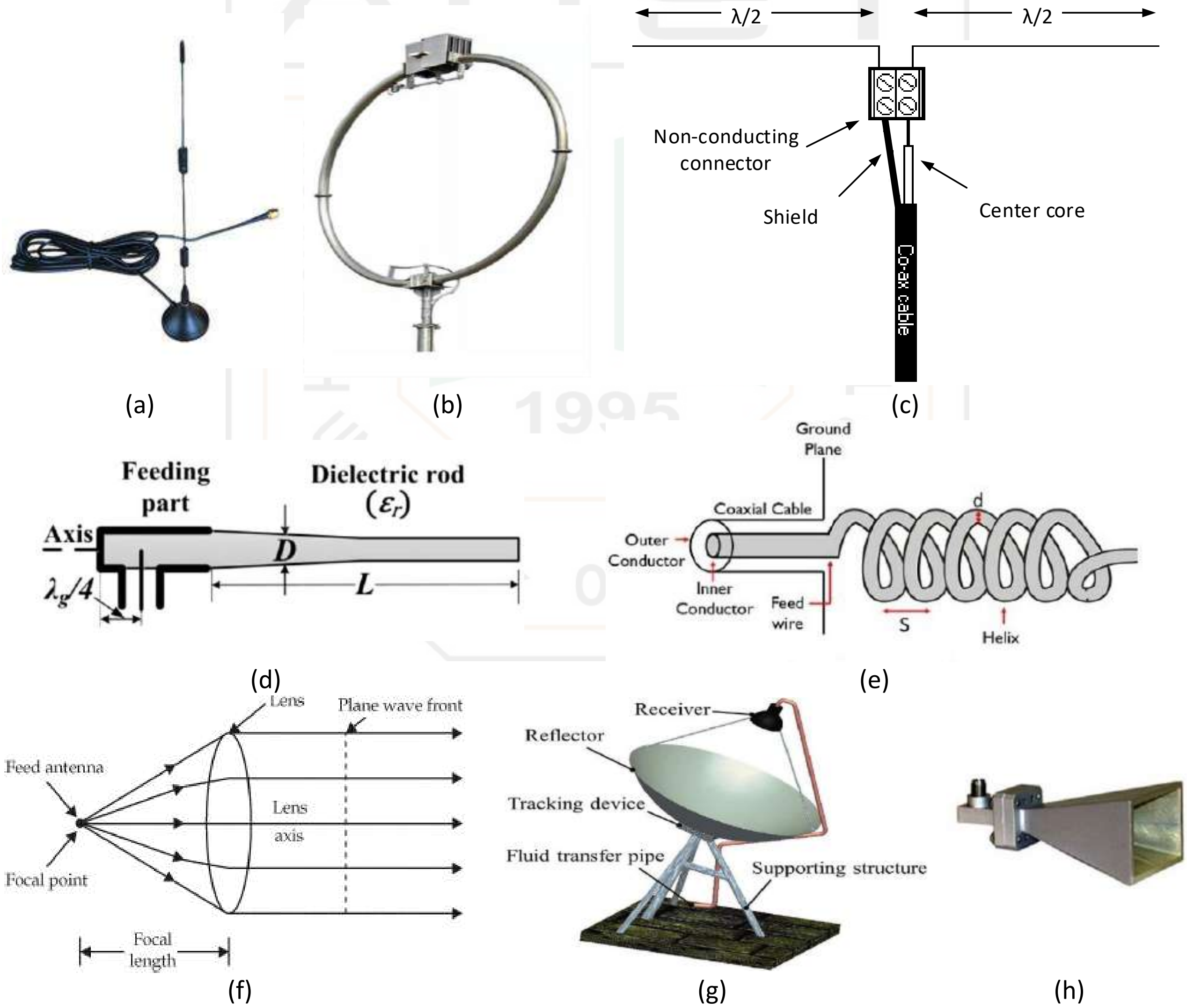


Fig 6.7: (a) Wire type (b) Loop antenna (c) (b)Dipole antenna (d) Dielectric rod antenna (e) Helix antenna (f) Lens antenna (g) Parabolic type (h) Aperture type antenna

Lens antenna: Convert a spherical wave into a plane wave.

Helix antenna: Used to generate circular polarization.

Dielectric rod antenna: It is a dielectric extension to the waveguide.

Parabolic type: Radiation from a short dipole/slot/horn directed towards a large conducing surface which reflects energy. If primary radiation is at infinity, geometric optics predicts parallel beam will emerge from such a reflector.

Aperture type: Radiation from an open area in a conducting surface. It works on the basis of Huygene principle such that any wave front is a source of secondary wave front. (eg. Waveguide horn). Most ground based satellite receiver dish have a smooth horn feed of narrow gain at focus of dish (0.5-1 m). There are different types such as flared H-plane, flared E-plane. If impedance is correct, all energy toward waveguide will radiate.

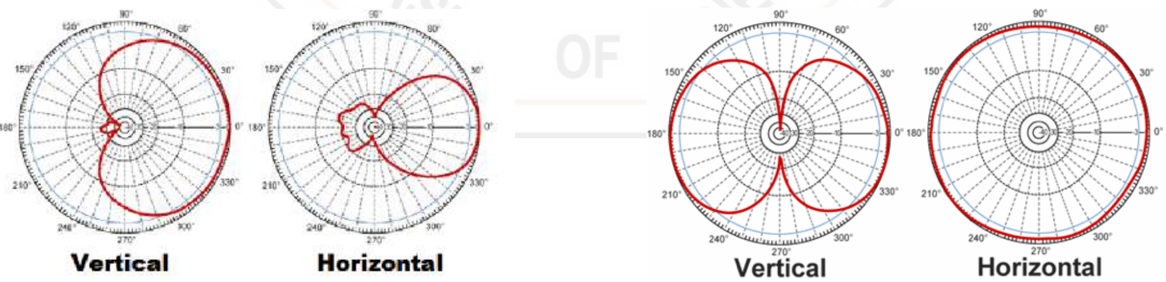
Antenna array: These are formed from multiples of the other kind of antennas. Active arrays have individual element individually driven by its own feed, whereas passive arrays have a primary radiator passing near field energy to parasitic elements. eg. Yagi, Broad side, End fire, Log periodic etc.

Non-Array antenna: Eg dipole, wire, loop, aperture, slot, helix, parabolic etc.

(a) (b) (c)

Figure: (a)Karl G. Jansky Very Large Array (Astronomy) (b) China YJ-26 radar (Military) (c)Millimeter wave phased array antenna from IBM research (Communication)

6.3.3 Antenna classification based on Radiation pattern



(a) (b)

Fig 6.8: (a)Directional antenna (b) Omnidirectional antenna

Directional antenna – This kind of antenna has the rotational variance around vertical axis. Example: Horn, Parabolic, Quad, Slot, Yagi-uda, Helical Dipole (when placed horizontally)

Omnidirectional antenna: This kind of antenna has rotational invariance around vertical axis (radiates uniformly in azimuthal plane. Example: Whip, monopole, Dipole (when placed vertically).

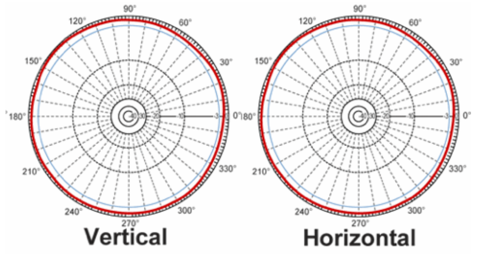


Fig 6.9: Isotropic antenna

Isotropic antenna: It's a kind of antenna which radiates equally in all direction. It is a reference radiator with which other antennas are compared. If the power supplied to this radiator is P watts, then the energy density at a distance R meter is P/(4πR2)

6.3.4 Some important laws and definition

Inverse square law:

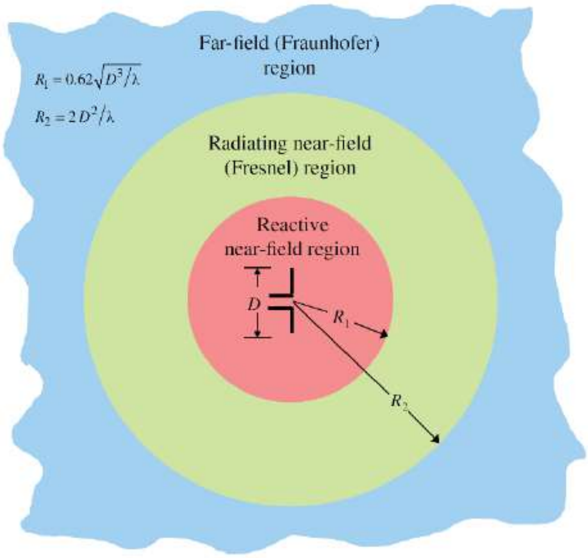
Let the total power radiated from a point source (e.g. antenna) is P_{rad}. At large distance from the source (compared to the size of source), this power is uniformly distributed over larger and larger spherical surfaces as the distance from the source increases. Since the surface are of the sphere of radius 'R' is $A=4\pi r^2$ then power density (which can be measured in dB) at distance 'R' is

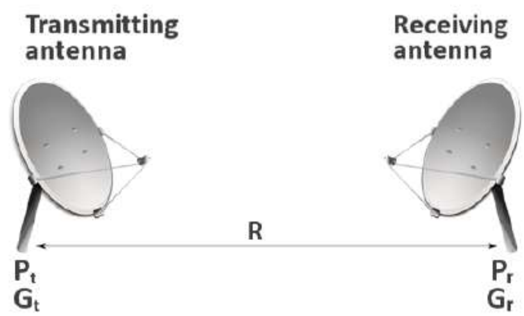
$$W = \frac{P_{rad}}{A} = \frac{P_{rad}}{4\pi R^2}$$

Near field region: In this case, the polar radiation pattern depends on the distance from the antenna and there is reactive power flow in and out of the region. Energy emerging here has an oscillatory longitudinal component and is transferred to and from the near field region.

Far field region: In this case, the polar radiation pattern is completely independent of distance from the antenna.

Rayleigh distance: The transition from near to far field happens at Rayleigh distance (far field distance) which is equal to $2d^2/\lambda$.





(a) (b)

Fig 6.10:(a)Radiation zone of an antenna (b)Transmission and receiving antenna

Friis radiation formula

For the microwave radio link shown in Fig 6.12(b)
P_t, P_r = transmitted & received power
G_t, G_r= transmit & receive antenna gain
R= distance between transmitter & receiver

$$P_r = P_t \; G_t G_r \left(\frac{\lambda}{4\pi R}\right)^2$$

6.4 Experiment procedure:

1. Keep both tripods at a minimal distance of 0.5 m from each other, center to center using measuring tape.
2. The minimal distance ensures that testing antennas are in the far field region.
3. Transmitted RF signal from Yagi (4el) antennas is intercepted by Yagi (3el) and sent to receiver. Measure the level in dBµV.
4. Note down the level reading at 0.5 m distance.
5. Take the reading at 0.7 m distance. Ensure that no scattering objects are in the vicinity of the antenna, which could reradiate and distort the field pattern and consequently the reading. Avoid any movement of persons while taking the readings.
6. Take further readings at 1m, 1.4m, 2m, 2.8 m, 4m, 5.6m, 8 m, and 11.3 m.
7. Readings can be distorted if the Yagi captures signal from its behind due to wall or from the ceiling etc.

6.5 Report:

1. Plot dB reading vs. distance.
2. Relate the results with theory.

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Experiment no: 7
Name of the Experiment: Study of the Characteristics Features of Various Antennas.

7.1 Objective:

1. To understand the concept of radiation pattern of an antenna.
2. To understand the concept of azimuthal plane, elevation plane, boresight direction, beam width etc.
3. To be familiarized with the radiation pattern of Omni-directional, directional, dipole and folded-dipole type antenna.
4. To confirm reciprocity theorem of antenna.
5. To differentiate between resonant and non-resonant antennas, calculate resonance frequency, VSWR and impedance of antenna.

7.2 Equipment list:

1. **Antenna Transmitter** (already discussed in the lab sheet of Expt. 6)

2. **Antenna Receiver** (already discussed in the lab sheet of Expt. 6)

3. **Stepper motor controller** (already discussed in the lab sheet of Expt. 6)

4. **Antenna Tripod** (already discussed in the lab sheet of Expt. 6)

5. **Yagi antenna** (already discussed in the lab sheet of Expt. 6)

6. **Directional & omnidirectional antenna** (already discussed in the lab sheet of Expt. 6)

7. **Dipole antenna:** It is the most common radiating structure which consists of a straight conductor broken at some point where it is excited by a voltage derived from transmission line etc. Dipole antennas that are much smaller than the wavelength of the signal are called Herzian, short or infinitesimal dipoles. They have very low radiation resistance and a high reactance. Dipole antennas which have half the wavelength of the signals are called half-wave dipole (Fig.7.1) A half wave dipole is cut to length according to formula $l(in\ feet) = \frac{468}{f}\ (MHz)$ or $l(in\ meet) = \frac{142.65}{f}\ (MHz)$.

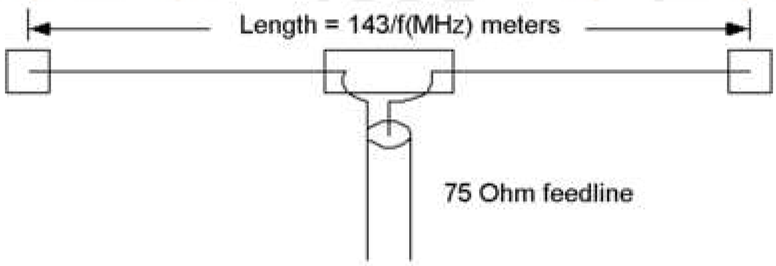


Figure: Half wave dipole Antenna

8. **Folded dipole antenna:** It is a half-wave dipole antenna where an additional parallel wire links the two ends of the half-wave dipole. It presents a driving point impedance of about 300 ohms (4 times of dipole) (Fig.7.2) Here addition of an extra rod raises the impedance because current in the extra rod mirrors the current in the driven rod, both currents being in the same direction. Since it can be considered as two parallel wire transmission line shorted at top and bottom ends where the folds are), the folded dipole presents an open circuit to any differential mode current which might be included in the transmission line considered as a parallel rod line. The current which radiates is twice the current delivered by feeder to the real radiation impedance. Therefore, the same total radiated power halving the feed current must quadruple the radiation resistance.

9. **Polarizer connector**

10. **Resonant antenna:** It corresponds to a transmission line that is exact number of half wavelengths long and is open at both ends. A low impedance source can easily be coupled to it at its low impedance point without causing any disturbance to the standing wave pattern.

11. **Non-resonant antenna**: It's a kind of non-resonant transmission line in which standing wave are suppressed by correct termination so that no power is reflected and only forward wave exist.

12. **Return loss bridge (FALCON):** It has three terminals -antenna, RX, TX. Here, RLB is matched to 75 ohms of receiver and transmitter and its reference impedance. It compares the antenna impedance to 75 ohms. If the antenna is 75 ohm then bridge is balanced and there is no output at the receiver end.

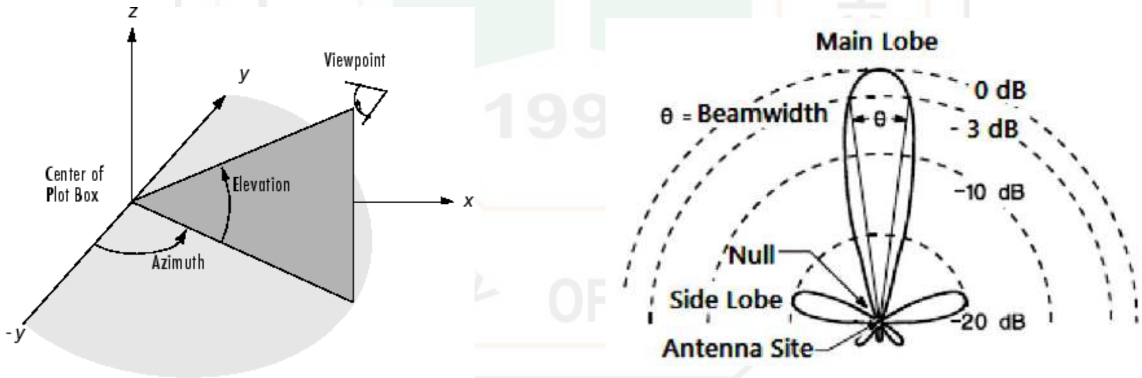
13. **Directional Coupler(AMITECH):** The directional coupler is matched to 50 ohms input impedance of receiver & 50 ohms output impedance of transmitter. It compares the antenna impedance to its internal reference impedance of 50 ohm (which is port 4). If the antenna is 50 ohm, then the bridge is balanced and there is no power at receiver end. The more antenna impedance differs from 50 ohms, the more will be output of receiver. Maximum receiver output will be at antenna impedance shorted or open circuited. For VHF and UHF, the null will be 35 dB and 15 dB respectively.

7.3 Theory

1. Antenna radiation pattern

The relative field strength/ radiation intensity/ power density/ average pointing vector in various direction or the general dependence of directivity on elevation and azimuthal plane is called the radiation pattern. (Fig,7.3). Radiation pattern of omni directional and directional antennas are shown in Fig.7.4. For the latter case the pattern consists of a major lobe in boresight direction and smaller lobes in other direction.

2. Some definitions related to antenna radiation pattern:



(a) (b)

Figure: (a) Azimuthal and Elevation plane (b) Radiation pattern of an antenna

Azimuthal plane: From the radiation point of antenna, if we look around horizontally then we get the azimuthal plane. Angle varies between 0 to +360°.

Elevation plane: From the radiation point of antenna, if we look up and down with respect to local horizon, then we get the elevation plane. Angle varies between -90° to +90°.

Boresight direction: The direction along which the radiation is most highly concentrated.

Beamwidth Angular separation between two half power points on the power density radiation pattern

Bandwidth: The frequency range over which satisfactory operation is possible or the frequency separation between two 3 dB points in q-curve.

Front to back ratio: Difference between the field strength etc. (in dB) in the boresight and in its 180° opposite direction

Antenna Gain: It measures the directive character of a given antenna. It is defined by the ratio of maximum radiation intensity from subject antenna and the radiation intensity from lossless perfect isotropic source / antenna having the same total accepted input power. If the direction is not specified, the value for gain is taken to mean the maximum value in any direction for the particular antenna. Mathematically,

$$G = \frac{4\pi A}{\lambda_0^2}$$

Where, A = capture area (intercept or absorption x- section of received antenna)

In case of horn antenna both at transmit and receive end,

$$G = \frac{4\pi r}{\lambda_0}\sqrt{\frac{P_{rec}}{P_{tr}}}$$

If two identical horn antennas are not available,

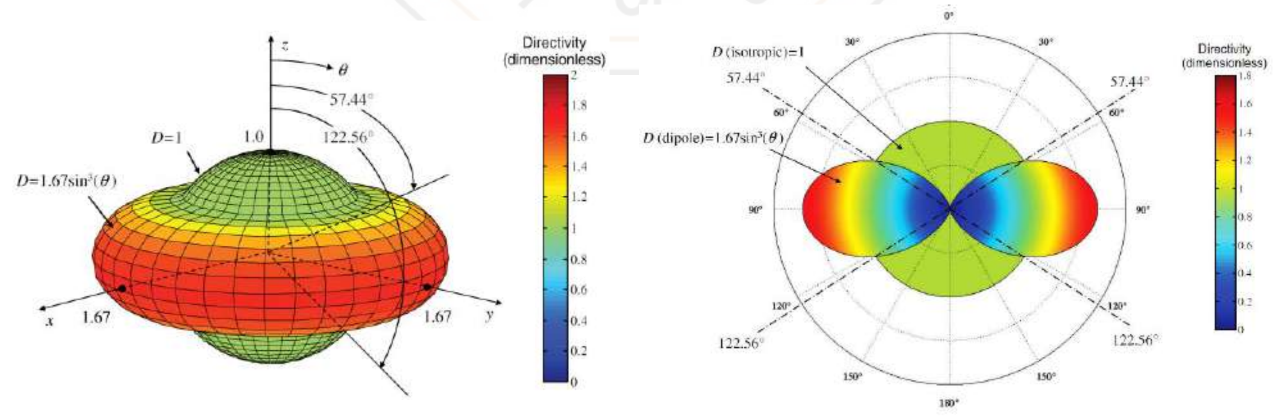
$$G = \frac{13325.71}{\theta_{az}\theta_{el}}$$

Where, θ_{az} and θ_{el} are azimuthal and elevation Beamwidth in degrees respectively.

Directivity: ratio of radiation intensity in any direction by a particular antenna to that by an isotropic antenna. Assume both antenna radiates same amount of power. In another angle, directivity is also defined as the ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all the directions. This averaged intensity is equal to the total power of the antenna divided by 4π. If the direction is not specified, the directivity refers to the direction of maximum radiation intensity. Also the directivity of an isotropic source is unity since its power is radiated equally well in all directions.

$$D = \frac{41000}{\theta_{az}\theta_{el}}$$

Where, θ_{az} and θ_{el} are azimuthal and elevation Beamwidth in degrees respectively.



(a) (b)

Figure: directivity patterns of a λ/2 dipole

It is apparent that when $57.44^\circ < \theta < 122.56^\circ$, the dipole radiator has greater directivity (greater intensity concentration) in those directions than that of an isotropic source. The maximum directivity of the dipole (relative to the isotropic radiator) occurs when $\theta = \pi/2$ and it is 1.67 (or 2.23 dB) more intense than that of the isotropic radiator (with the same radiated power).

Efficiency: The ratio of gain and directivity.

Radiation resistance: The resistive impedance presented by an antenna to the transmission line / free space impedance is called radiation resistance. Energy transferred to antenna & environment by resistive power flow heats up surrounding. Mathematically,

$$Radiation\ resistance = \frac{power\ radiated\ by\ antenna}{(current\ at\ feed\ point)^2}$$

Reciprocity theorem (Fig. 7.5): Consider two separate sources, J1, M1 and J2, M2 which generates fields E1, H1 and E2, H2 respectively in the volume V enclosed by the closed surface S. The equation corresponding to reciprocity theorem is given in the following table for two separate cases

When S encloses no charges	When S bounds a perfect conductor
$\int_s (\vec{E}_1 \times \vec{H}_2) \cdot \overrightarrow{dS} = \int_s (\vec{E}_2 \times \vec{H}_1) \cdot \overrightarrow{dS}$	$\int_v (\vec{E}_1 \cdot \vec{J}_2 - \vec{H}_1 \cdot \vec{M}_2) \cdot dv = \int_v (\vec{E}_2 \cdot \vec{J}_1 - \vec{H}_2 \cdot \vec{M}_1) \cdot dv$

The theorem states that the system response E1 & E2 is not changed when the source and observation points are interchanged.

For antenna, it means that the transmitting and receiving radiation patterns of an antenna are equal.

The physical reason is that the only difference between outgoing and incoming waves lies in the arrow of time. Since the electromagnetic equations are invariant except for the signs of magnetic field and current, under time reversal, there can be the difference between transmit and receive mode in the physical current and field distribution.

Antenna resonance: The tendency of a system to absorb more energy when the frequency of its oscillations matches the system natural frequency of vibration is called resonance. Electrical resonance occurs in an electrical system when the impedance between the input and output of the circuit is at minimum (or when the transfer function is minimum). Resonance of a circuit involving capacitors and inductors occurs when the collapsing magnetic field of the inductor generates an electric current in its windings that charges the capacitor and discharging the capacitor provides an electric current that builds the magnetic field in the inductor and the process is repeated continuously. In some cases, the inductive reactance and capacitive reactance of the circuit are of equal magnitude causing electrical energy to oscillate between the magnetic field of inductor and electrical field of capacitor. If L & C are inductive and capacitance respectively, then $\omega L = \frac{1}{\omega C}$ and $\omega = \frac{1}{\sqrt{LC}}$

$$RL = -20 \log|\Gamma|$$

$$VSWR = \frac{1 + |\Gamma|}{1 - |\Gamma|} = \frac{Z_0}{R_i}$$

7.4 Experiment procedure:

Omni directional antenna (Fig.7.6)

1. Connect a polarization connector to the tripod and a dipole antenna to the polarization connector. Dipole antenna should rest in vertical position.
2. Set transmitting frequency at 600 MHz, attenuator low/high for FALCON/ AMITECH
3. Connect the monopole antenna to the stepper tripod and set the frequency of receiver to 600 MHz and attenuator upward for maximum sensitivity.
4. Set the distance between antennas to be around 1 m (a distance between 1-1.5 m ensures multipath reflection is minimized, otherwise reading will vary, variation higher than 6 dB is not good. Keep the antennas in same horizontal plane & remove any stray object in the LOS)
5. Using stepper tripod, rotate the vertical monopole antenna around its axis & take readings.

Directional antenna

Follow the same procedure as before but, (i) connect dipole antenna directly to transmitter tripod and (ii) connect Yagi antenna in place of monopole antenna in the receiver tripod, (iii) reading should be taken for both azimuthal & elevation plane.

Dipole antenna

Follow the same procedure as before but, connect another dipole antenna in place of yagi antenna.

Folded dipole antenna

Follow the same procedure as before. But, connect a folded dipole antenna in place of dipole antenna at the receiver.

AMITECH approach	FALCON approach
1. Set up the directional coupler for forward power measurement as Fig. 7.7 2. Connect the TX at input port of directional coupler (RF IN, port 1), RX at coupled port (SAMPLE, port 3), antenna at output port (RF OUT, port 3). Take forward power reading. 3. Connect the antenna at input port of directional coupler (RF IN, port 1), RX at coupled port (SAMPLE, port 3), TX at output port (RF OUT, port 2). Take reverse power reading. Difference between the two reading is the return loss. 4. Follow step 3-6 of the FALCON approach.	1. Connect the return loss bridge to the transmitter tripod through the TX RF connector. 2. Connect the dipole antenna to the RLB at the connector, connect the receiver to the RLB at ANT connector. 3. For frequencies set from 500 to 750 MHz, record receiver readings, there will be distinct decrease in level due to bridge null where antenna resonates. Locate that frequency. 4. Change antenna dimensions and locate the new null frequencies. 5. Repeat the same for folded dipole antenna. 6. From the return loss table, calculate antenna impedance, VSWR.

Antenna reciprocity test

1. Connect Yagi to receiver stepper and dipole antenna to transmitter tripod, keep antennas in horizontal plane at 1 meter distance and both are set at frequency 600 MHz.
2. Rotate the Yagi around its axis and take readings.
3. Interchange Yagi and dipole antenna, rotate dipole antenna & take readings

7.5 Report:

1. Plot radiation pattern of different antennas studied during the experiment.
2. Calculate Beamwidth, front to back ratio, side lobe level, directivity and gain of antennas studied.
3. Relate the results with theory.
4. Present other results studied during the experiment, relate them with theory.

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Experiment No: 11
Name of the Experiment: Introduction and calibration of Vector Network Analyzer

(a)Objective:

1) Necessity of transmission line theory in Microwave and millimeter wave devices
2) Challenges of Measurement in microwave frequency.
3) Introduction to Scattering Parameter and its usefulness
4) Introduction to Vector Network Analyzer
5) Understanding Error Model of a 2-port Network.
6) Calibration method and SOLT calibration

(b)Theory

2.1 Lumped element model and Distributed element model (Transmission line Model):

The *lumped element model* simplifies electrical network by considering the electrical network as combination of resistors, capacitors, and inductors, etc. joined by perfectly conducting wires. The model also assumes the energy transfer from source to load is instant.

Practically energy transfer is not instant, but EM field propagate in the circuit at propagation velocity $\left(v_p = \frac{c}{\sqrt{\varepsilon_r}}\right)$. The wavelength λ, is defined as a function of the propagation speed (V_p or c) and the sine wave generator frequency (f_0) in Equation

$$\lambda = \frac{v_p}{f_0}$$

When the frequency (f_0) is low, the wavelength is large, and the length of the cable is negligible compared to the size of the wavelength. As a result, the measured voltage and current are independent of the location on the cable. As the energy transfer in the network is instant the voltage in point 'a' and point 'b' is the same (there is no voltage drop or phase change). This situation is illustrated in Fig 1(a), and the circuit is referred to as being a lumped element circuit or lumped.

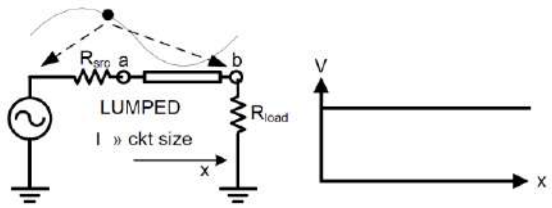


Fig 1(a): Circuit response at low frequency

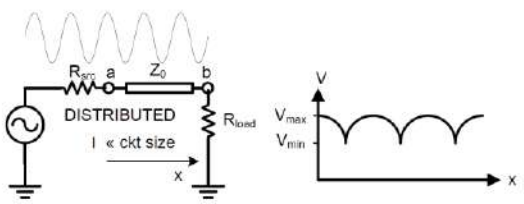


Fig 1(b): Circuit response at high frequency

When the frequency (f_0) of the source increases, the wavelength is reduced. Thus, as frequency increases, the wavelength eventually becomes similar in size or even smaller than the length of the cable. In a scenario where the wavelength of the signal is similar or smaller in size to the length of the cable, the measured voltage and current will depend on the position, as shown in Fig 1(b) Thus, measuring the voltage with a voltage probe is invalid because the result will be dependent on the probe's position. In this scenario, the circuit must be treated as a distributed element circuit rather than as a lumped circuit.

Analysis of a distributed circuit is more complex and involves the use of transmission line theory. In transmission line theory, electrical power traveling along the line can be considered as a voltage (E-field) and current (H-field) traveling and relation is imposed by the electrical properties of the line. the cable itself will behave such that it is characterized by an inherent impedance that does not change as long as the properties of the line or cable do not change. This impedance is called the characteristic impedance (Z).

As the electrical power hits the termination (R_{load}), the voltage to current relationship is now imposed by the impedance of the load. Under the condition where the load impedance is equal to the characteristic impedance, the power is fully absorbed. If the load impedance is different from the characteristic impedance, the ratio of voltage and current will change at the point where the transmission medium occurs. As a result, the load will not absorb all the power, resulting in a portion of the power traveling back towards the source. They are known as incident and reflected wave. The ratio of them is known as reflection coefficient:

$$\Gamma(\omega_0) = \frac{V^+(\omega_0)}{V^-(\omega_0)} = \frac{Z_L - Z_0}{Z_L + Z_0}$$

At each point along transmission line, there are two waves traveling. They are a) forward wave moving towards load b) reflected wave moving generator/signal source keeping a variable phase relationship between them. So, there are some points where both waves are in phase & in some they are in antiphase. Corresponding voltages in those points will be Vmax and Vmin respectively. Ratio of Vmax & Vmin is called VSWR

2.2 Scattering parameter/ S-parameter:

The S-parameters describe the magnitude and phase relationship between incident and reflected waves. They relate to familiar measurements such as gain, loss, and reflection coefficient. They are defined in terms of voltage traveling waves, which are relatively easy to measure. S-parameters don't require connection of undesirable loads to the device under test. The measured S-parameters of multiple devices can be cascaded to predict overall system performance. If desired, H, Y, or Z-parameters can be derived from S-parameters. And very important for RF design, S-parameters are easily imported and used for circuit simulations in electronic-design automation (EDA) tools. S-parameters are the shared language between simulation and measurement.

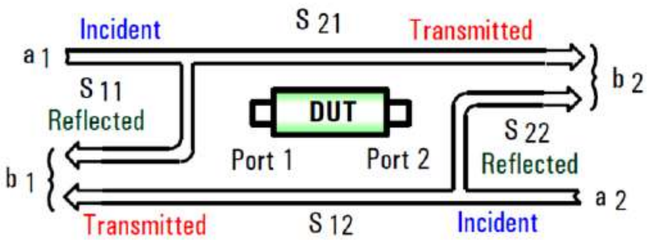


Fig 2: Signal Path in an ideal two port network

For a two-port network shown in Fig. 2 a1 and a2 are the incident wave in port 1 and port 2, b1 and b2 are reflected/ transmitted wave in port 1 and port 2 respectively. Because the system is *linear* mathematically it can be written as:

$$b_1 = S_{11}a_1 + S_{12}a_2$$

$$b_2 = S_{21}a_1 + S_{22}a_2$$

when the characteristic impedance (Z_0) is equal to 50 Ohms, and if a 50 Ohm termination is present at port 2 (in fig 3 a_2 is reduced to zero) resulting in equations for S_{11} and S_{21}. This principle can be applied in the reverse direction as well. By setting a_1 to zero, equations for S_{22} and S_{12} can be obtained.

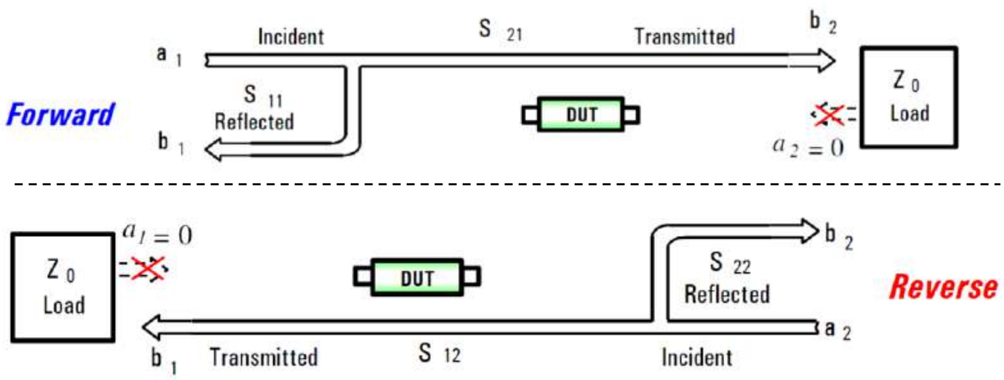


Fig 3: Definition of S parameter from forward and reverse transmission

Defining incident wave as $a_1 = V_1^+$ and $a_2 = V_2^+$ and reflected/transmitted waves as $b_1 = V_1^-$ and $b_2 = V_2^-$. So for forward transmission the s parameters are

$$S_{11} = \frac{Wave\ reflected\ from\ port\ 1}{Wave\ Incidnet\ \ at\ port\ 1} = \frac{V_1^-}{V_1^+} = \frac{b_1}{a_1}\bigg|_{a_2=0}$$

$$S_{21} = \frac{Wave\ transmitted\ \ to\ port\ 2}{Wave\ Incidnet\ at\ port\ 1} = \frac{V_2^-}{V_1^+} = \frac{b_2}{a_1}\bigg|_{a_2=0}$$

And for reverse transmission the s parameters

$$S_{22} = \frac{Wave\ reflected\ from\ port\ 2}{Incidnet\ wave\ at\ port\ 2} = \frac{V_1^-}{V_2^+} = \frac{b_2}{a_2}\bigg|_{a_1=0}$$

$$S_{12} = \frac{reflected\ wave\ from\ port\ 1}{Incidnet\ wave\ at\ port\ 2} = \frac{V_1^-}{V_2^+} = \frac{b_1}{a_2}\bigg|_{a_1=0}$$

The numbering convention for S-parameters is that the first number following the "S" is the port where the signal emerges, and the second number is the port where the signal is applied. So, S_{21} is a measure of the signal coming out port 2 relative to the RF stimulus entering port 1. When the numbers are the same (e.g., S_{11}), it indicates a reflection measurement, as the input and output ports are the same. The incident terms (a1, a2) and output terms (b1, b2) represent voltage traveling waves. An N-port device has N^2 S-parameters. So, a two-port device has four S-parameters. For 1 port network the scattering matrix is $[S_{11}]$. Similarly for 2 port network the matrix is $\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix}$ and so on.

2.3 Relationship between S parameter and other quantity

	Port 1	**Port 2**
S_{xx}	Input port match, $\Gamma_{in} = S_{11}$	Output port match, $\Gamma_{out} = S_{22}$
	Return loss at the input port, $\mathrm{RL}_{in} = -20\log_{10}\lvert S_{11}\rvert\ \ (\mathrm{dB})$	Return loss at the output port, $(\mathrm{RL})_{out} = -20\log_{10}\lvert S_{22}\rvert\ \ (dB)$
	$(VSWR)_{in} = \dfrac{1+\lvert S_{11}\rvert}{1-\lvert S_{11}\rvert} = \dfrac{1+\lvert\Gamma_{in}\rvert}{1-\lvert\Gamma_{in}\rvert}$	$(VSWR)_{out} = \dfrac{1+\lvert S_{22}\rvert}{1-\lvert S_{22}\rvert} = \dfrac{1+\lvert\Gamma_{out}\rvert}{1-\lvert\Gamma_{out}\rvert}$
S_{xy}	$Scalar\ logarithmic\ gain, g = Insertion\ loss, IL = -20\log_{10}\lvert S_{21}\rvert\ \ (dB)$	$Scalar\ logarithmic\ reverse\ gain, g_{rev} = -20\log_{10}\lvert S_{12}\rvert\ \ (dB)$
		$Reverse\ Isolation, I_{rev} = \lvert g_{rev}\rvert = \lvert -20\log_{10}\lvert S_{12}\rvert\rvert$

2.4 Vector Network Analyzer (VNA):

Network Analyzer is an instrument used to measure impedance. At lower frequencies impedance can be measured with a sine wave generator, a volt meter, a current meter. The ratio of voltage and current will give out the impedance of the network. At higher frequency, the voltage and current vary with position due to the standing wave produced by the interaction of transmitted and reflected wave (explained in the previous section). Thus, impedance measurement at higher frequencies is done with measurement of incident and reflected waves. In fact, the VNA is able to measure the amplitude and phase differences between incident and reflected waves, using one of the waves as a reference.

The primary use of a VNA is to determine the S-parameters of numerous passive components, including cables, filters, switches, diplexers, duplexers, triplexers, couplers, bridges, transformers, power splitters, combiners, circulators, isolators, attenuators, antennas, and many more. In addition, VNAs can also characterize active devices such as transistors and amplifiers using S-parameters, as long as they are operating in their linear mode of operation.

2.5 Architecture of a 2-port Vector Network Analyzer:

The fundamental principle of a vector network analyzer is to measure the amplitude and phase of both incident and reflected waves at the various ports of the DUT. The general design of a VNA is to stimulate an RF network at a given port with a stepped or swept continuous wave (CW) signal and to measure the travelling waves, not only at the stimulus port but at all the ports of the network terminated with specific load impedances, typically 50 Ohms or 75 Ohms. A typical but simplified VNA architecture is illustrated in Fig 4.

A typical VNA have one or more signal source (SRC) with controllable frequency. The test port contains some signal separation hardware (e.g. directional coupler) to split out the incident and reflected travelling waves. The test set can contain switches to route the signal source to the different test ports and terminates other ports with specific load impedances. The ‘ref’ port typically measure the incident wave from the coupled line and the test port measure the reflected signal.

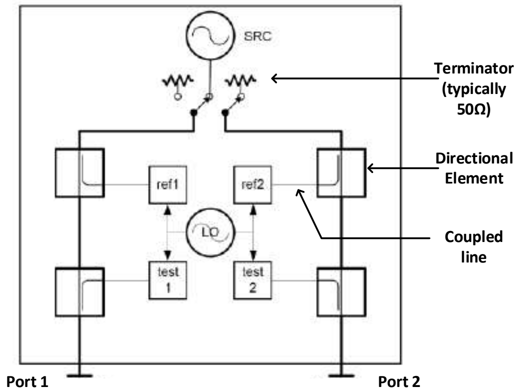
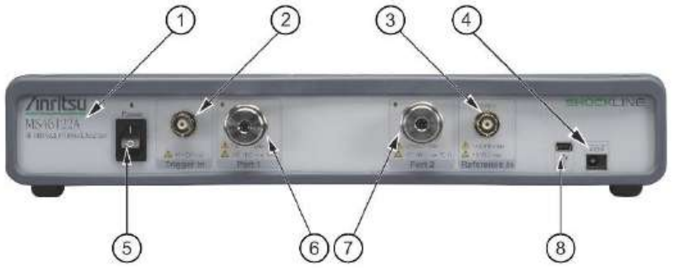


Fig 4: Simplified VNA block diagram



1. ID Plate – Instrument model number identification	5. Operate / Standby Power Switch
2. Trigger In	6. Test Port 1 – K(m)
3. 10 MHz External Reference Input	7. Test Port 2 – K(m)
4. DC Power Supply Connector	8. USB Port

Fig 5: Anritsu MS43122A-20 Vector Network Analyzer

2.6 VNA Errors and Calibration:

In the VNA measurement there are three types of errors: random, systematic and drift. Random errors vary with time and are thus unpredictable and cannot be removed by calibration. Typical random errors include those caused by instrument noise, and the repeatability of switches, cables, and connectors. In contrast, systematic errors occur in a reproducible manner. They are caused by imperfections in the VNA, can be characterized, and thus can be removed mathematically through calibration. Drift occurs after a calibration has been performed because of changes in VNA performance arising from variations in ambient temperature.

A VNA is only as useful as the accuracy with which it makes measurements, and this requires the instrument to be calibrated. The calibration process employs a technique called *vector error correction.* Vector-error correction is the process of characterizing systematic error terms by measuring known calibration standards, and then removing the effects of these errors from subsequent measurements mathematically. The process of removing these errors requires the errors and measured quantities to be measured vectorically (Magnitude and Phase).

Each network analyzer can be separated into an *Error Network* (or linear error model) and an *ideal network* analyzer. The parameters of the error network are considered 'error terms' and can be directly interpreted as raw system data. Correcting system errors is the primary goal of calibration, and any remaining errors are expressed by the effective system data and depend on the accuracy of the error terms and the repeatability of the measurement process.

The calibration process determines the *error terms*, requires a test system consisting of a VNA, cables and a *calibration standard*. These calibration standards are one-port and two-port networks that have known characteristics. It is impossible to manufacture a calibration standard that has ideal properties, so the deviations of the standards are sent to the VNA as ‘characteristic data’. This data is provided as data files to the VNA software. After calibration, the analyzer mathematically computes the error terms using the values it measured during the calibration process along with the characteristic data of the standards. It is then possible to correct the raw measured values in later measurements and calculate systematic error free S-parameters for the device under test.

2.7 Error Model:

Fig 6 show simplified error model of the *Error network.* There are three class of systematic error in an Error network. Signal leakage, Signal reflection and Frequency response. The errors relating to signal leakage are directivity and crosstalk. Errors related to signal reflections are source and load match. The errors related to frequency response of the receivers are reflection and transmission tracking. The full two-port error model includes all six of these terms for the forward direction and the same six (with different data) in the reverse direction, for a total of twelve error terms. This is why two-port calibration (3 receiver VNA architecture) is referred as twelve term error correction.

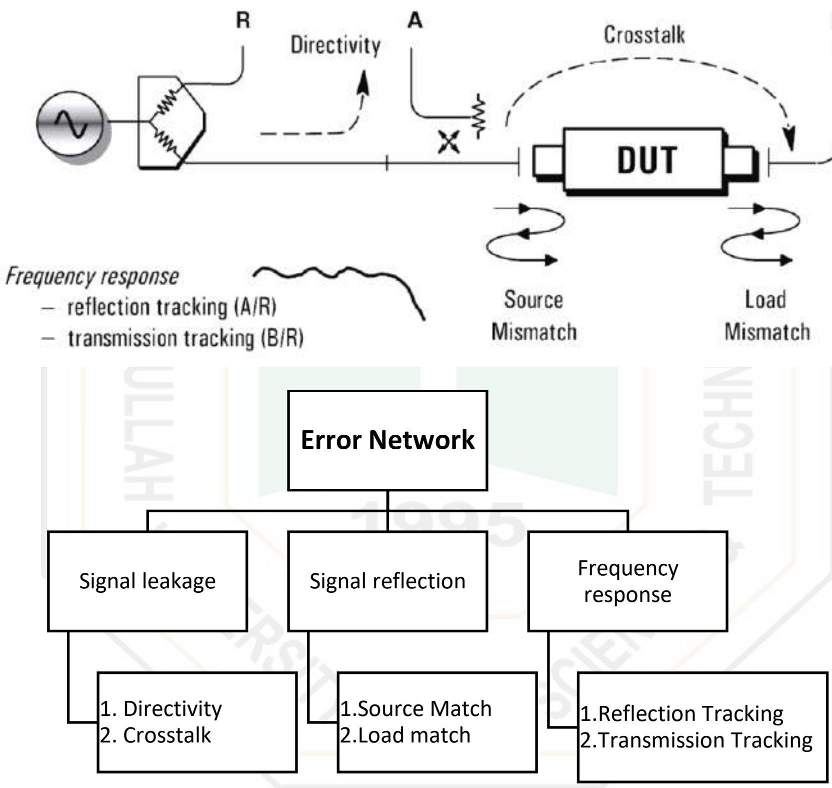


Fig 6: Classic Two-Port 12 term Error Model of VNA

The interface between the error network and the device under test is called the *reference plane*. When using coaxial calibration standards, the reference plane is the mating plane of the outer conductor, an example of which is shown in Fig 7.

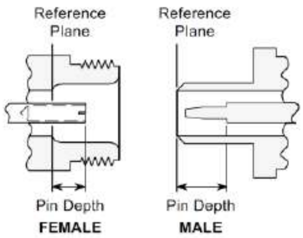
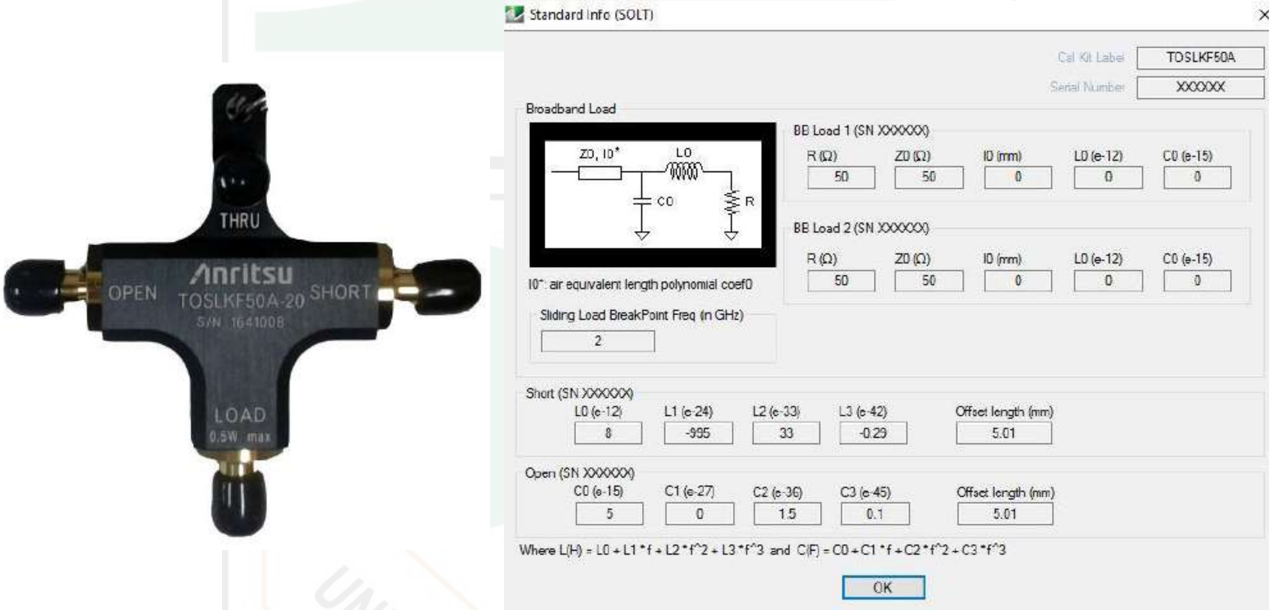


Fig 7: Location of the reference plane in a Type-N connector

2.8 Calibration standard and SOLT Calibration:

There are many different types of VNA calibration methods like SOLT, SSLT, SSST, LRL, ALRM and so on. For our purpose, we will use the most common calibration technique called SOLT (**S**hort, **O**pen **L**oad, **T**hrough). SOLT calibration technique use four calibration standards (short, open, load and through) to acquire all 12 term which is required for full two port calibration (see Appendix A for detail). The characteristics data of the SOLT standard can be seen in figure 8 with the actual cal kit and the data sheet of the kit.



TOSLKF50A-20 Calibration Kit Specifications

Through (Thru)	Spec
Length	16.07 mm
Return Loss (DC to 10 GHz)	≥ 34 dB
Return Loss (10 to 20 GHz)	≥ 32 dB
Insertion Loss (DC to 20 GHz)	≤ 0.025 x √(f/GHz) dB

Open	Spec
Length	5.01 mm
C0 (1E-15) F	5.000
C1 (1E-27) F/Hz	0.000
C2 (1E-36) F/Hz^2	1.500
C3 (1E-45) F/Hz^3	0.100
Phase (DC to 10 GHz)	≤ ± 1.5°
Phase (10 to 20 GHz)	≤ ± 3.0°

Short	Spec
Length	5.01 mm
L0 (1E-12) H	8.000
L1 (1E-24) H/Hz	−995.000
L2 (1E-33) H/Hz^2	33.000
L3 (1E-42) H/Hz^3	−0.290
Phase (DC to 10 GHz)	≤ ± 1.5°
Phase (10 to 20 GHz)	≤ ± 2.5°

Load	Spec
DC Resistance	50 Ω ± 0.25 Ω
Return Loss (DC to 10 GHz)	≥ 42 dB
Return Loss (10 to 20 GHz)	≥ 36 dB
Max Power	0.5 W

Fig 8: SOLT Calibration kit, characteristic data in ShockLine Software and data sheet of TOSLKF50A-20 coaxial calibration kit from Anritsu

Short: A coaxial short (Fig 9) can be constructed that has near ideal characteristics, with a total reflection of magnitude 1. The reflection coefficient of the short is dependent only on its length offset, which represents the length between the 'reference plane' and the short. The loss occurring over this length can generally be ignored. Modeling the short in a VNA requires that only its electrical length be entered into the instrument, but in some cases the model can be extended using the polynomial coefficients L_0 to L_3 to account for parasitic inductance.

$$L = L_0 + L_1 f + L_2 f^2 + L_3 f^3$$

Fig 9: Internal construction and equivalent circuit diagram of the short standard

Open: A coaxial open standard (Fig 10) is constructed using a closed design to avoid effects caused by entry of stray electromagnetic energy. At the open end of the inner conductor, a frequency-dependent fringing capacitance is formed. Even if an open standard could physically be constructed with a length of 0, fringing capacitances would result in a negative imaginary part for S11at higher frequencies.

$$C = C_0 + C_1 f + C_2 f^2 + C_3 f^3$$

Fig 10: Internal construction and equivalent circuit diagram of the open standard

Match/ Load: A match is a precision broadband impedance that has a value corresponding to the system impedance (50 Ohm in this case). An implementation in which the inner conductor terminates into a resistively-coated substrate is shown in Fig 11.

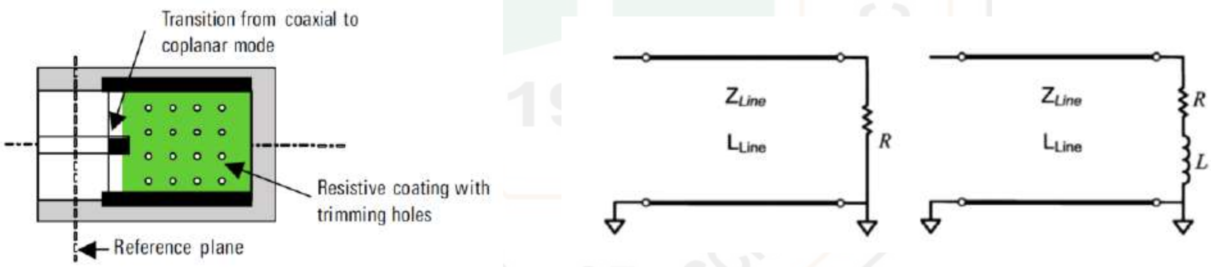


Fig 11: Internal construction and equivalent circuit diagram of the load standard

Through: A through (Fig 12) is a two-port standard that allows direct connection of two test ports with low loss. The characteristic quantities of a through are insertion loss and electrical insertion length. the through is assumed to be ideally matched. If connectors of the same type but of a different gender are used, the two test ports can be directly connected to produce a through connection. This special case of a through has an electrical insertion length of 0 (zero) mm. Through standard is modeled as a transmission line length with some frequency dependent loss. A root-f frequency dependence of that loss is assumed. If 0 is entered for f_0 (the reference frequency), the loss is assumed to be constant with frequency.

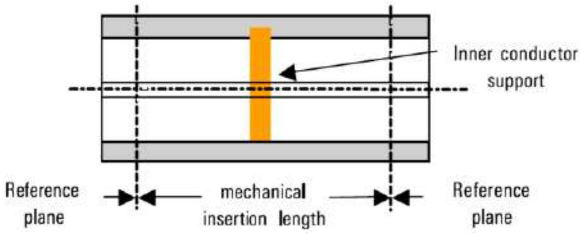


Fig 12: Thru standard

(c)Experimental Procedure:

1. Connect both the wire to Test Port 1 and Test port 2 (see Fig 5).
2. Turn on the power supply.
3. Click to open "ShockLine" software. ShockLine is a software made by Anritsu which can be used as the interface of the MS46122A-20 Vector Network Analyzer.

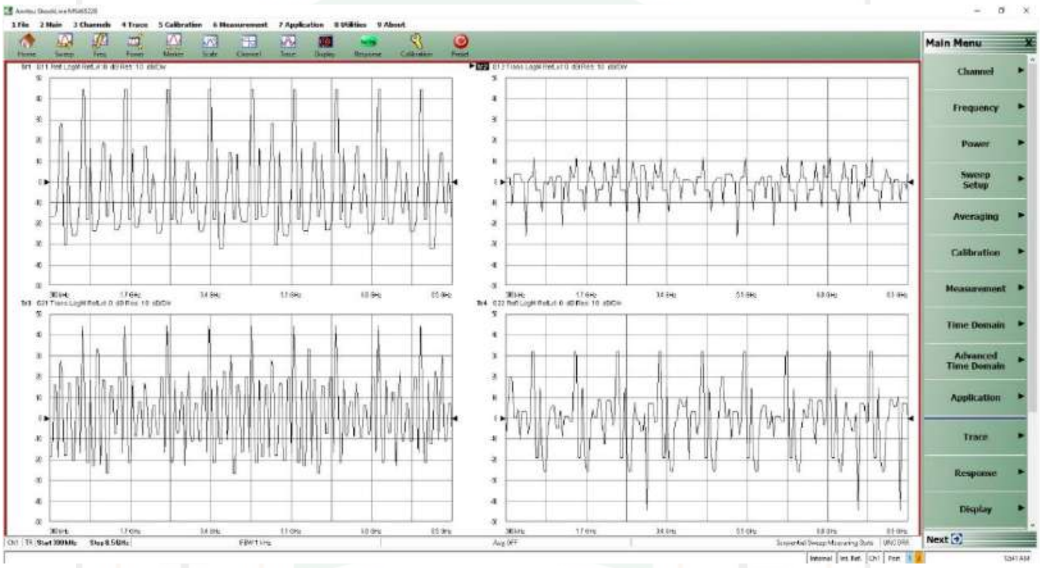


Fig 13: Anritsu ShockLine software interface

4. The default interface should look like Fig 13.
5. Click 'Preset'. This will change all the setting of the VNA to its default setting. In the status bar (bottom of the screen) the Start Stop frequency of the VNA, frequency, IF bandwidth, Measurement status can be seen at a glance. As the VNA is in default status the calibration is void. (can be seen as "UNCORR" in the status bar)
6. To change the start and stop frequency of the sweep, click frequency on the right side of the screen or Click 'Main>Frequency' from the main menu.

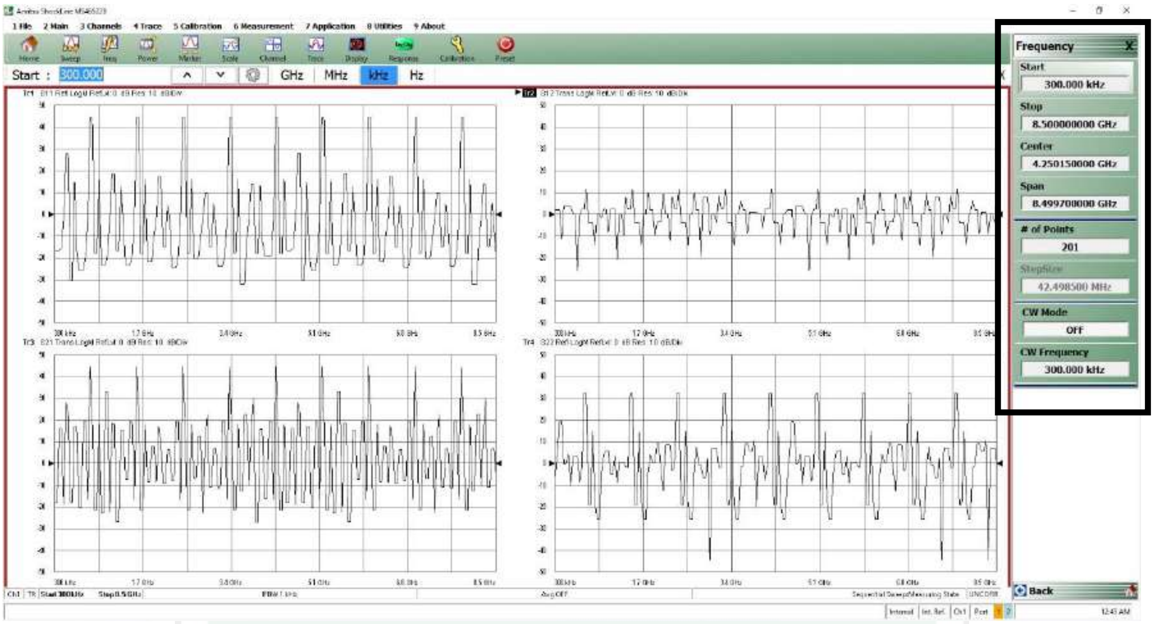


Fig 14: Start and Stop frequency selection

7. Enter the Start and Stop frequency in the dialog box (Fig 14), also enter the 'number of points'. By increasing the number of points the resolution of the measurement can be increased but it will also slow down the sweeping speed.
8. To start calibration, click 'calibrate' from the main menu. In the calibration menu click Calibrate>Manual Cal> 2 Port Cal (Fig 15). In 'Two Port Cal' menu port 1, port 2 is calibrated one at a time using the Open, Short, Load standard. Then Thru calibration is done.

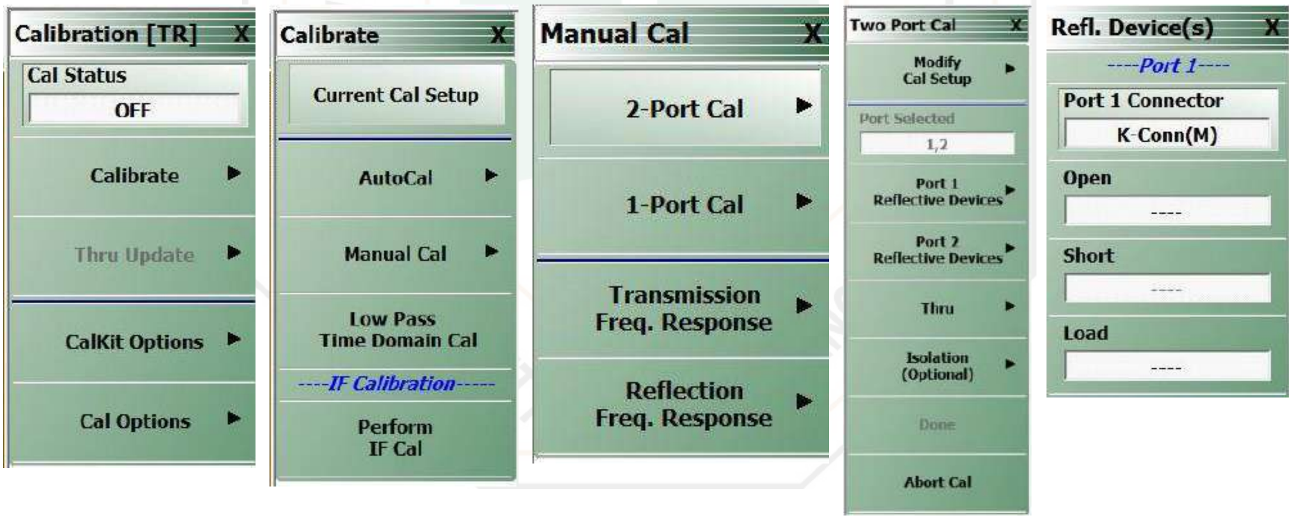


Fig 15: Calibration Sub menu

9. To calibrate port 1, click 'Port 1 Reflective Devices'.
10. In the Refl. Device(s) sub-menu click 'port 1 Connector'
11. Set all the parameter according to the Fig 16. Select "TOSLKF50A" as the Cal kit for test port 1 and test port 2. Click Ok.

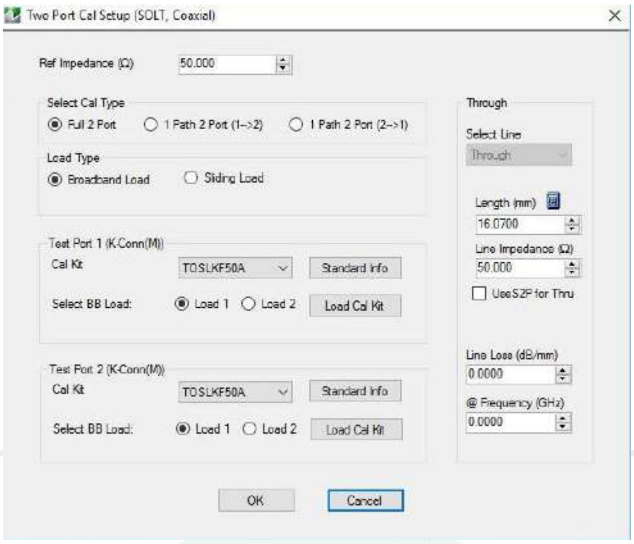


Fig 16: Calibration process

12. Now connect the Open standard at the end of the cable and click 'open' in the 'refl. Device(s)' menu. After the calibration, there will a small check mark on the side of open menu.
13. Then disconnect the open standard and connect the short standard. Click 'short'. After the calibration, there will a small check mark on the side of Short menu.
14. Then disconnect the short standard and connect the Load standard. Click 'Load'. After the calibration, there will a small check mark on the side of Load menu.
15. If all three standards are calibrated successfully the menu will look like Fig 17.
16. Now for port 2 calibration, repeat steps 12-14 for port 2.
17. For thru calibration connect thru standard in-between port 1 and in port 2. Click thru in 'two port cal' Menu. After the calibration, there will be a check mark on side of the thru menu.
18. If all is done right after the Thru calibration there will be a dialog box saying "Please click "done" button to complete the calibration". Click ok.
19. Click Done in the 'Two port cal' Menu (fig 17)

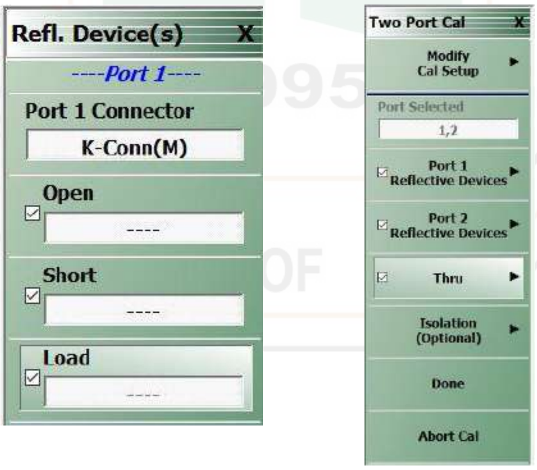


Fig 17: Menu after complete two port calibration

20. If the calibration is done the Status bar will show "CORR" as the Measuring State.

Fig 18: Software interface after Calibration

21. Verify all the calibration during or after the calibration by connecting the short, open and load to port 1 and 2. View the impedance in smith chart format and compart with fig 19.

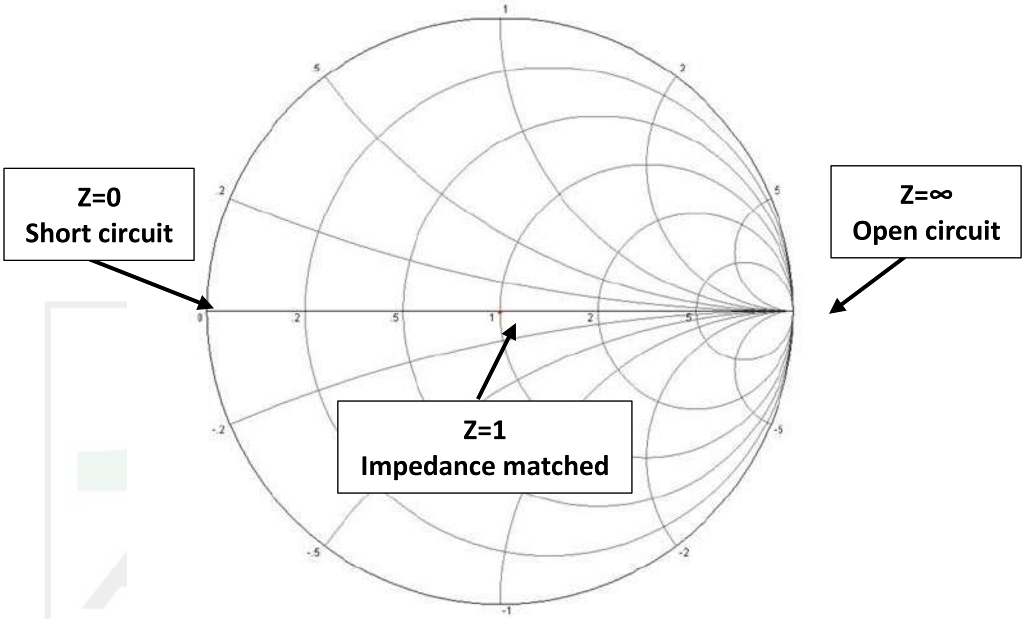


Fig 19: Ideal Position for Open, Load and short circuit in Smith Chart

22. To view the measurement in smith chart format, connect load/short/open in port 1 or 2 then click Display>Trace Format> Smith (R+jX) >Impedance.

(d)Report:

1. What is a VNA?
2. What type of connector does the VNA and cable use? Why?
3. Explain in brief the Error model of a VNA.
4. Why it is absolutely essential to calibrate a VNA before using it for measurement?
5. Ideally the trace in Smith chart should follow the r and x circle but why the trace in VNA spiral inwards?

Experiment no: 12
Name of the Experiment: **Study on Transmission line matching circuit using vector network analyzer**

12.1 Objective:

1. To understand the theory of impedance matching.
2. To understand the design of impedance matching networks by using analytical method (transmission line theory)
3. To understand the design of impedance matching networks by using graphical method (Smith Chart)

12.2 Equipment:

1. Vector network analyzer
2. ETEK MSA-2003-01 Module

12.3 Theory:

Impedance matching is very important with radio and microwave transmission lines. Otherwise, standing waves lead to increased losses and corresponding transmitter malfunction. A line terminated in its characteristic impedance has a standing-wave ratio of unity and transmits a given power without reflection. Also, transmission efficiency is optimum where there is no reflected power.

Matching a transmission line has a special meaning, one differing from that used in circuit theory to indicate equal impedance seen looking both directions from a given terminal pair for maximum power transfer. In circuit theory, maximum power transfer requires the load impedance to be equal to the complex conjugate of the generator. This condition is sometimes referred to as a conjugate matching. In transmission line problems matching means simply terminating the line in its characteristic impedance.

Impedance matching or tuning is important for the following reasons:

1. Maximum power is delivered when the load is matched to the line (assuming the generator is matched), and power loss in the feed line is minimized.
2. Impedance matching sensitive receiver components (antenna, low-noise amplifier, etc.) may improve the signal-to-noise ratio of the system.
3. Impedance matching in a power distribution network (such as an antenna array feed network) may reduce amplitude and phase errors.

Three methods for Impedance matching are:

1. Matching with Lumped Elements (L Networks)
2. Stub Tuning
3. The Quarter-Wave Transformer

The equation of reflection coefficient, load impedance and characteristic impedance are written as

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0} \tag{1.1}$$

$$Return\ Loss\ = -20\log|\Gamma|\ \text{dB} \tag{1.2}$$

For lossy transmission line we know: $\gamma = \alpha + j\beta$; γ=propagation constant, α=attenuation constant, β=phase constant. For lossless case $\alpha = 0$. The input impedance of a transmission line as shown in Fig. 12.1 terminated at a load Z_L is:

$$Z_{in} = Z_0 \left[\frac{Z_L + Z_0\, j\tan(\beta l)}{Z_0 + Z_L\, j\tan(\beta l)}\right] \tag{1.3}$$

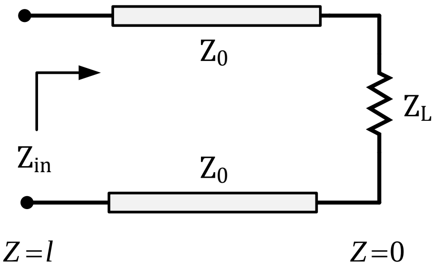


Fig 12.1: input impedance of a transmission line terminated in a load

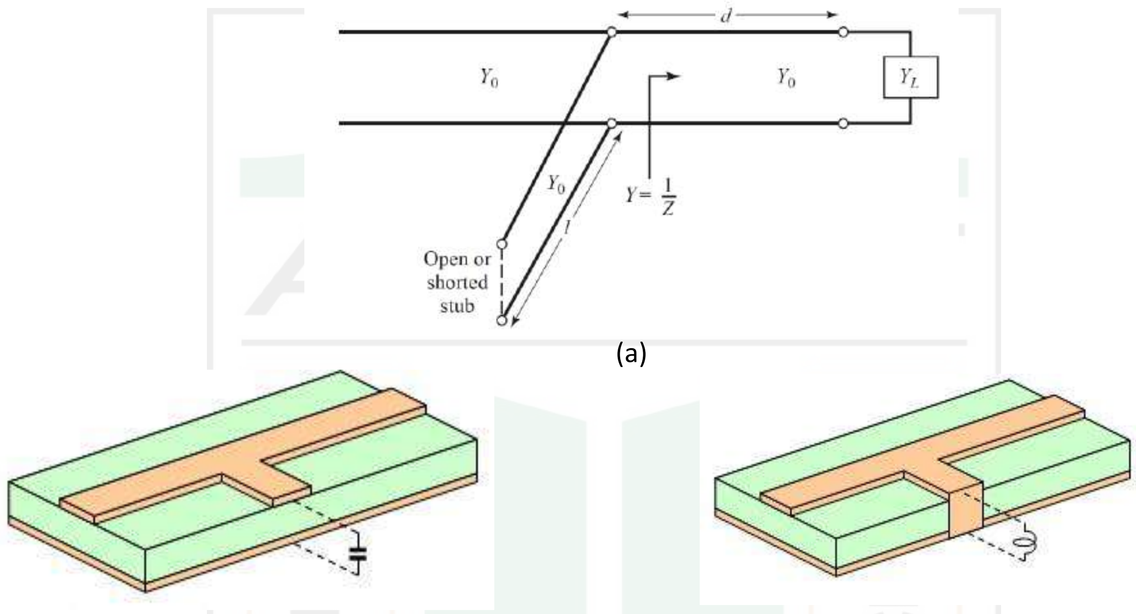
Where,
Z_0 =The characteristic impedance of transmission line
Z_L =The load impedance
l = the distance to load (Length of transmission line between source and load)
β =the wave number $(2\pi/\lambda)$
Let us consider several special cases from equation (1.3):

Table 12.1: Transmission line

Sl.	Conditions	Mathematical Relation	Remarks
1.	$l = \lambda/4$ (Quarter wave transformer)	$Z_{in} = Z_0 \left[\frac{Z_L + Z_0\, j \tan\left[\frac{2\pi}{\lambda} * \frac{\lambda}{4}\right]}{Z_0 + Z_L\, j \tan\left[\frac{2\pi}{\lambda} * \frac{\lambda}{4}\right]}\right]$ $\Rightarrow Z_0^2 = Z_{in} Z_L$ $\Rightarrow Z_0 = \sqrt{Z_{in} Z_L}$	If the input impedance Z_{in} and load impedance Z_L are given, we can design a $\lambda/4$ transmission line with characteristic impedance $\sqrt{Z_{in} Z_L}$ to match between Z_{in} and Z_L. Thus the $\lambda/4$ transmission line is also called as $\lambda/4$ impedance transformer.
2.	$Z_L = 0$ where $\frac{(n-1)\lambda}{4} \leq l \leq \frac{(2n-1)\lambda}{2}$ $n = 1,2,3 \ldots$	$Z_{sc} = Z_{in}\vert_{Z_L=0}$ $= jZ_0 \tan \beta l$ $\cong j\omega L, When\ l < \lambda/4$	A short-circuited terminated transmission line act as an inductive device (we can also see this while calibrating VNA with a short). Thus, we can use it to substitute a parallel inductive device. Moreover, the trace will move in counter clock-wise on the Smith-Chart.
3.	$Z_L = 0$ and $l = \lambda/4$ Short circuit condition	$Z_{sc} = jZ_0 \tan \beta l$ $= jZ_0 \tan\left(\frac{2\pi}{\lambda} * \frac{\lambda}{4}\right)$ $= \infty$	The $\lambda/4$ short-circuited terminated transmission line can be seen as an open circuit. In implementation, the higher the characteristic impedance of transmission line is, the better the result we can obtain.
4.	$Z_L = \infty$ where $\frac{(2n-1)\lambda}{4} \leq l \leq \frac{n\lambda}{2}$ $n = 1,2,3 \ldots$	$Z_{oc} = \lim_{Z_L \to \infty} Z_{in}$ $= \frac{Z_0}{j \tan(\beta l)}$ $\cong \frac{1}{j\omega C}, When\ l < \frac{\lambda}{4}$	An open-circuited terminated transmission line can be seen as a capacitance device (we can also see this while calibrating VNA with an open). Thus, we can use it to substitute a parallel capacitance device. Moreover, the trace will move in clock-wise on the Smith-Chart.

5.	$Z_L = \infty$ and $l = \lambda/4$ Open circuit condition	$Z_{oc} = \lim_{Z_L \to \infty} Z_{in}$ $= -jZ_0 \cot\left(\frac{2\pi}{\lambda} * \frac{\lambda}{4}\right)$ $= 0$	The λ/4 open-circuited terminated transmission line act as a short circuit. In implementation, the lower the characteristic impedance of transmission line is, the better the result we can obtain.

<u>12.3.1 Stub Matching</u>: This is a method of impedance matching where a small section of short/open circuited transmission line is connected in shunt Fig 12.2(a) or series with main transmission line. The distance or position of the stub d from the load and length of the stub l_s (short stub) or l_o(open stub) are chosen such that the reflected wave produced by shunting impedance or stub is equal and opposite to reflected wave already existing on the line.



(a)

(b) (c)

Figure 12.2: (a) Single-stub tuning circuits (b) Open circuited stub (c)Short circuited stub

<u>Equation related to Single stub match:</u>

Let a load $Z_L = R + jX_L$ connected to a transmission line. To impedance match the load to a impedance of Z_0 the distance d of the stub from the load will be:

$$\frac{d}{\lambda} = \begin{cases} \frac{1}{2\pi}\tan^{-1} t & for\ t \geq 0 \\ \frac{1}{2\pi}(\pi + \tan^{-1} t) & for\ t < 0 \end{cases}$$

Here $t = -\frac{X_L}{2Z_0}$

Also the length of the stub is:

For an open-circuited stub $\frac{l_0}{\lambda} = -\frac{1}{2\pi}\tan^{-1}\left(\frac{B}{Y_0}\right)$ (1.4)

For a short-circuited stub $\frac{l_s}{\lambda} = \frac{1}{2\pi}\tan^{-1}\left(\frac{Y_0}{B}\right)$ (1.5)

Here,

$B =$ Susceptence of the transmission line which can be derived as,

$$B = \frac{R_L^2 t - (Z_0 - X_L t)(X_L + Z_0 t)}{Z_0[R_L^2 + (X_L + Z_0 t)^2]}$$

and $Y_0 = Admittance$ and $Z_0 = \frac{1}{Y_0}$

From here is quite clear that the stub matching calculation from analytical formula is quite complex and cumbersome. This can also be done using Smith Chart which is relatively easier**.**

12.4 Procedure:

Case 1: Quarter wave (λ/4) impedance transformer

1. Refer to the circuit diagram of λ/4 impedance transformer in Fig. 12.3 or Figure MSA 1-1 in ETEK MSA-2003-01 trainer module.
2. Let the load impedance $Z_L = 150\Omega$ to be matched to $Z_{in} = 50\Omega$. Calculate Z_0.
3. Connect the DUT to the VNA and see the Return loss in Log magnitude format.
4. Using the Marker function of the Network Analyzer mark the frequencies at 2350MHz, 2400MHz and 2450MHz respectively.
5. Also view the S_{11} in Smith Chart format.
6. Record all the measured results in graph or table.
7. Based on the S_{11} plot determine the f of the matching DUT and hence calculate λ/4. Try to measure the length from SMA port of the DUT to the load.

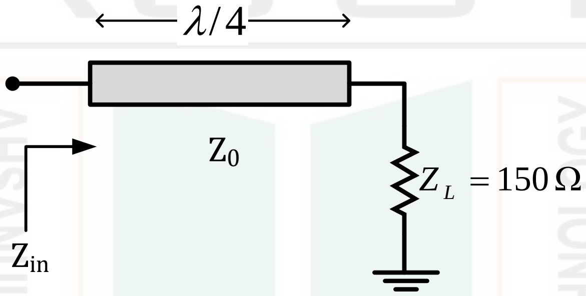


Fig. 12.3: λ/4 impedance-transformer

Case 2: Single Short stub shunt tuning

1. Refer to the circuit diagram of single-port short stub in Fig. 12.4 or Figure MSA 1-2 in ETEK MSA-2003-01 trainer module.
2. Let the load impedance $Z_L = 150\Omega$ to be matched to $Z_{in} = 50\Omega$.
3. Connect the DUT to the VNA and see the Return loss in Log magnitude format.
4. Using the Marker function of the Network Analyzer mark the frequencies at 2350MHz, 2400MHz and 2450MHz respectively.
5. Also view the S_{11} in Smith Chart format.
6. Record all the measured results in graph or table.
7. Based on the S_{11} plot determine the f at which the DUT is matched. Try to measure the length of d and l of the setup.

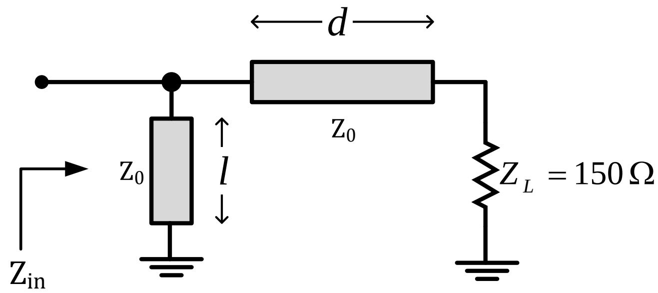


Fig 12.4: Diagram for single short stub shunt matching network

<u>Case 3: Single open stub shunt tuning</u>

1. Refer to the circuit diagram of single-port short stub in Fig. 12.5 or Figure MSA1-2 of ETEK MSA-2003-01 module.
2. Let the load impedance $Z_L = 150\Omega$ to be matched to $Z_{in} = 50\Omega$.
3. Connect the DUT to the VNA and see the Return loss in Log magnitude format.
4. Using the Marker function of the Network Analyzer mark the frequencies at 2350MHz, 2400MHz and 2450MHz respectively.
5. Also view the S_{11} in Smith Chart format.
6. Record all the measured results in graph or table.
7. Based on the S_{11} plot determine the f at which the DUT is matched. Try to measure the length of d and l of the setup.

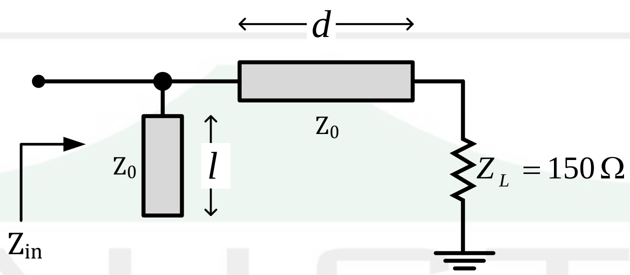


Fig. 12.5: Diagram for single open stub shunt matching network

12.5 Report:

1. Discuss and explain different results obtained in the experiment.
2. Explain when transmission line act as a capacitor or inductor.
3. What are some of the drawbacks of impedance matching using quarter wave transformer technique (Case 1)?
4. Using Smith chart calculate the distance of the stub from load and stub length for both short stub (Case 2) and open stub (Case 3).

Experiment number: 13
Experiment Name: Design and simulation of a Microstrip Patch Antenna (MPA)

13.1 Objective:

1. Understand the theoretical concept of Microstrip patch antenna (MPA)
2. Understand the design aspects of MPA
3. Understand the design steps of MPA
4. Simulate MPA using CST studio software
5. Understand the characteristics of MPA using CST studio software

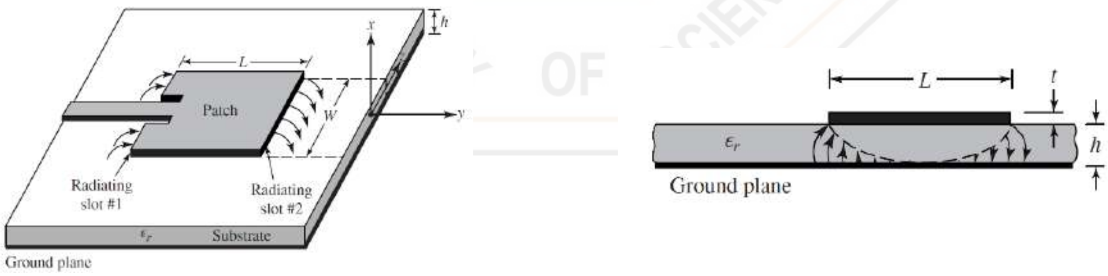
13.2 Theory

13.2.1 Introduction

Microstrip Patch antennas (MPA) are low profile, low cost, low weight, robust, printed circuit manufacturable antennas and are compatible with MMIC design. These high-performance antennas have tremendous prospect in aircraft, space craft, satellite, missile, present and future wireless technology (such as LTE, WiMax, 3G, 4G, 5G, 6G) applications. They can be designed by sophisticated software simulation technologies for different frequency, polarization, radiation characteristics so that they can be integrated into wide bandwidth high speed complicated communication network in a flexible and comfortable manner. Major disadvantages are low efficiency, low power and low Q factor. The concept of MPA was initiated during 1953 and received much attention from 1970.

13.2.2 Structure of MPA:

As shown in Fig. 13.1, the antenna structure consists of dielectric substrate on ground plane, microstrip metallic patch and feeding line. Electric energy received by feed line is converted in electromagnetic energy and is radiated through thin metallic patch (which can have different shapes shown in Fig. 13.2). Thickness, width and length of patch are denoted as t, w, L and substrate height as ‘h’ (small fraction of wavelength, $0.003\lambda << h << 0.05\,\lambda$) with dielectric constant between $2.2 < \varepsilon_r < 12$. Thick substrates with lower dielectric constant are required for better efficiency, larger bandwidth. Thin substrates with higher dielectric constant are suggested for undesired coupling minimization. This antenna is a broad side radiator (maximum radiation perpendicular to patch). End fire radiation can be also be possible. For rectangular patch, L is to be in the range $\lambda/3 < L < \lambda/2$.



(a) (b)

Fig. 13.1: Microwave Patch antenna (a)Top View (b) Side view

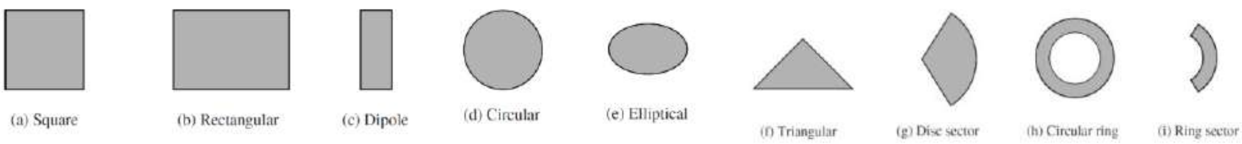


Fig. 13.2: Different type of patch design

13.2.3 Feeding Technology:

There are many feeding techniques developed so far (Fig. 13.3a). Such as microstrip line feed, probe feed, aperture coupled feed, proximity feed etc. Corresponding equivalent circuits are shown in Fig. 13.3b

Microstrip line feed is easy to fabricate, simple to match by controlling inset position and simple to model. But there is bandwidth constraint for higher thickness substrate, but for probe feed, inner conductor is connected with patch and outer conductor to ground plane. Its characteristic more or less similar to that of probe feed but it is difficult to model Aperture coupled feed take care of the cross-polarization problem of previous one by using non contacting aperture. But it is difficult to fabricate. Have narrow bandwidth. On the bottom side of lower substrate, there is a microstrip feed line whose energy is coupled to a patch through a slot on the ground plane separating the two substrates. It allows independent optimization of the feed mechanism. Typically matching is done by controlling the width of the feed line and length of slot. Finally, proximity coupled feeding provides the best bandwidth, it is easy to model but difficult to fabricate. The length of feeding stub and width to line ration of patch is used to control the match.

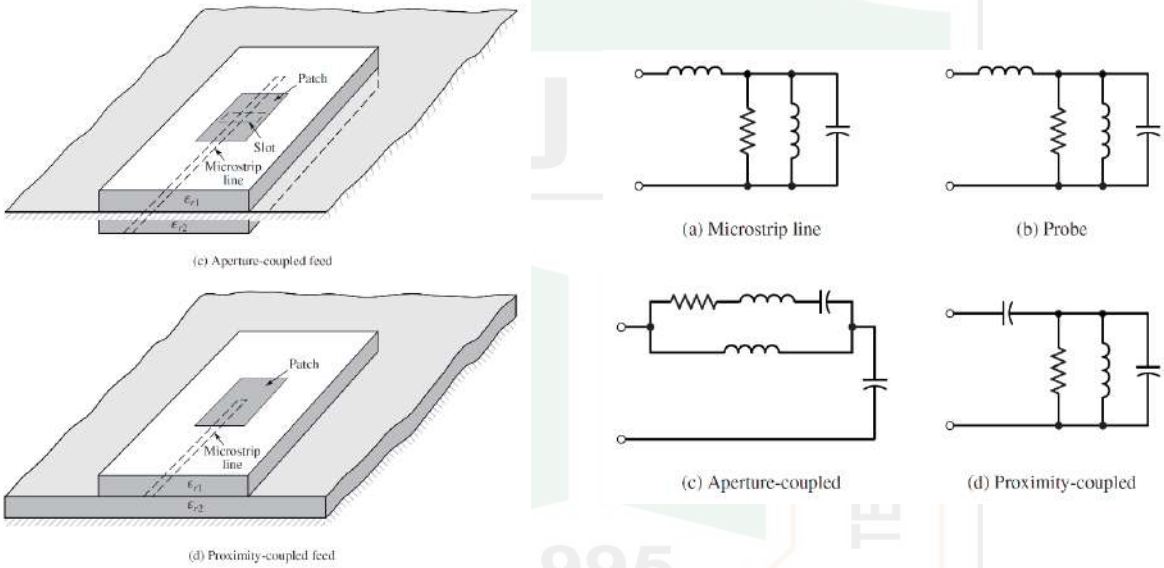


Fig. 13.3: (a)Microstrip antenna feeding technique (b)Their circuit equivalent

13.2.4 Fringing effects:

It is be mentioned that dimensions of the patch are finite along length and width. So the field at the edge of patch undergoes fringing (Fig. 13.4)

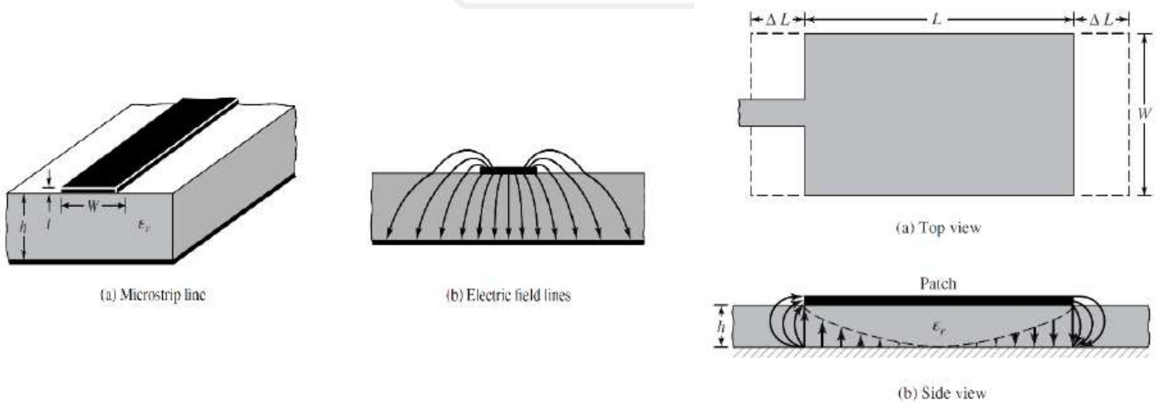


Fig. 13.4: Fringing effect of a MPA

13.2.5 Important Design equations:

For low frequencies the effective dielectric constant is essentially constant. At intermediate frequencies its values begin to monotonically increase and eventually approach the values of the dielectric constant of the substrate. The initial values (at low frequencies) of the effective dielectric constant are referred to as the static values, and they are given by

For $W/h \geq 1$

$$\varepsilon_{re} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left(1 + 12\frac{h}{w}\right)^{-0.5} \tag{13.1}$$

Because of the fringing effects, electrically the patch of the microstrip antenna looks greater than its physical dimensions. For the principal E-plane (xy-plane), this is demonstrated in Fig. 13.4 where the dimensions of the patch along its length have been extended on each end by a distance ΔL, which is a function of the effective dielectric constant ε_{eff} and the width-to-height ratio (W/h). A very popular and practical approximate relation for the normalized extension of the length is

$$\frac{\Delta L}{h} = 0.412\frac{\left(\varepsilon_{eff} + 0.3\right)\left(\frac{W}{h} + 0.264\right)}{\left(\varepsilon_{eff} - 0.258\right)\left(\frac{W}{h} + 0.8\right)} \tag{13.2}$$

Since the length of the patch has been extended by ΔL on each side, the effective length of the patch is now (L = $\lambda/2$ for dominant TM_{010} mode with no fringing)

$$L_{eff} = L + 2\Delta L \tag{13.3}$$

For the dominant TM010 mode, the resonant frequency of the microstrip antenna is a function of its length. Usually, it is given by

$$(f_r)_{010} = \frac{v_0}{2L\sqrt{\varepsilon_r}} \tag{13.4}$$

where v_0 is the speed of light in free space. Since (eq. 13.4) does not account for fringing, it must be modified to include edge effects and should be computed using

$$(f_{rc})_{010} = q\frac{v_0}{2L\sqrt{\varepsilon_r}} \tag{13.5}$$

Were,

$$q = \frac{(f_{rc})_{010}}{(f_r)_{010}} \tag{13.6}$$

13.2.6 MPA Design steps

Based on the simplified formulation that has been described, a design procedure is outlined for practical designs of rectangular microstrip antennas. The procedure assumes specified information regarding the dielectric constant of the substrate (ε_r), the resonant frequency (f_r), and the height of the substrate (h). Then determine W, L

Specify:

$$\varepsilon_r, f_r (in\ Hz) and\ h$$

Determine:

$$W, L$$

13.2.7 Design procedure:

1. For an efficient radiator, a practical width that leads to good radiation efficiencies is

$$W = \frac{v_0}{2f_r}\sqrt{\frac{2}{\varepsilon_r + 1}} \tag{13.7}$$

where v_0 is the free-space velocity of light.

2. Determine the effective dielectric constant of the microstrip antenna using equation (13.1)
3. Once W is found, determine the extension of the length ΔL using equation (13.3).
4. The actual length of the patch can now be determined by solving equation (13.5) for L, or

$$L = \frac{1}{2f_r\sqrt{\varepsilon_{eff}}\sqrt{\varepsilon_0\mu_0}} - 2\Delta L \tag{13.8}$$

13.3 Patch Antenna Simulation Using CST

13.3.1 CST Software features and Application Area:

CST Studio Suite is a software package for electromagnetic and multiphysics simulation used in leading technology and engineering companies around the world. CST has specialized solvers for applications such as motors, circuit boards, cable harnesses and filters. CST also has support for multithreaded CPU, Cluster computing, GPU and hardware acceleration for complex problems. Some of the application area and industry covered by CST are:

1. Aerospace and Defense
   Installed antenna performance, Lightning strike and environmental electromagnetic effects (E3), Radar, Co-site interference
2. Construction, Cities and Territories
   Building shielding, Cabling, Lightning protection
3. Energy and Materials
   High-voltage components, Generators and motors, Solar panel optimization, Transformers
4. Industrial Equipment
   RFID, Non-destructive testing (NDT), Motors and actuators, Welding and lithography
5. Life Sciences
   MRI, Implant safety, Wearable devices, RF diathermy, X-ray tubes
6. High Tech
   Antenna performance, Microwave and RF components, Electromagnetic compatibility (EMC), Signal and power integrity (SI/PI), Touchscreens, Cables and connectors, Specific absorption rate (SAR) exposure
7. Transportation and Mobility
   Antenna installed performance, Cable harness, Automotive radar, Electric motors, Wireless charging, Onboard electronics, Sensors

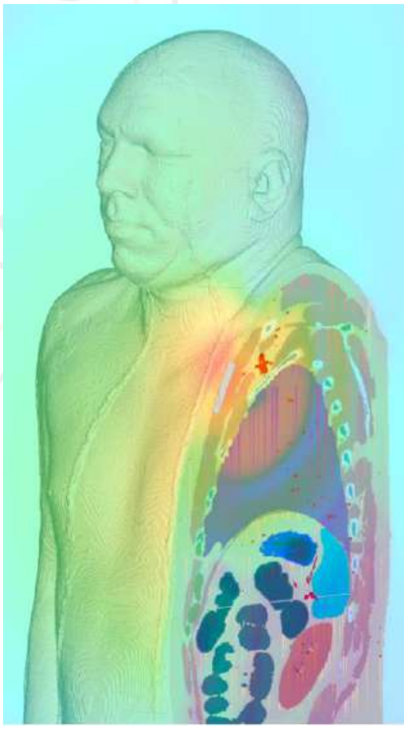
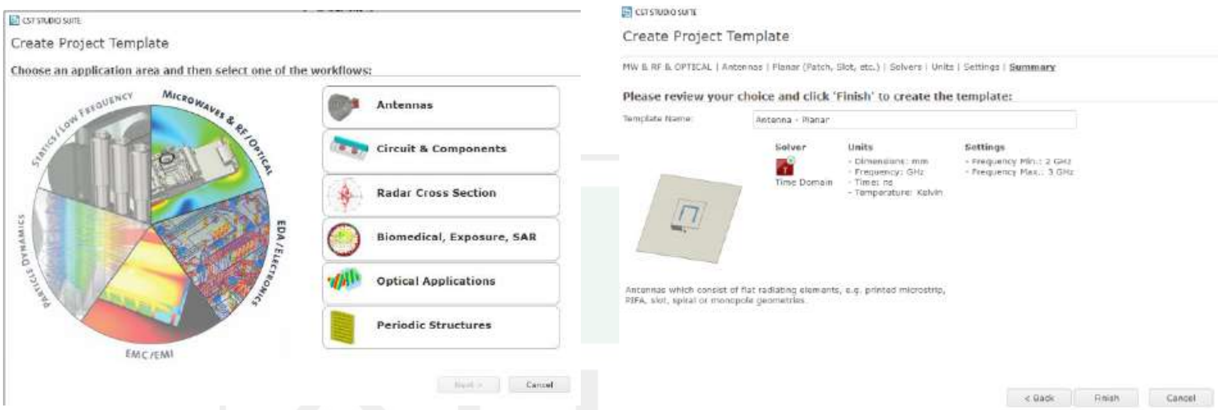


Fig. 13.5: Electric field from a pacemaker antenna inside the human body. (Life Science field) (Courtesy: CST Brochure)

13.3.2 Modeling the MPA

Let the dielectric constant for FR4 material $\varepsilon_r = 4.3$,Hight of the copper layer, Ch=.035mm and Hight of the FR4, h=1.6mm. Using the parameter mentioned above calculate the L and W of the 2.4GHz patch antenna. Also using microwave line calculator, it was found width of a 50ohm microstrip line is, Fw=3.12mm.

1. Open CST Studio suite
2. Click 'New Template'. A window called 'Create Project Template' will appear (Fig. 13.6a).



(a) (b)

Fig. 13.6: (a)New template wizard (b)Summery of the wizard

3. In the 'Choose an application area and then select one of the workflows' section click 'Microwave &RF/Optical' then click 'Antennas'. Click 'Next'.
4. In 'Please select a workflow' section click 'Planar (Patch, Slot, etc.)'. Click 'Next'.
5. In 'The recommended solvers for the selected workflow are:" section select 'Time Domain'. Click 'Next'.
6. In 'Please select the units:' section keep the default setting. Click 'Next'.
7. In 'Please select the Settings' section enter '2GHz' as 'Frequency Min.:' and '3GHz' as 'Frequency Max.:' . As we want to create an antenna at 2.4GHz that is why we have chosen this parameter. We can change this later on. Click 'Next'.
8. Finally, a window will appear that will show the summary of the setting (Fig. 13.6b). Click 'Finish'.
9. The main Window of the CST will appear (Fig. 13.7).

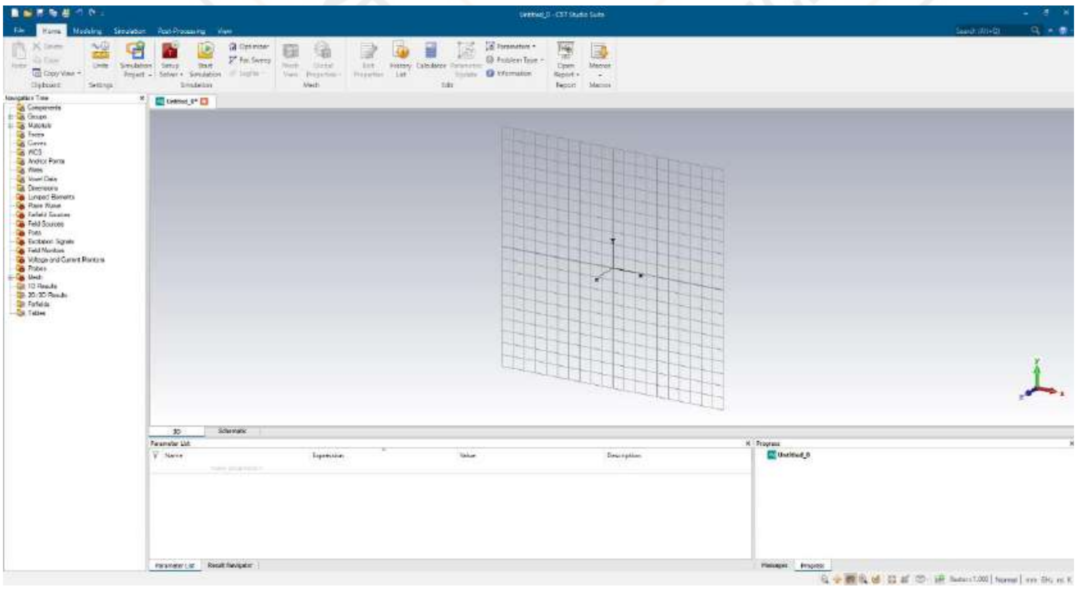


Fig. 13.7: Main interface of CST studio suite

10. On the bottom part of the Main interface of CST, create and enter the following variable in the "Parameter List" section as shown in Table 13.1:

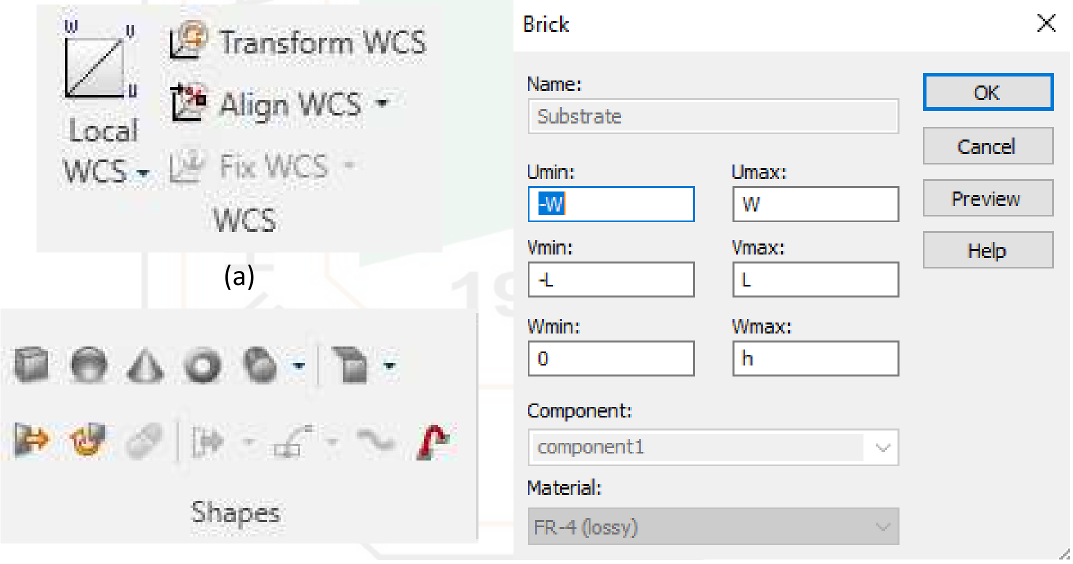
Table 13.1: Parameter for Patch antenna

Name	Expression	Description
W	38	Width of the patch
L	29.5	Length of the patch
h	1.6	Hight of the substrate
Ch	0.035	Hight of the copper layer
Fw	3.12	50ohm Feedline width

11. On the Top ribbon click "Modeling" tab (Fig. 13.8).

Fig. 13.8: Modeling Ribbon

12. Click 'Local WCS' From the WCS group (Fig. 13.9a). We can see the axis change from Global x,y,z to local U,V,W axis.
13. Click 'Brick' from Shapes group (Fig. 13.9b). Click on the main window and press Esc in keyboard. A window will appear (Fig. 13.9c). Enter parameter as shown Table 13.2 for substrate. Select FR-4 (lossy) as the material. Click 'Ok'. This will be the main substrate (FR4) board.



(a) (b) (c)

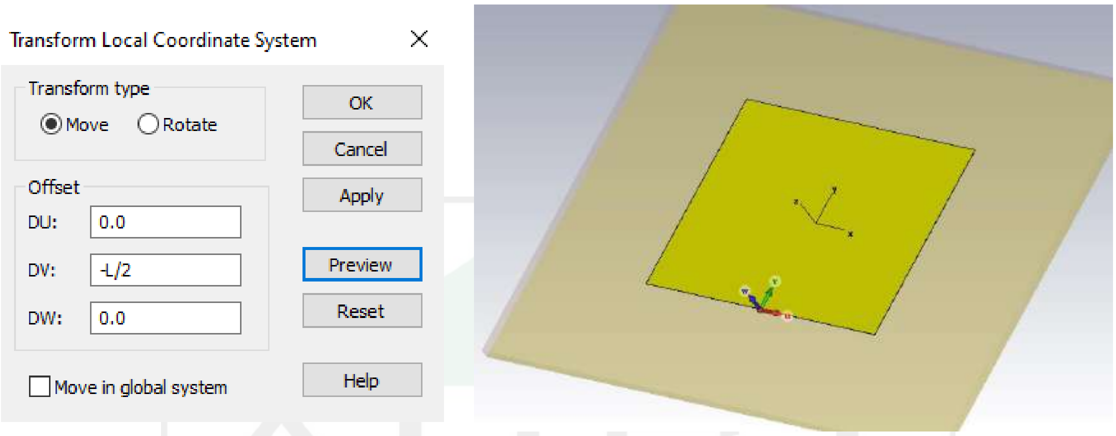
Fig. 13.9: (a)WCS group (b)Shapes Group (c)Settings for Brick

14. Repeat the previous step and create the Ground copper layer.

Table 13.2: Parameter for Patch antenna

Name	Umin	Umax	Vmin	Vmax	Wmin	Wmax	Material
Substrate	-W	W	-L	L	0	h	FR4(lossy)
Ground	-W	W	-L	L	0	-Ch	Copper (annealed)
Patch	-W/2	W/2	-L/2	L/2	0	Ch	Copper (annealed)
Feedline	-Fw/2	Fw/2	0	-L/2	0	Ch	Copper (annealed)

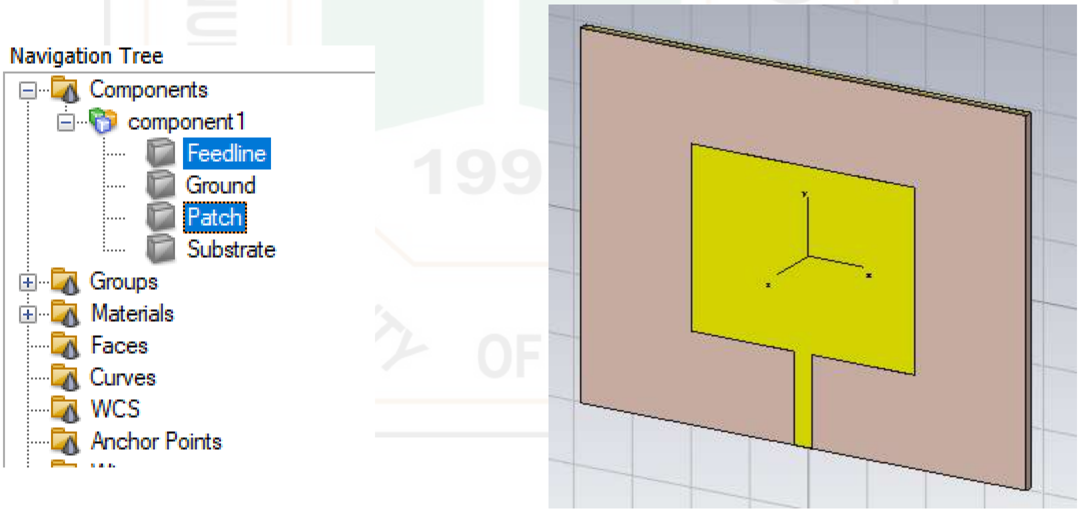
15. Now click 'Pick Face' from Picks group. Double click on the top of the substrate. Top Face of the substrate will be highlighted. From WCS group click Align WCS>Align WCS with selected face. This will align the Local WCS reference frame with the top face.
16. Now create a Brick and create the patch according to the Table 13.2.
17. Now from WCS group, click Transform WCS (Fig. 13.10a). Put 'DV=-L/2'.Click Ok. This will move the WCS axis in V axis by -L/2 distance and should look like Fig 13.10b.



(a) (b)

Fig. 13.10: (a) WCS Transform setting (b) WCS frame after transformation

18. Now click on the Brick and create the Feedline using the values mentioned in the Table 13.2.
19. Now Select Feedline and Patch both from the Navigation Tree> Components> Component1 (Fig. 13.11a).
20. From Modeling Ribbon click Tools>Boolean>add. This will join two separate brick Feedline and Patch and make a single compound object (Fig. 13.11b).

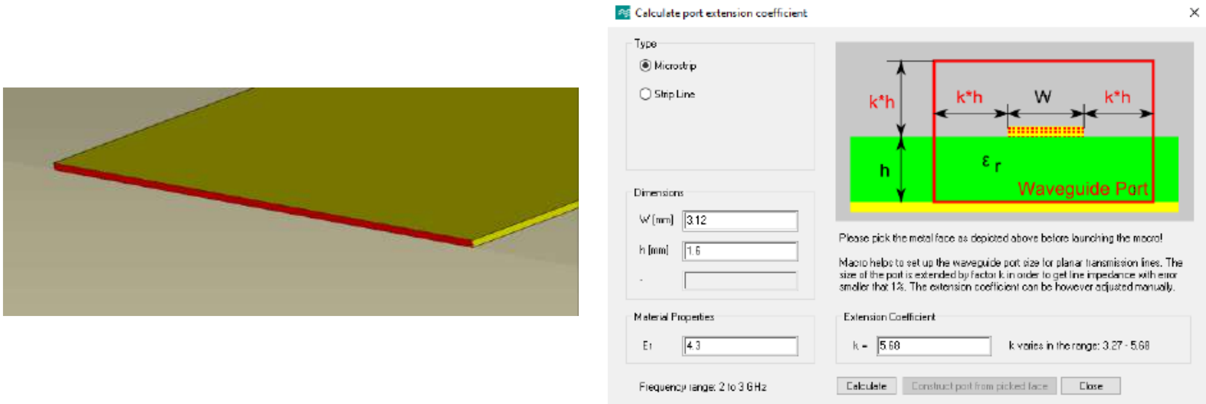


(a) (b)

Fig. 13.11: (a)Selecting two brick object in Navigation Tree (b)After Boolean operation

13.3.3 Input excitation:

1. Click 'Picks>Pick Face' from Picks group. Double click on the end of Feedline as shown in (Fig. 13.12a).



(a) (b)

Fig. 13.12:(a) Select feedline face for creation of Port (b) Port extension coefficient calculation window

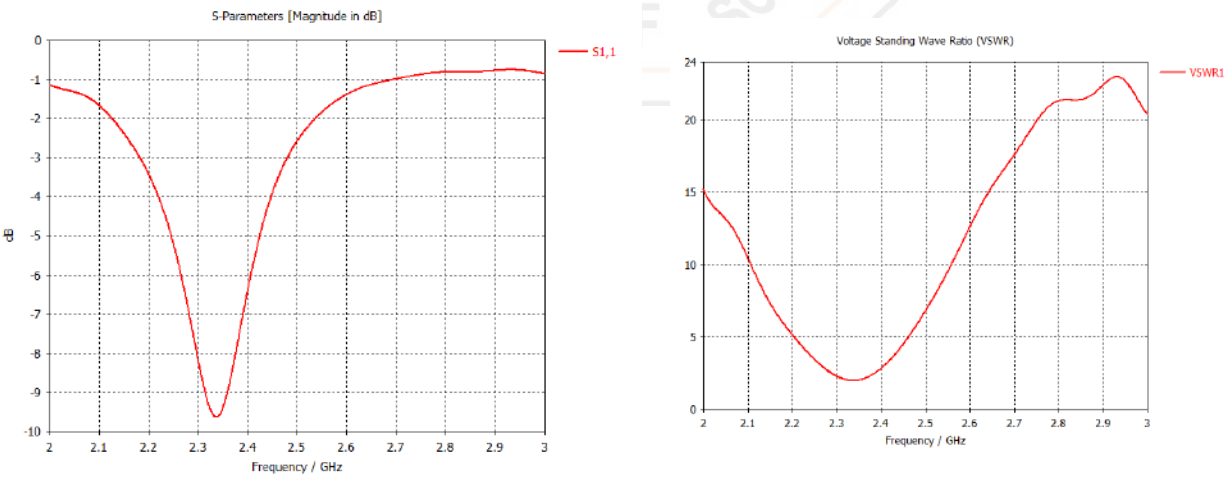
2. Click Home Ribbon. Click “Macros>Solvers> Ports>Calculate port extension coefficient” (Fig. 13.12b). Click ‘Calculate’. Extension co efficient will be calculated. Click ‘Construct port from picked face’. A port will be created based on the Dimension dan material property of the Face. CST will use the port to excite the entire structure. To see the port from ‘Navigation Tree’ click ‘Ports>Port 1’.

13.3.4 Monitors and Simulation Setting:

1. Click ‘Simulation’ ribbon.
2. Click ‘Field Monitor’ from ‘Monitor’ group.
3. In the monitor we can select the type of output we want to see and also at what frequency. To see the Electric field distribution at 2.4GHz, select ‘E-Field’ and enter ‘2.4’ next to frequency. Click Apply. A new entry will be created in ‘Navigation Tree>Field Monitor’.
4. Include ‘Farfield /RCS’ field monitor’ at 2.4 in the simulation.
5. To start the simulation, Click ‘Setup Solver’ in the ‘Solver’ group.
6. In the ‘Time Domain Solver parameters’ window Select ‘Port 1’ from ‘source type’ dropdown box. Keep all the other setting to the default position. The accuracy/ time taken to solve the problem can be modified by varying various parameter of the window.
7. Click ‘Start’ to start the simulation.

13.3.5 Result Post processing:

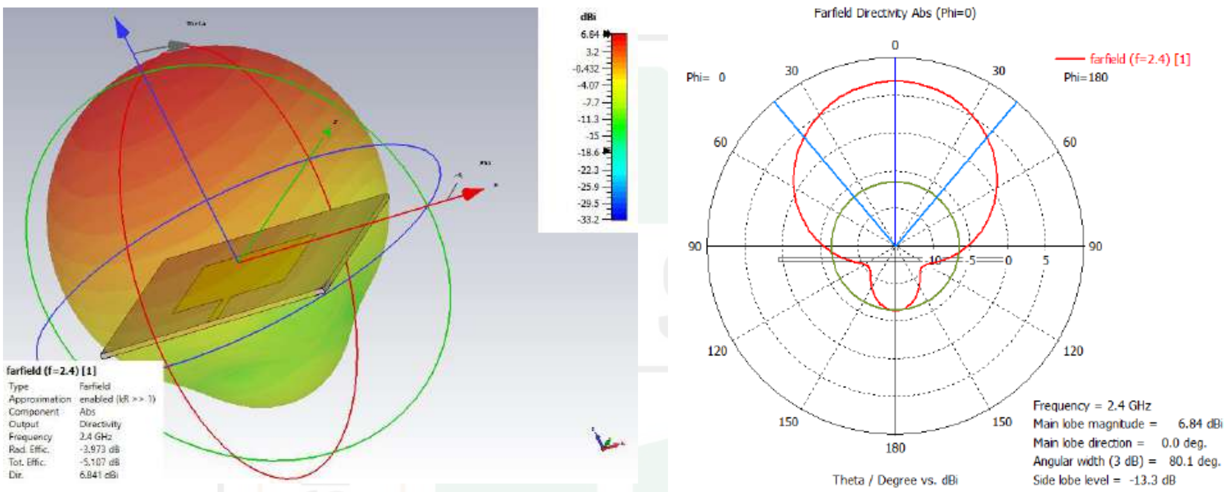
1. To see the S_{11} parameter of the antenna, Click Navigation Tree>1D Result>S-Parameter (Fig. 13.13a).



(a) (b)

Fig. 13.13: (a)S11 Plot of the MPA (b)VSWR of the antenna

2. VSWR or any other 1D result can be seen in this manner (Fig. 13.13b).
3. To find out the lowest point of the curve Click ‘1D Plot’ Ribon. In the ‘Markers’ section Click ‘Axis Marker> Move Marker to minimum’.
4. To see the data at any point of the curve, click ‘Curve Marker>Add Curve Markers’. Double click on the curve to see the x and y axis data on that point.
5. To see the Radiation pattern of the MPA, Click Navigation Tree>Farfields>farfield (f=2.4). The 3D radiation pattern can be seen (Fig. 13.14a).
6. To see the 1D Polar plot, click ‘Properties’ from the ‘Farfield’ ribbon.
7. In the Window select ‘1D Polar’ from the Plot type dropdown box. Enter Phi=0. Click ‘Ok’. This will create the 1D polar plot at Phi=0 axis (Fig. 13.14b).



(a) (b)

Fig. 13.14: (a)3D plot of the radiation pattern (b) 1D polar radiation plot of the MPA

13.4 Report

1. Calculate and simulate a MPA aimed at ____ frequency. Consider ____ as ϵ_r, ____ as the hight of the substrate and ___ as the hight of the copper layer. Also consider the ______shape patch and Recessed microstrip-line feed as the feedline technique.

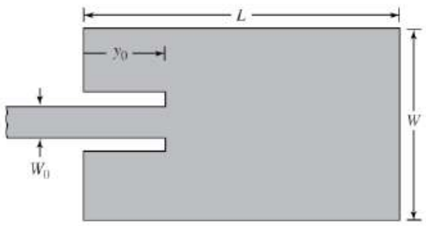


Fig. 13.15: Recessed microstrip-line feed

2. What is the effect of using the Recessed microstrip-line feed technique shown compared to the one used in the lab?
3. Discuss about the MPA characteristics from your simulation results.

13.5 Prepared by

1. Dr. A.K.M. Ehtesanul Islam, (Theory section)
2. Md. Aminur Rahman, (Simulation Section)

Experiment number: 14
Experiment Name: Low pass filter synthesis, verification and measurement

14.1 Objective:

1. Understand the theory behind Low pass filter
2. Understand the challenges behind circuit design in microwave frequency
3. Understand the design steps for lumped element and Microstrip based LPF
4. Simulate Microstrip and lumped element circuit in QUCS.
5. Perform a full electromagnetic simulation in CST Studio suite

14.2 Theory:

Network Synthesis:

A network analysis means that a network is analyzed using various network theory like KVL, KCL, Norton etc. to predict the response of the network. In contrast in Network synthesis a desired output is first selected and then the network is designed such that response of the network gives out the desired output.

Filter design by the insertion loss method

A perfect filter would have zero insertion loss in the passband, infinite attenuation in the stopband, and a linear phase response (to avoid signal distortion) in the passband. Practically it's not possible to create a filter like that and there is where various method is used to optimize the design a filter. In the insertion loss method, a filter response is defined by its insertion loss (IL) or power loss ratio, P_{LR}:

$$Power\ Loss\ ratio, P_{LR} = \frac{Power\ available\ from\ source}{Power\ delivered\ to\ load} = \frac{P_{in}}{P_L} = \frac{1}{1 - |\Gamma(\omega)|^2}$$

$$IL = 10\log(P_{LR}) = -10\log\left[1 - |\Gamma(\omega)|^2\right]$$

It can be shown that $|\Gamma(\omega)|^2$ is an even function of ω, so it can be expressed as a polynomial in ω^2. Thus:

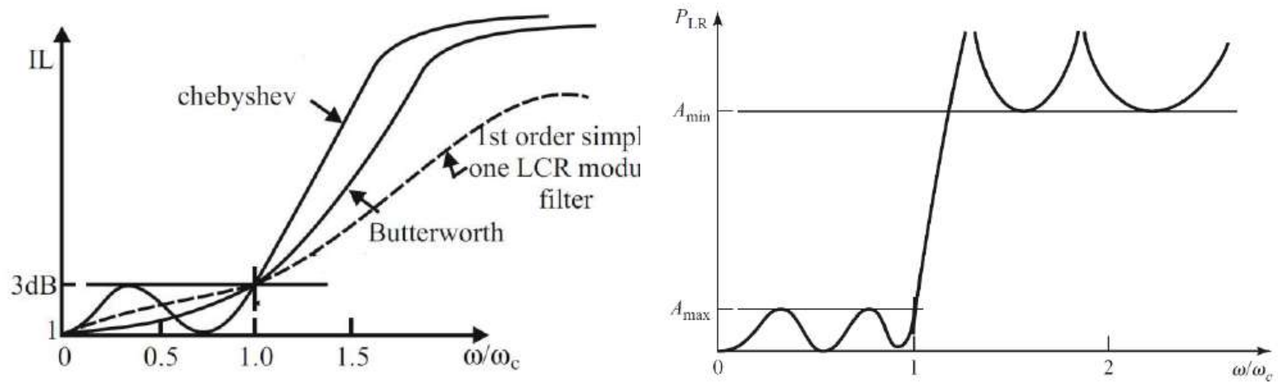
$$|\Gamma(\omega)|^2 = \frac{M(\omega^2)}{M(\omega^2) + N(\omega^2)} = 1 + \frac{M(\omega^2)}{N(\omega^2)}$$

So,

$$IL = 10\ log\left[\frac{M(\omega^2)}{N(\omega^2)}\right]$$

By selecting the polynomial ratio M/N the IL vs. frequency gives out various response. Some are:

(i) Butterworth (called maximally flat)
(ii) Chebyshev (called equal-ripple)
(iii) Ecliptic function
(iv) Linear phase



(a) (b)

Figure 14.1: Low pass filter Response (a) Insertion loss (IL) of Butterworth (maximally flat) and chebyshev (equal-ripple) (c) Power loss (P_{LR}) of Elliptic-function

Low pass filter and filter Order:

Binomial or Butterworth response provides the flattest possible passband response for a given filter complexity, or order. For a low-pass filter, it is specified by

$$P_{LR} = 1 + k^2 \left(\frac{\omega}{\omega_c}\right)^{2N} \quad (14.1)$$

$$IL(dB) = 10\ log\left[1 + k^2 \left(\frac{\omega}{\omega_c}\right)^{2N}\right] \quad (14.2)$$

where N is the order of the filter, ω_c is the cutoff frequency and K is a constant. If we select K=1 at cut off frequency $\omega = \omega_c$, the $IL = 10\log(2) = 3dB$. The passband extends from $\omega = 0$ to $\omega = \omega_c$. at the band edge the power loss ratio is $1 + k^2$ (3dB point in IL diagram). Also the rate of IL with depends on 2N in the equation.

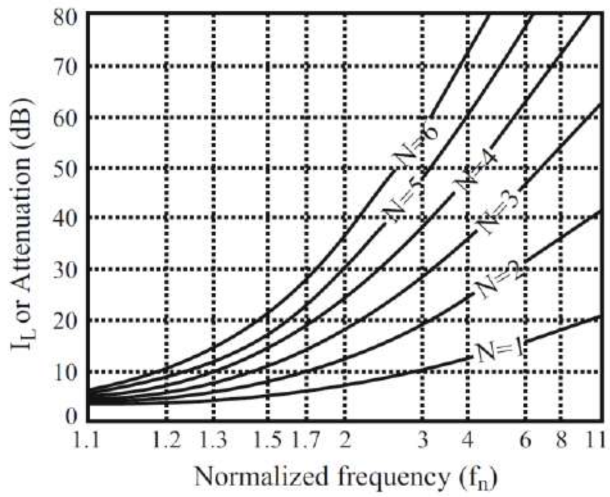


Figure 14.2: Butterworth filter (maximally flat) filter. Insertion loss in a low pass filter versus normalized frequency (fn) (for n = 1–6)

For $\omega < \omega_c$ it has low insertion loss

(i) For $\omega > \omega_c$ the insertion loss increases monotonically with frequency.

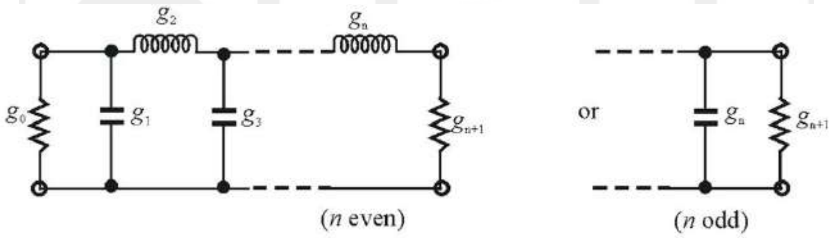
(ii) For $\omega \gg \omega_c$: $IL \simeq 10\log\left[K^2\left(\frac{\omega}{\omega^2}\right)^{2N}\right] = 20N\ log\left(\frac{\omega}{\omega^2}\right)$ for K=1, which shows that the insertion loss increases at the rate of 20N dB/decade.

Prototype filter

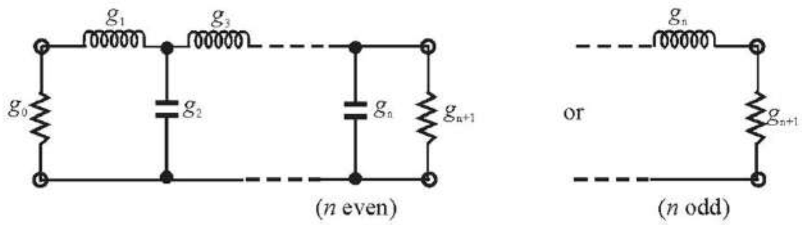
A low-pass *Prototype filter* is defined as a filter whose element values are **normalized** to make the source resistance or conductance equal to one, denoted by g_0 = 1, and the cutoff angular frequency to be unity, denoted by $\Omega_c = 1$(rad/s).

Two possible realizations of π-type and T-type of generic normalized low pass filter are shown in Fig. 14.3. The element values of L, C components in Fig. 14.3 are numbered as g_0 (for generate) to g_{n+1} (load). The elements in the circuit alternate between series inductance and shunt capacitance. These elements are defined as:

g_0 = Generator resistance in π-type filters (Fig. 14.3a) or generator conductance T–type filters (Fig. 14.3b)
g_n = Inductance for series inductor of the filter or capacitance for shunt capacitor of the filter (n = 1, 2, 3, … N)
g_{n+1} = Load resistance for p-type filter or load conductance for T-type filter.



(a)



(b)

Figure 14.3: (a) π-type (b)T-type prototype filter for LPF

For *Butterworth* response of the LPF the normalized prototype *element value* are:

$$g_0 = 1.0$$
$$g_i = 2\sin\left(\frac{(2i-1)\pi}{2N}\right)\ for\ i = 1\ to\ N \quad (14.3)$$
$$g_{N+1} = 1.0$$

Here N is the order of the filter.

Lumped Element value for practical LPF: Frequency and element transformation

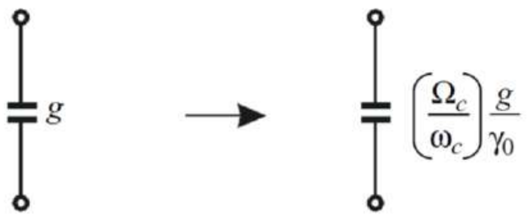
Previously for low pass ***prototype*** filters, which have a normalized source resistance/conductance $g_0 = 1.0$ and a cutoff frequency of $\Omega_c = 1$ we obtained normalized prototype *element values (g)*.To obtain actual lumped values for ***practical*** filters from the normalized prototype *element value, frequency* and *element transformations* need to made.

For inductor the equation becomes:

$$L = \left(\frac{\Omega_c}{\omega_c}\right)\gamma_0 g$$
$$\Rightarrow L = \left(\frac{\Omega_c}{2\pi f_c}\right)\left(\frac{Z_0}{g_0}\right) g$$
$$\Rightarrow L = \left(\frac{Z_0}{2\pi f_c}\right) g \qquad (14.4.1)$$

Here the g is the *element value* of an inductance. In case of a capacitor the transformation equation becomes:

$$C = \left(\frac{\Omega_c}{\omega_c}\right)\frac{g}{\gamma_0}$$
$$\Rightarrow C = \left(\frac{\Omega_c}{2\pi f_c}\right)\left(\frac{g_0}{Z_0}\right) g$$
$$\Rightarrow C = \left(\frac{1}{Z_0 2\pi f_c}\right) g \qquad (14.4.2)$$



In this case the g is the *element value* of a capacitor.

Here Impedance scaling factor for a 50ohm line is $\gamma_0 = \frac{Z_0}{g_0} = \frac{50}{1} = 50$.

Stepped Impedance implementation:

In microwave frequency the lumped element values required for any filter becomes very difficult to get from the market in some cases impossible to buy out of the store and need to custom made. Also the frequency response and size of the lumped element also need to be taken into account. For that purpose, in very high frequency Microstrip based filter implementation is used.

Stepped Impedance filters use alternating sections of very high and very low characteristic impedance lines. They are also known as hi-Z, Low-Z filters. They are easier to design and take up less space than a similar low-pass filter using stubs. As there are approximations involved their electrical performance is not as good, so they can't be used where a sharp cutoff is required.

A short Length of high-impedance (Z_c) lossless line terminated at both ends by relatively low impedance, Z_0 (Fig 14.4a) is represented by a π-equivalent circuit (Fig 14.4b).

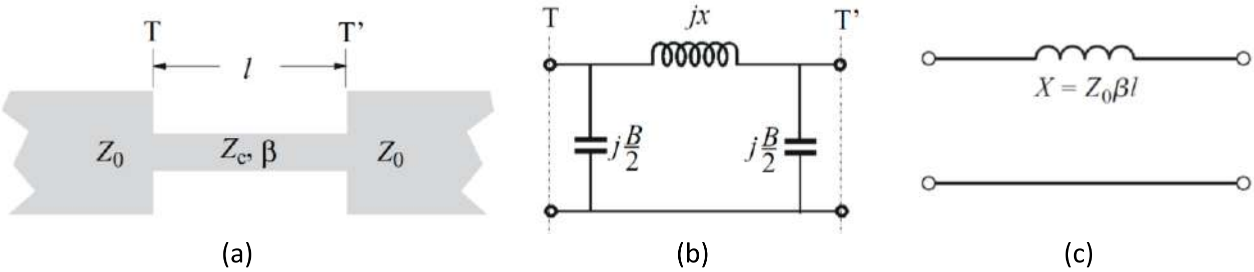


Fig. 14.4: (a)High-impedance Transmission line (b) Equivalent circuit (c) Approximated circuit

The Circuit parameter for Fig 14.4b :

$$X = Z_c \sin(\beta l)$$
$$\Rightarrow \frac{X}{Z_c} = \sin\left(\frac{2\pi}{\lambda_{gh}} l\right)$$
$$\Rightarrow l = \frac{\lambda_{gh}}{2\pi} \sin^{-1}\left(\frac{X}{Z_c}\right) \quad (14.5.1)$$

$$\frac{B}{2} = \frac{1}{Z_c} \tan\left(\frac{\pi}{\lambda_{gh}} l\right)$$

Assume short length of line $(l < \lambda_g/8)$ and large characteristic impedance ($Z_c \gg Z_0$) we can ignore the shunt capacitance, so $X \simeq Z_0\beta l$ and $B \simeq 0$. Which implies a series inductor (Fig: 14.4 c) and the equation 14.5.1 becomes

$$l = \left(\frac{\lambda_{gh}}{2\pi}\right)\left(\frac{X}{Z_c}\right)$$
$$l = \left(\frac{\lambda_{gh}}{2\pi}\right)\left(\frac{Z_0}{Z_h}\right) g \quad (14.5.2)$$

Here $Z_c = Z_h/Z_0$= Impedance of the high impedance line. λ_{gh}= Guided or effective wavelength of the high impedance line, $X = g_k$= Normalized element value of the inductor.

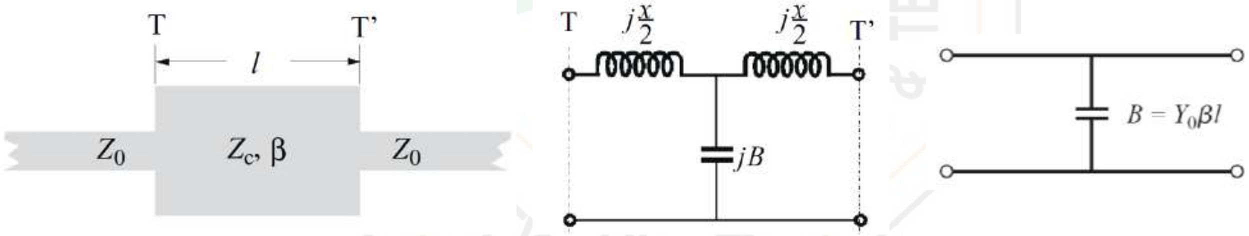
Length can also be calculated using

$$X \simeq Z_0 \beta l$$

$$\beta l = \frac{1}{Z_0} X = \left(\frac{R_0}{Z_h}\right) X \quad (14.5.3)$$

Here βl is the electrical length (in degree) of the high impedance (inductive line) and $X = g_k$= Normalized element value of the inductor.

For a short length of low-impedance (Z_c) lossless line terminated at either end by relatively high impedance (Z_0) (Fig 14.5a) is represented by a T-equivalent circuit with the circuit parameters



(a) (b) (c)

Fig 14.5: (a)Low Impedance Lossless line (b) Equivalent circuit (c) Approximated circuit

The Circuit parameter for Fig 14.5b :

$$B = Y_0 \sin(\beta l)$$
$$\Rightarrow B = \frac{1}{Z_c} \sin\left(\frac{2\pi}{\lambda_{gl}} l\right)$$
$$\Rightarrow l = \frac{\lambda_{gl}}{2\pi} \sin^{-1}(BZ_c) \quad (14.6.1)$$

$$\frac{x}{2} = Z_c \tan\left(\frac{\pi}{\lambda_g} l\right)$$

Now assume short length of line $\left(l < \lambda_g/8\right)$ and small characteristic impedance ($Z_c \ll Z_0$) we can ignore the series reactance, so $X \simeq 0$ and $B \simeq Y_0\beta l$. Which implies a shunt capacitor (Fig: 14.5c) and the equation can be approximated by

$$l = \left(\frac{\lambda_{gl}}{2\pi}\right)(BZ_c)$$

$$l = \left(\frac{\lambda_{gl}}{2\pi}\right)\left(\frac{Z_l}{Z_0}\right) g \qquad (14.6.2)$$

Here $Z_c = Z_l/Z_0$= Impedance of the low impedance line. λ_{gl}= Guided or effective wavelength of the low impedance line, $B = g_k$= Normalized element value of the capacitor.

Length can also be calculated using

$$B \simeq Y_0\beta l$$

$$\beta l = \frac{1}{Y_0}B = Z_0 B = \left(\frac{Z_l}{R_0}\right)B \qquad (14.6.3)$$

Here βl is the electrical length (in degree) of the low impedance (capacitive line) and $B = g_k$= Normalized element value of the capacitor.

Here in both the cases $\theta = \beta l$ is known as the electrical length of the line for a physical length l . The phase constant as it is used in Microstrip line it is known as effective phase constant $\beta = \frac{2\pi}{\lambda_g}$. The guided or effective wavelength is defined as $\lambda_g(mm) = \frac{300}{f(GHz)\sqrt{\varepsilon_{re}}}$ and ε_{re} is the effective permittivity of the Microstrip line. (**See Appendix G**). To determine the physical length following formula can be used:

$$l = \frac{\lambda_g}{2\pi} \times \theta(in\ radian)$$

So, the series inductors of a low-pass prototype can be replaced with high-impedance line sections ($Z_0 = Z_h$), and the shunt capacitors can be replaced with low-impedance line section ($Z_0 = Z_l$). The ratio Z_h/Z_l should be as large as possible, so the actual values of $Z_h$and Z_l are usually set to the highest and lowest characteristic impedance that can be practically fabricated.

14.3 Filter synthesis:

In this section we see step by step method by which a LPF is calculated and then transformed into a Microstrip implementation. Two different ways to implement the stepped impedance method is also shown, one with the entire calculation and another use different line calculator tools.

14.3.1 Determine the element value for the Prototype LPF

Step 1: Determine the degree of the LPF

To determine the degree of Butterworth low pass filter, put k=1 in equation 14.2 we can get

$$N \geq \frac{\log(10^{0.1L_{AS}} - 1)}{2\log\left(\frac{\omega}{\omega_c}\right)} \tag{14.7}$$

L_{AS}=minimum stopband attenuation or insertion loss (dB), ω= frequency at which insertion loss was calcualted.

Step 2: Determine the normalized *element value* of the *Prototype* LPF using following equations.

$$g_0 = 1.0$$

$$g_i = 2\sin\left(\frac{(2i-1)\pi}{2N}\right)\ for\ i = 1\ to\ N$$

$$g_{N+1} = 1.0$$

Step 4: Select π or T type topology for the prototype LPF (Fig. 14.6a).

14.3.2 Lumped element filter synthesis

Use equation $L = \left(\frac{Z_0}{2\pi f_c}\right) g$ and $C = \left(\frac{1}{Z_0 2\pi f_c}\right) g$ to determine the value for L and C of the filter from various value of g_k (Fig. 14.6b).

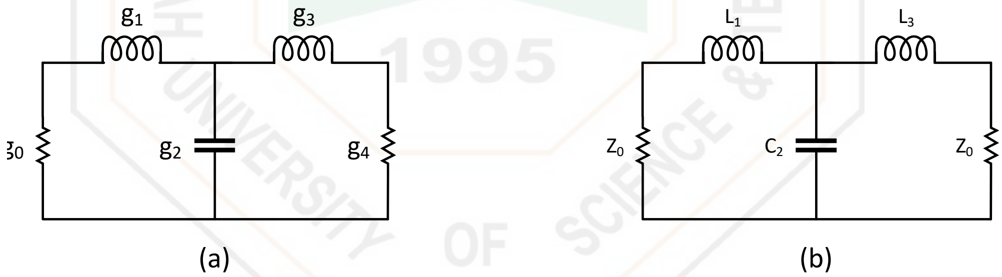


Figure 14.6: *Prototype* LPF with normalized *element value* (b) Practical value with Lumped element

14.3.3 Stepped impedance implementation in microstrip: rigorous calculation (Without any software)

Step 1: Select the value of the high impedance Z_h and low impedance Z_l line for the stepped impedance LPF.

Step 2: As we know both high and low Impedance value; height of the Microstrip substrate h, dielectric constant of the substrate ε_r now using equation g 2.1 or g 2.2 (see appendix G) we can solve for the width of the element W.

So determine the width of the inductor and capacitor line W_l and W_h.

Step 3: Determine the effective dielectric constant, ε_{re} for both capacitive and inductive line using equation g1.1 or g 1.2. Use the W_l and W_h from the previous step.

Step 4: Determine the effective wavelength for the capacitive and inductive line use $\lambda_g(mm) = \frac{300}{f(GHz)\sqrt{\varepsilon_{re}}}$ equation. Use the effective dielectric constant value from previous step.

As the effective dielectric constant are different for capacitive and inductive line, the λ_g for capacitive and inductive line will also be different.

Step 5: To determine the length of the transmission line use $l = \left(\frac{\lambda_{gl}}{2\pi}\right)\left(\frac{Z_l}{Z_0}\right) g$ and $l = \left(\frac{\lambda_{gh}}{2\pi}\right)\left(\frac{Z_0}{Z_h}\right) g$ using the g_k values determined for capacitor and inductors respectively.

14.3.4 Stepped impedance implementation in microstrip: Using line calculator software

Step 1: Select the value of the high impedance Z_h and low impedance Z_l line for the stepped impedance LPF.

Step 2: Determine the electrical length using $\beta l = \left(\frac{Z_l}{R_0}\right) B$ and $\beta l = \left(\frac{R_0}{Z_h}\right) X$ for capacitor and inductor respectively. Here B and X are the g_k values for capacitor and inductor respectively.

Step 3: Using any Microstrip line calculator (Txline, QUCS or any web-based tool like emtalk.com/mscalc.php) synthesize the width and length of the inductor and capacitor.

For this, use the value of the electrical length, impedance of the line, cutoff frequency and Microstrip parameter like substrate height and dielectric constant of the substrate.

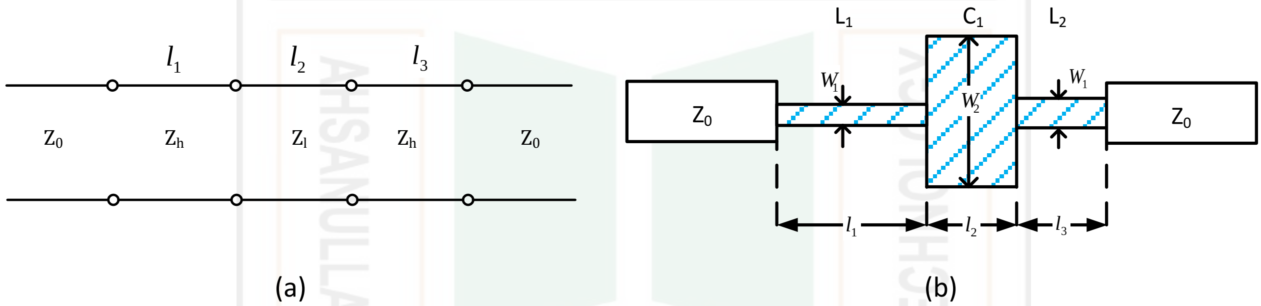


Figure 14.7: (a) Stepped impedance implementation. (b) Microstrip layout of Stepped impedance implementation.

14.4.1 Example: Microwave Engineering by David M. Pozar: Example 8.6

Design a stepped-impedance low-pass filter having a maximally flat response and a cutoff frequency of 2.5 GHz. It is desired to have more than 20 dB insertion loss at 4 GHz. The filter impedance is 50Ω; the highest practical line impedance is 120Ω and the lowest is 20Ω.

Consider the effect of losses when this filter is implemented with a microstrip substrate having d = 0.158 cm, $\varepsilon_r = 4.2$, tan δ = 0.02, and copper conductors of 0.5 mil thickness.

Solution: Table: Calculated output for Example 8.6

Section	Normalized element value		Lumped Value		Impedance of the section	Electrical length βl (deg.)	Width of the section $W_i(mm)$	Length of the section $l_i(mm)$
1	g_1	0.517	C_1	$\left(\frac{g_1}{Z_0\omega_c}\right) = .658pF$	Z_l =20	$\left(\frac{Z_l}{R_0}\right) g_1 = 11.8$	11.3	2.05
2	g_2	1.414	L_2	$\left(\frac{Z_0}{\omega_c}\right) g_2 = 4.5nH$	Z_h =120	$\left(\frac{R_0}{Z_h}\right) g_2 = 33.8$	0.428	6.63
3	g_3	1.932	C_3	$\left(\frac{g_3}{Z_0\omega_c}\right) = 2.46pF$	Z_l =20	$\left(\frac{Z_l}{R_0}\right) g_3 = 44.3$	11.3	7.69
4	g_4	1.932	L_4	$\left(\frac{Z_0}{\omega_c}\right) g_4 = 6.15nH$	Z_h =120	$\left(\frac{R_0}{Z_h}\right) g_4 = 46.1$	0.428	9.04
5	g_5	1.414	C_5	$\left(\frac{g_5}{Z_0\omega_c}\right) = 1.8pF$	Z_l =20	$\left(\frac{Z_l}{R_0}\right) g_5 = 32.4$	11.3	5.63
6	g_6	0.517	L_6	$\left(\frac{Z_0}{\omega_c}\right) g_6 = 1.64nH$	Z_h =120	$\left(\frac{R_0}{Z_h}\right) g_6 = 12.3$	0.428	2.41

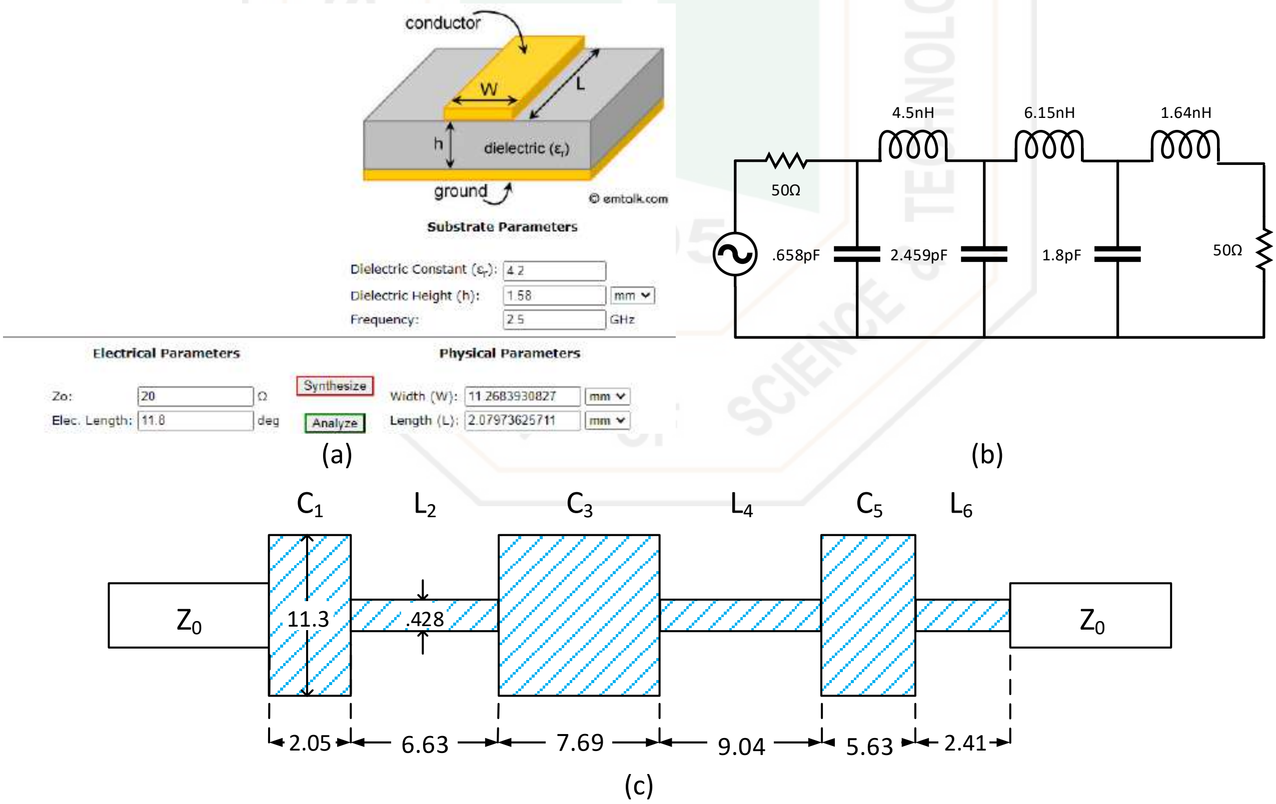
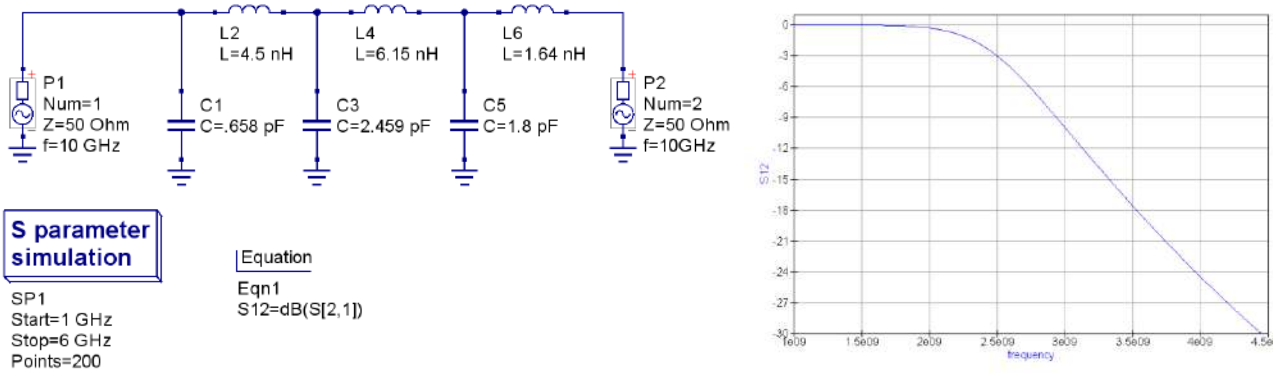


Figure 14.8: (a) emtalk.com line calculator (b)Lumped element (c)Microstrip implementation

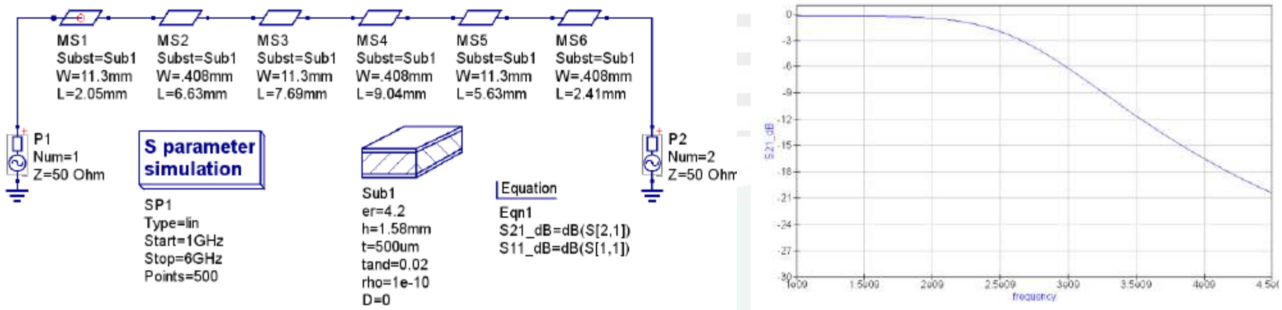
14.4.2 Simulation using QUCS (Lumped element):



(a) (b)

Figure 14.9: Simulation done using lumped element value (a) Circuit with lumped element (b) Output S12 (dB)

14.4.3 Simulation using QUCS (Stepped-impedance implementation):



(a) (b)

Figure 14.10: Simulation done using Microstrip implementation (a) Microstrip circuit (b) Output S12 (dB)

14.4.4 Full Electromagnetic simulation using CST Studio suite:

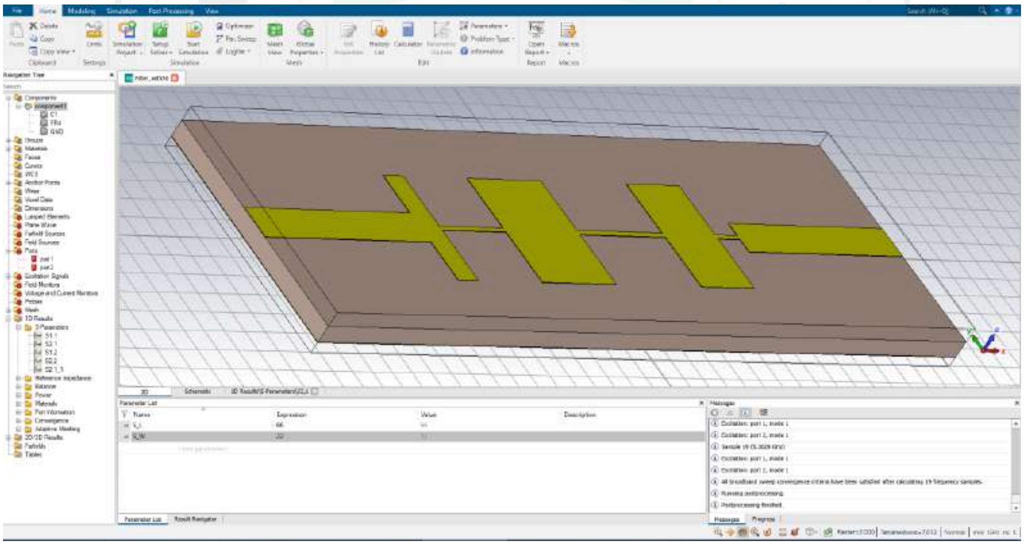


Figure 14.11:(a) Microstrip LPF in CST

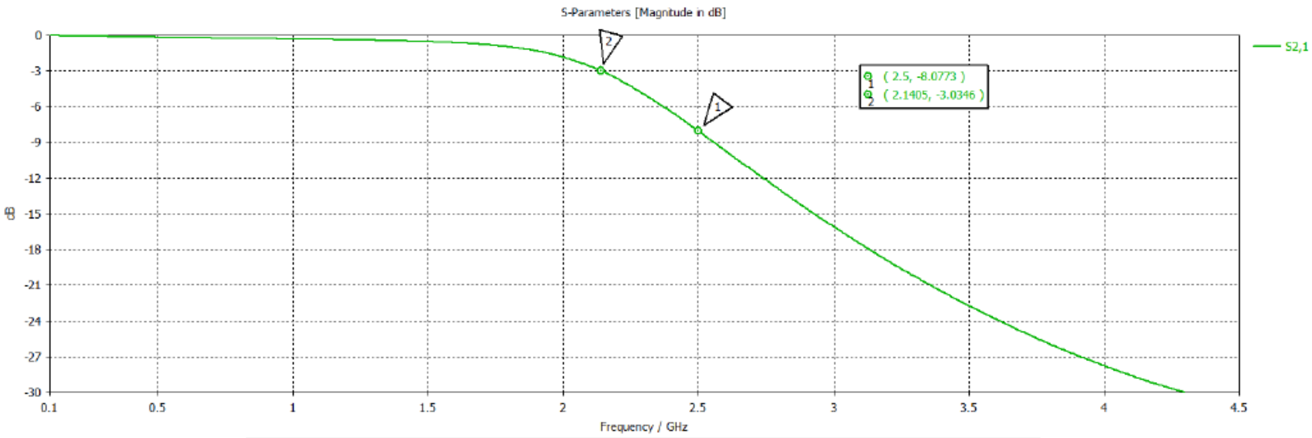


Figure 14.11: (b) S21(dB) of the stepped impedance LPF in CST

14.5 Report:

1. What are the difficulty with lumped element filter?
2. Name some other ways the LPF can be implemented in Microstrip other then stepped impedance method.
3. Why there is a deviation between 3 simulated output.
4. Design a stepped-impedance low-pass filter having a maximally flat response and a cutoff frequency of ___GHz. It is desired to have more than ___ dB insertion loss at ___ GHz. The filter impedance is 50Ω; the highest practical line impedance is ___Ω and the lowest is ___Ω.
   Assume typical FR4 material as substrate
5. Simulate the filter mentioned above in _____ software using Lumped element/stepped impedance method.

14.6 Project:

- Design the filter above a PCB board using PCB etching / PCB CNC milling/ copper tape over and solder SMA connector on both end and characterize it using the Network analyzer or a RF signal source and spectrum analyzer.

14.7 Prepared by:

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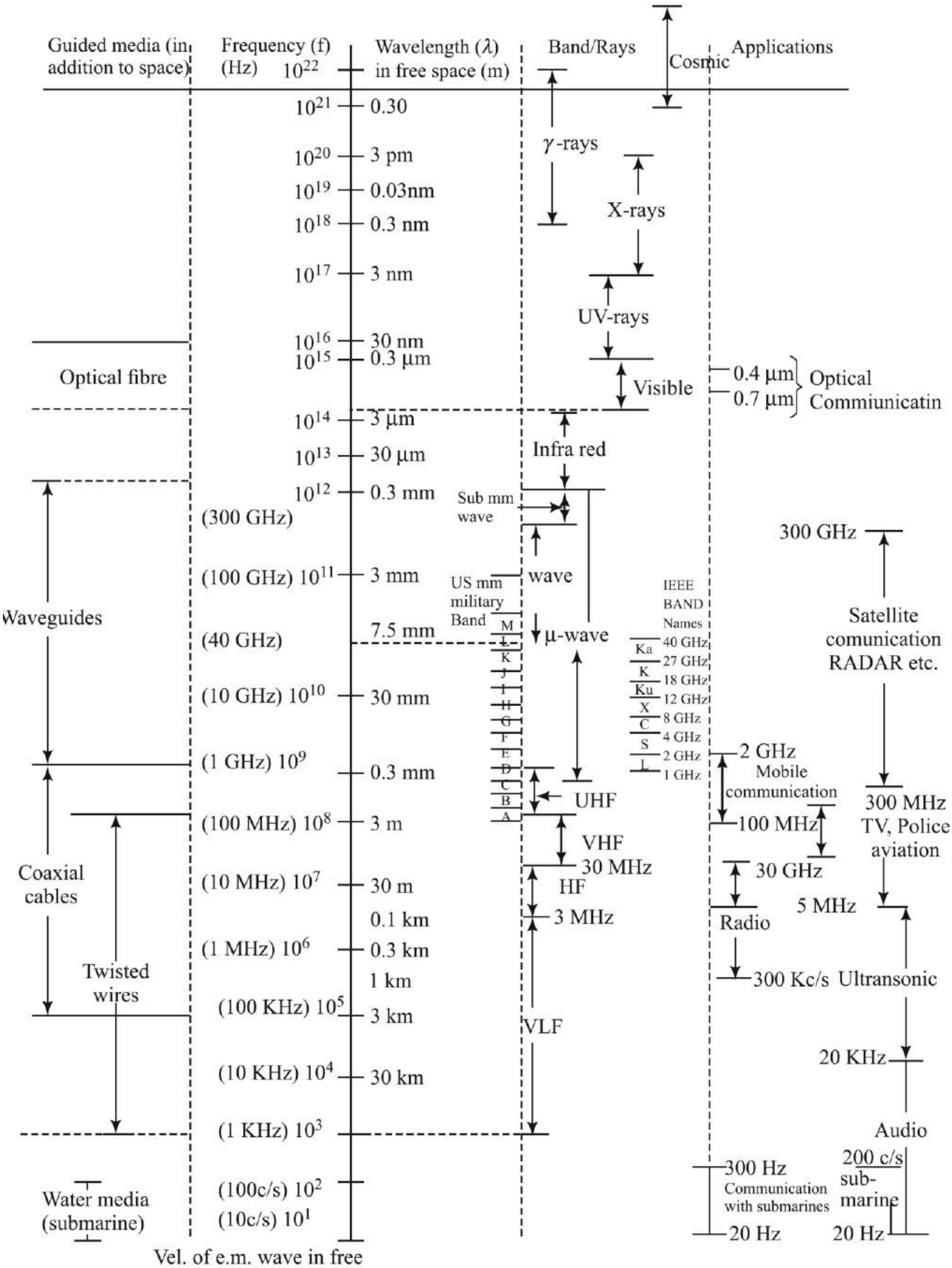
1995

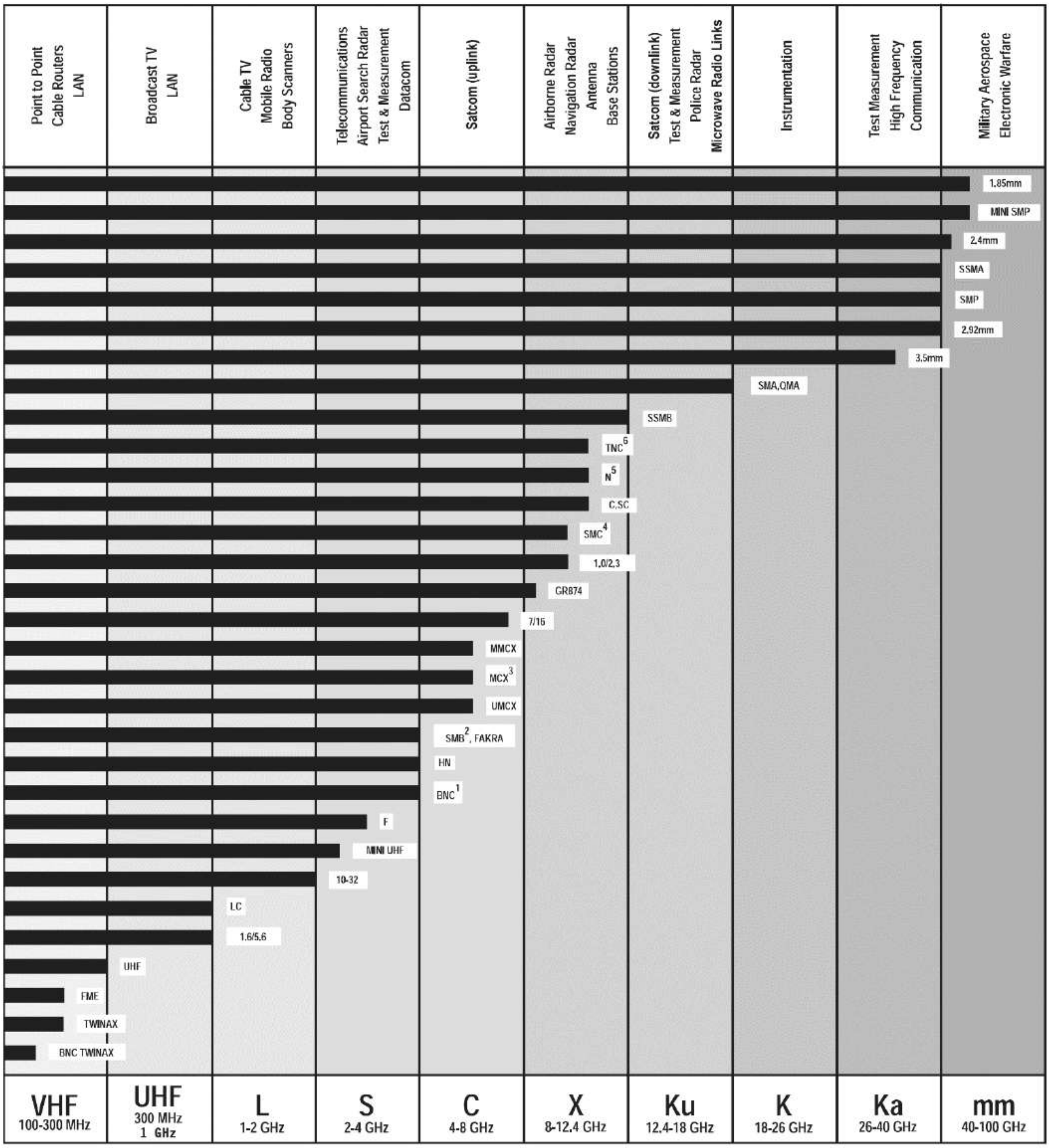
Appendix A1: Full electromagnetic spectrum



Vel. of e.m. wave in free space = $c = \lambda f = 3 \times 10^8$ m/sec.

Reference: Microwave, Radar & RF Engineering, Prakash Kumar Chaturvedi

Appendix A2: Microwave Connector type



Refrence: Pasternack Inc.

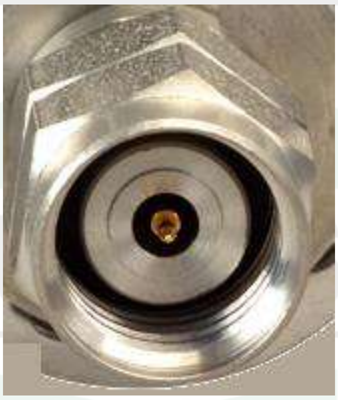
Appendix A3: Microwave connectors

BNC

N

SMA

K



V

1 mm

Table: List of most popular microwave and mm wave connectors (see appendix A2 for detailed list)

Name	Separation between the conductors (mm)	Maximum Frequency (GHz)	Coupling Torque (N-cm)	Dielectric
W1*	1	110	45	Air
V	1.85	70	90	Air
	2.4	50	90	Air
K	2.92	40	90	Air
APC-3.5	3.5	34	90	Air
N	-	18	135	TFE
SMA	-	18	56	Teflon
APC-7	7			
BNC	6.35	4	N/A	PTFE

* W1 is the name given by Anritsu which is equivalent to 1mm connector.

Appendix B1: National Frequency Allocation Plan (NFAP)

72-108 MHz

<table>
<tr><th colspan="3">Allocation to services by ITU</th><th rowspan="2">National Allocations</th><th rowspan="2">Usage</th></tr>
<tr><th>Region 1</th><th>Region 2</th><th>Region 3</th></tr>
<tr><td rowspan="3"></td><td>72-73
FIXED
MOBILE</td><td rowspan="3"></td><td rowspan="3"></td><td rowspan="3"></td></tr>
<tr><td>73-74.6
RADIO ASTRONOMY
5.178</td></tr>
<tr><td>74.6-74.8
FIXED
MOBILE</td></tr>
<tr><td>74.8-75.2</td><td colspan="2">AERONAUTICAL RADIONAVIGATION
5.180 5.181</td><td>74.8-75.2 (SHRD)
AERONAUTICAL
RADIONAVIGATION
5.180</td><td>1. ILS marker radio beacons (ground-air). Further details are available in ICAO Annex 10, volume 1, chapter 3, sections 3.1.7 and 3.6)</td></tr>
<tr><td rowspan="4">75.2-87.5
FIXED
MOBILE except aeronauticalmobile
5.175 5.179 5.187</td><td colspan="2">75.2-75.4
FIXED
MOBILE
5.179</td><td>75.2-75.4 (GOVT)
FIXED
MOBILE</td><td>1. Long range fixed and mobile applications including PMR</td></tr>
<tr><td>75.4-76
FIXED
MOBILE</td><td rowspan="2">75.4-87
FIXED
MOBILE
5.182 5.183 5.188</td><td>75.4-76.25 (GOVT)
FIXED
MOBILE</td><td>1. Long range fixed and mobile applications including PMR</td></tr>
<tr><td rowspan="3">76-88
BROADCASTING
Fixed
Mobile
5.185</td><td>76.25-87 (CIVIL)
FIXED
MOBILE</td><td>1. Simplex operation mode PMR in 76.5-77MHz, in 77-82 MHz paired with 82.0-87MHz
2. Rural cordless telephone extension in 47.25–47.5 MHz paired with 76.25–76.5MHz</td></tr>
<tr><td rowspan="3">87-100
FIXED
MOBILE
BROADCASTING</td><td rowspan="4">87-108 (CIVIL)
BROADCASTING</td><td rowspan="4">1. VHF FM analog sound broadcasting with 100 kHz channel spacing</td></tr>
<tr><td rowspan="2">87.5-100
BROADCASTING
5.190</td></tr>
<tr><td>88-100
BROADCASTING</td></tr>
<tr><td>100-108</td><td colspan="2">BROADCASTING
5.192 5.194</td></tr>
</table>

Reference: Bangladesh National Frequency Allocation Plan (NFAP)

Appendix C1: Power to Voltage conversion table

P (dBm)	P (mW)	V_{RMS} (V)	Vp (V)[1]	Vpp (V)	P (dBm)	P (mW)	V_{RMS} (V)	Vp (V)[1]	Vpp (V)
-30	0.001	0.007	0.010	0.020	0	1.000	0.224	0.316	0.632
-29	0.001	0.008	0.011	0.022	1	1.259	0.251	0.355	0.710
-28	0.002	0.009	0.013	0.025	2	1.585	0.282	0.398	0.796
-27	0.002	0.010	0.014	0.028	3	1.995	0.316	0.447	0.893
-26	0.003	0.011	0.016	0.032	4	2.512	0.354	0.501	1.002
-25	0.003	0.013	0.018	0.036	5	3.162	0.398	0.562	1.125
-24	0.004	0.014	0.020	0.040	6	3.981	0.446	0.631	1.262
-23	0.005	0.016	0.022	0.045	7	5.012	0.501	0.708	1.416
-22	0.006	0.018	0.025	0.050	8	6.310	0.562	0.794	1.589
-21	0.008	0.020	0.028	0.056	9	7.943	0.630	0.891	1.783
-20	0.010	0.022	0.032	0.063	10	10.000	0.707	1.000	2.000
-19	0.013	0.025	0.035	0.071	11	12.589	0.793	1.122	2.244
-18	0.016	0.028	0.040	0.080	12	15.849	0.890	1.259	2.518
-17	0.020	0.032	0.045	0.089	13	19.953	0.999	1.413	2.825
-16	0.025	0.035	0.050	0.100	14	25.119	1.121	1.585	3.170
-15	0.032	0.040	0.056	0.112	15	31.623	1.257	1.778	3.557
-14	0.040	0.045	0.063	0.126	16	39.811	1.411	1.995	3.991
-13	0.050	0.050	0.071	0.142	17	50.119	1.583	2.239	4.477
-12	0.063	0.056	0.079	0.159	18	63.096	1.776	2.512	5.024
-11	0.079	0.063	0.089	0.178	19	79.433	1.993	2.818	5.637
-10	0.100	0.071	0.100	0.200	20	100.000	2.236	3.162	6.325
-9	0.126	0.079	0.112	0.224	21	125.893	2.509	3.548	7.096
-8	0.158	0.089	0.126	0.252	22	158.489	2.815	3.981	7.962
-7	0.200	0.100	0.141	0.283	23	199.526	3.159	4.467	8.934
-6	0.251	0.112	0.158	0.317	24	251.189	3.544	5.012	10.024
-5	0.316	0.126	0.178	0.356	25	316.228	3.976	5.623	11.247
-4	0.398	0.141	0.200	0.399	26	398.107	4.462	6.310	12.619
-3	0.501	0.158	0.224	0.448	27	501.187	5.006	7.079	14.159
-2	0.631	0.178	0.251	0.502	28	630.957	5.617	7.943	15.887
-1	0.794	0.199	0.282	0.564	29	794.328	6.302	8.913	17.825
0	1.000	0.224	0.316	0.632	30	1000.000	7.071	10.000	20.000

[1] For Square wave signal, $V_P = V_{RMS}$.

Note: The converted voltages in the above table are for R = 50 Ω

$P_{(dBm)} = 10 \log_{10} P_{(mW)}$

$P_{(mW)} = 10^{(P_{(dBm)}/10)}$

$P_{(mW)} = [V_{RMS\,(V)}]^2 * 10^3/R$

$V_{RMS\,(V)} = \sqrt{(P_{(mW)} \cdot R / 10^3)}$

$V_P = \sqrt{2} * V_{RMS}$, for sinusoid – Fig. a.

$V_P = V_{RMS}$, for square wave – Fig. b.

$V_{P\text{-}P} = 2 * V_P$

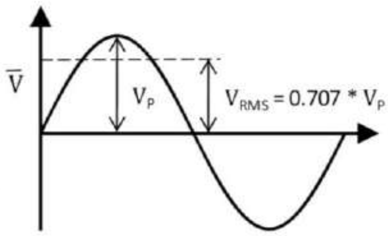


Fig. a. Sinusoidal signal

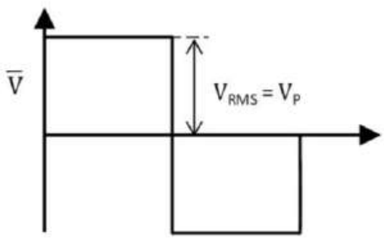
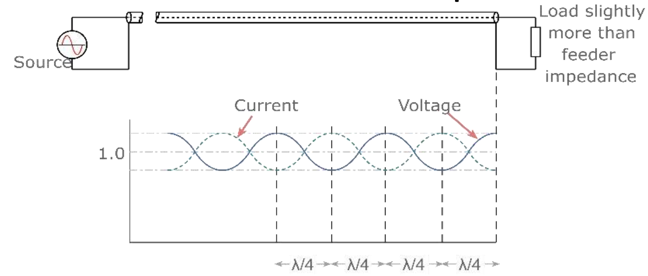
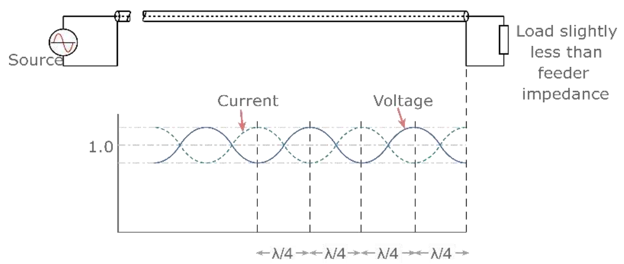


Fig. b. Square wave signal

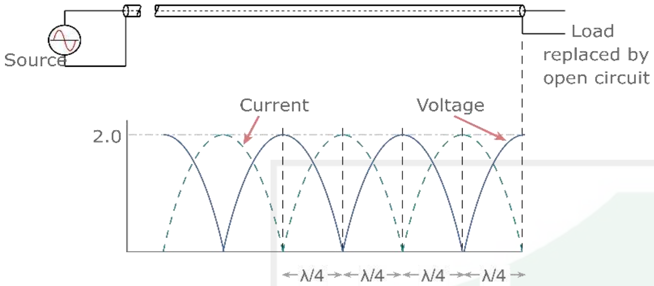
Apendix C2: Standing wave and VSWR



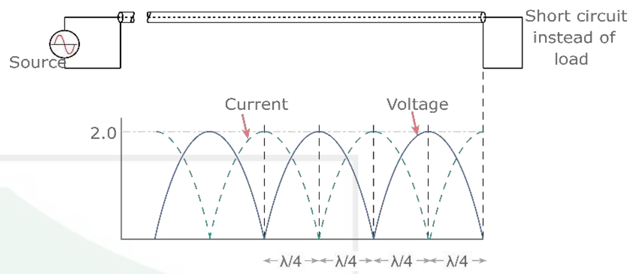
$Z_L > Z_0$



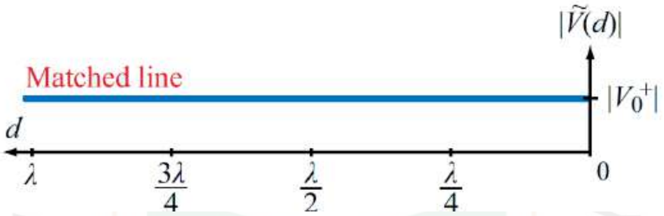
$Z_L < Z_0$



$\Gamma = 1\ or\ Z_L = \infty$ **(Open Circuit)**



$\Gamma = -1\ or\ Z_L = 0$ **(Short Circuit)**

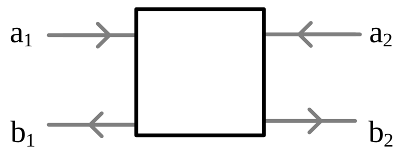


$\Gamma = 0$ or $Z_L = Z_0$ **(Match)**

Voltage Standing Wave Ratio $VSWR = \frac{\lvert V_{max}\rvert}{\lvert V_{min}\rvert}$	**Reflection coefficient** $\lvert\Gamma\rvert = \frac{VSWR-1}{VSWR+1}$	**Return Loss (dB)** $-20\log\lvert\Gamma\rvert$	**Reflected power (%)** $100 * \lvert\Gamma\rvert^2$	**Mismatch loss (dB)** $-10 * \log_{10}(1-\lvert\Gamma\rvert^2)$	**Comment**
1	0.000	∞	0.00	0.000	Matched load
1.5	0.200	13.979	4.00	0.177	
2	0.333	9.542	11.11	0.512	See example
3	0.500	6.021	25.00	1.249	
4	0.600	4.437	36.00	1.938	
5	0.667	3.522	44.44	2.553	
7	0.750	2.499	56.25	3.590	
9	0.800	1.938	64.00	4.437	
10	0.818	1.743	66.94	4.807	
15	0.875	1.160	76.56	6.301	
20	0.905	0.869	81.86	7.413	
100	0.980	0.174	96.08	14.066	
∞	Γ =1	0	100	∞	Open Load
∞	Γ =-1	0	100	∞	Short Load

Example:
If 1000 watts (60 dBm/30 dBW) is applied to an antenna with a VSWR value of 2, the return loss would be 9.54 dB. Therefore, 111.1 watts would be reflected and 888.9 watts (59.488 dBm/29.488 dBW) would be transmitted, so the mismatch loss would be 0.512 dB.

Apendix E: Lossless Two Port network:



For a lossless network, Power leaving =power entering

$$|b_1|^2 + |b_2|^2 = |a_1|^2 + |a_2|^2 \qquad \text{(E1)}$$

$$\Rightarrow \begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

$$\Rightarrow [b] = [S][a] \qquad \text{(E2)}$$

Now, we can

$$[b^T][b^*] = [b_1 \quad b_2] \begin{bmatrix} b_1^* \\ b_2^* \end{bmatrix}$$

$$= [b_1 b_1^* + b_2 b_2^*]$$

$$= |b_1|^2 + |b_2|^2$$

Substituting the value in Eq. E1, we have

$$[b^T][b^*] = [a^T][a^*] \qquad \text{(E3)}$$

Also, we know that

$$[b] = [S][a]$$

So $[b]^T = [a]^T[S]^T$ and also $[b^*] = [S^*][a^*]$

So, substituting the value in Eq. E3

$$\Rightarrow [a]^T[S]^T[S^*][a^*] = [a^T][a^*]$$

$$\Rightarrow [a]^T[S]^T[S^*][a^*] - [a^T][a^*] = 0$$

$$\Rightarrow [a]^T\{[S]^T[S^*] - [I]\}[a^*] = 0$$

Here a is power input, so $a \neq 0$,

$$\boxed{[S]^T[S^*] = [I]}$$

For example, for a two port network, the S-matrix

$$S = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix}, [S]^T = \begin{bmatrix} S_{11} & S_{21} \\ S_{12} & S_{22} \end{bmatrix}, [S^*] = \begin{bmatrix} S_{11}^* & S_{12}^* \\ S_{21}^* & S_{22}^* \end{bmatrix}$$

So,

$$[S]^T[S^*] = [I]$$

$$\Rightarrow \begin{bmatrix} S_{11} & S_{21} \\ S_{12} & S_{22} \end{bmatrix} \begin{bmatrix} S_{11}^* & S_{12}^* \\ S_{21}^* & S_{22}^* \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$$

From the equation it can be said,

$$S_{11}S_{11}^* + S_{21}S_{21}^* = 1, \Rightarrow |S_{11}|^2 + |S_{21}|^2 = 1$$

$$S_{12}S_{12}^* + S_{22}S_{22}^* = 1, \Rightarrow |S_{12}|^2 + |S_{22}|^2 = 1$$

$$S_{11}S_{12}^* + S_{21}S_{22}^* = 0$$

$S_{12}S_{11}{}^{*} + S_{22}S_{21}{}^{*} = 0$

Last 4 equations can be written as a general form:

$$\sum_{k=1}^{N} S_{ki}S_{ki}^{*} = 1 \, and \sum_{k=1}^{N} S_{ki}S_{kj}^{*} = 0 \;\; Here \; i \neq j$$

So any Network that satisfy the generalized equation can be said as a *Lossless* Network.

AUST AHSANULLAH UNIVERSITY OF SCIENCE & TECHNOLOGY 1995

Appendix D: Detail 12 term Error Model of the VNA

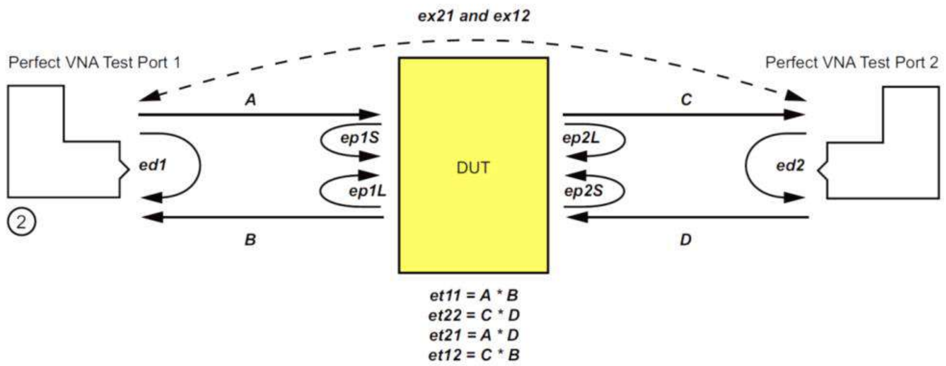


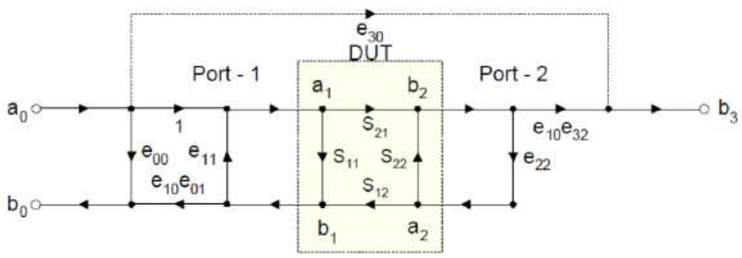
Fig d1: Simplified 12 Term error model diagram [1]

- Directivity (ed1 and ed2) describes the finite directivity of the bridges or directional couplers in the system. Partially includes some internal mismatch mechanisms that contribute to effective directivity.
- Source match (ep1S and ep2S) describes the return loss of a driving port.
- Load match (ep1L and ep2L) describes the return loss of a terminating port.
- Reflection tracking (et11 and et22) describes the frequency response of a reflect measurement including loss behaviors due to the couplers, transmission lines, converters, and other components.
- Transmission tracking (et12 and et21) is the same as above but for the transmission paths. The tracking terms are not entirely independent and this fact is used in some of the calibration algorithms.
- Isolation (ex12 and ex21) takes into account certain types of internal (non-DUT dependent) leakages that may be present in hardware. It is largely present for legacy reasons and is rarely used in practice since this type of leakage is typically very small in modern VNAs.

2 Port SOLT Calibration [1]:

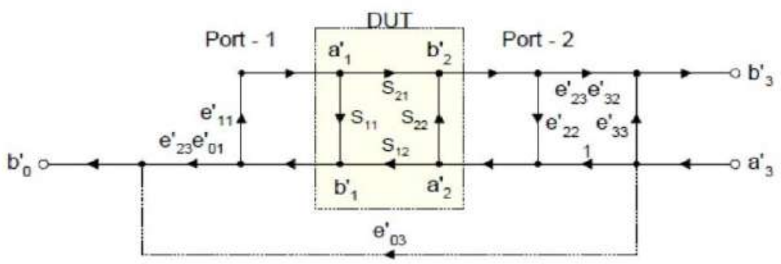
Calibration procedure consists in measuring 7 different reference standards (2 Opens, 2 Shorts, 2 matches and a Thru) with known reflection and/or transmission values from a TOSM calibration kit. In this paper reference standards are considered to have ideal values as follows

$\Gamma_{OPEN} = 1, \Gamma_{SHORT} = -1, \Gamma_{MATCH} = 0, S_{THRU} = \begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix}$



e00 : Directivity (F)
e11 : Port-1 Source Match (F)
e10e01 : Reflection Tracking (F)
e10e32 : Transmission Tracking (F)
e30 : Leakage (Crosstalk)(F)
e22 : Port-2 Load Match (F)

Fig. d2: Forward 12-term error model flow chart



e'33: Directivity (R)
e’11: Port-1 Load Match (R)
e’23e’32: Reflection Tracking (R)
e’23e’01: Transmission Tracking (R)
e’03: Leakage (Crosstalk) (R)
e’22: Port-2 Source Match (R)

Fig d3: Reverse 12-term error model flow chart

Solving measured S-parameters from figures d2 and d3

$$S_{11M} = \frac{b_0}{a_0} = e_{00} + \frac{e_{10}e_{01} \cdot (S_{11} - e_{22}\Delta_S)}{1 - e_{11}S_{11} - e_{22}S_{22} + e_{11}e_{22}\Delta_S} \quad .....(1)$$

$$S_{21M} = \frac{b_3}{a_0} = e_{30} + \frac{e_{10}e_{32} \cdot S_{21}}{1 - e_{11}S_{11} - e_{22}S_{22} + e_{11}e_{22}\Delta_S} \quad .....(2)$$

$$S_{22M} = \frac{b_3'}{a_3'} = e'_{33} + \frac{e'_{23}\, e'_{32} \cdot (S_{22} - e'_{11}\,\Delta_S)}{1 - e'_{11}\,S_{11} - e'_{22}\,S_{22} + e'_{11}\,e'_{22}\,\Delta_S} \quad .....(3)$$

$$S_{12M} = \frac{b_0'}{a_3'} = e'_{03} + \frac{e'_{23}\, e'_{01} \cdot S_{12}}{1 - e'_{11}\,S_{11} - e'_{22}\,S_{22} + e'_{11}\,e'_{22}\,\Delta_S} \quad .....(4)$$

where

SxxM: Measured, i.e. uncorrected, S-parameters

Sxx: Corrected S-parameters

$\Delta_S = S_{11}S_{22} - S_{12}S_{21}$

Port 1 Calibration: By performing Open, short and Match calibration to Port 1 the following forward error terms are calculated from eq. (1)

$$e_{00} = S_{11M}\left(match_1\right)$$

$$e_{11} = \frac{S_{11M}(open_1) + S_{11M}(short_1) - 2 \cdot e_{00}}{S_{11M}(open_1) - S_{11M}(short_1)}$$

$$e_{10}e_{01} = \frac{-2 \cdot [S_{11M}(open_1) - e_{00}] \cdot [S_{11M}(short_1) - e_{00}]}{S_{11M}(open_1) - S_{11M}(short_1)}$$

$$a_3 \qquad 1 - e'_{11}\,S_{11} - e'_{22}\,S_{22} + e'_{11}\,e'_{22}\,\Delta_S$$

$$S_{12M} = \frac{b_0'}{a_3'} = e'_{03} + \frac{e'_{23}\, e'_{01} \cdot S_{12}}{1 - e'_{11}\,S_{11} - e'_{22}\,S_{22} + e'_{11}\,e'_{22}\,\Delta_S}$$

Port 2 Calibration: By performing Open, short and Match calibration to Port 2 the following reverse error terms are calculated from eq. (3)

$$e'_{33} = S_{22M}\left(match_2\right)$$

$$e'_{11} = \frac{S_{22M}(open_2) + S_{22M}(short_2) - 2 \cdot e'_{33}}{S_{22M}(open_2) - S_{22M}(short_2)}$$

$$e'_{23}e'_{32} = \frac{-2 \cdot [S_{22M}(open_2) - e'_{33}] \cdot [S_{22M}(short_2) - e'_{33}]}{S_{22M}(open_2) - S_{22M}(short_2)}$$

Isolation Ports Calibration: Conning Load to port 1 and port 2 is optionally made only when very low transmission parameters must be measured. In most cases this error term is neglected.

$$e_{30} = S_{21M}\left(match_{1,2}\right)$$

$$e'_{03} = S_{12M}\left(match_{1,2}\right)$$

Calibration between Ports: By Connecting thru standard in-between port 1 and port 2 and Transmission Tracking error terms are calculated from (1), (2), (3) and (4) as follows

$$e_{22} = \frac{S_{11M}(Thru) - e_{00}}{S_{11M}(Thru)\cdot e_{11} - \Delta e}$$

$$e_{10}e_{32} = \left[S_{21M}(Thru) - e_{30}\right]\cdot(1 - e_{11}e_{22})$$

$$e'_{11} = \frac{S_{22M}(Thru) - e'_{33}}{S_{22M}(Thru) - \Delta e'}$$

$$e'_{23}e'_{32} = \left[S_{12M}(Thru) - e'_{03}\right]\cdot(1 - e'_{33}e'_{11})$$

where

$$\Delta e = e_{00}.e_{11} - e_{01}e_{10}$$

$$\Delta e' = e'_{33}.e'_{22} - e'_{23}e'_{32}$$

Calibration	**Reference**	**Error to be corrected**	**Description**
Port 1	Open	e_{11}	Source Match (F)
	Short	$e_{10}e_{01}$	Reflection Tracking (F)
	Match	e_{00}	Directivity (F)
Port 2	Open	e'_{11}	Source Match (R)
	Short	$e'_{23}e'_{32}$	Reflection Tracking (R)
	Match	e'_{33}	Directivity (R)
Isolation of ports	$Match_1$	e_{30}	Crosstalk (F)
	$Match_2$	e'_{03}	Crosstalk (R)
Calibration between Ports	Thru	e_{22}	Load Match (F)
		e'_{11}	Load Match (R)
		$e_{10}e_{32}$	Transmission Tracking (F)
		$e'_{23}e'_{01}$	Transmission Tracking (R)

Table: Summery of the 2-port calibration process

Solving equations (1) to (4), corrected S-parameters of the DUT can be expressed as follows

$$S_{11} = \frac{A_{11}\cdot(1 + A_{22}\cdot e'_{22}) - e_{22}\cdot A_{21}\cdot A_{12}}{D}$$

$$S_{21} = \frac{A_{21}\cdot\left[1 + A_{22}\cdot(e'_{22} - e_{22})\right]}{D}$$

$$S_{22} = \frac{A_{22}\cdot(1 + A_{11}\cdot e_{11}) - e'_{11}\cdot A_{21}\cdot A_{12}}{D}$$

$$S_{12} = \frac{A_{12}\cdot\left[1 + A_{11}\cdot(e_{11} - e'_{11})\right]}{D}$$

where

$$N_{11} = \frac{S_{11M} - e_{00}}{e_{10}e_{01}}$$

$$N_{12} = \frac{S_{12M} - e'_{03}}{e'_{23}e'_{01}}$$

$$N_{21} = \frac{S_{21M} - e_{30}}{e_{10}e_{32}}$$

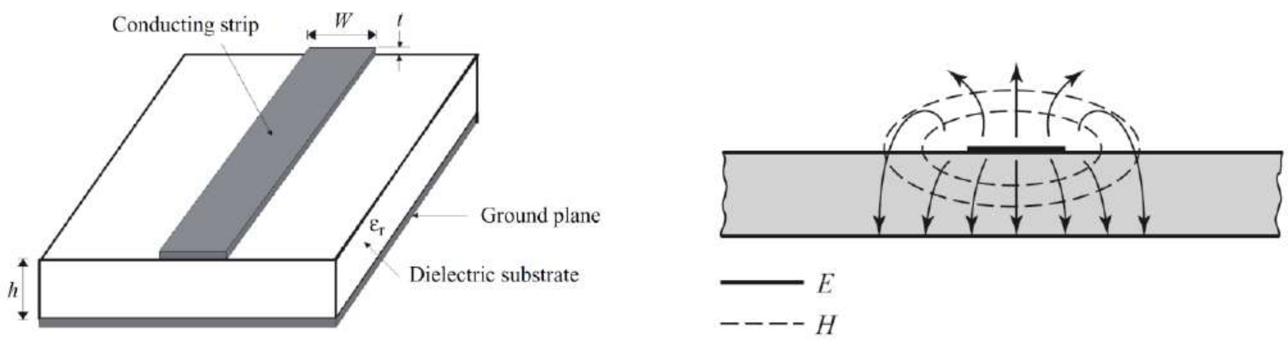
$$N_{22} = \frac{S_{22M} - e'_{33}}{e'_{23}e'_{32}}$$

$$D = (1 + A_{11}\cdot e_{11})\cdot(1 + A_{22}\cdot e'_{22}) - A_{21}\cdot A_{12}\cdot e_{22}\cdot e'_{11}$$

Reference :

1. Understanding VNA Calibration- Anritsu
2. D. C. DeGroot, K. L. Reed and J. A. Jargon, "Equivalent Circuit Models for Coaxial OSLT Standards," 54th ARFTG Conference Digest, Atlanta, GA, USA, 1999, pp. 1-13. doi: 10.1109/ARFTG.1999.3

Appendix G: Microstrip Line



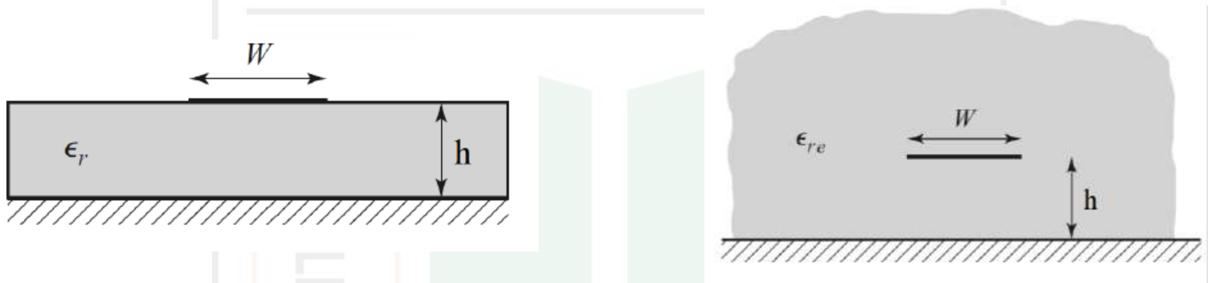
(a) (b)

Figure G1: Microstrip transmission line. (a) Geometry. (b) Electric and magnetic field lines

When the longitudinal components of the fields for the dominant mode of a microstrip line remain very much smaller than the transverse components, they may be neglected. In this case, the dominant mode then behaves like a TEM mode and the TEM transmission-line theory is applicable for the microstrip line as well. This is called the quasi-TEM approximation and is valid over most of the operating frequency ranges of Microstrip.

Effective Dielectric Constant

The effective dielectric constant can be interpreted as the dielectric constant of a homogeneous medium that equivalently replaces the **air and dielectric regions** of the Microstrip line as shown in figure bellow.



(a) (b)

Figure G2: Equivalent geometry of a quasi-TEM microstrip line. (a) Original geometry. (b) Equivalent geometry, where the dielectric substrate of relative permittivity ε_r is replaced with a homogeneous medium of effective relative permittivity ε_{re}

For very thin conductors (that is, t → 0), the closed-form expressions that provide an accuracy better than 1% are given as follows:

<u>For W/h ≤ 1</u>

$$\varepsilon_{re} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left\{\left(1 + 12\frac{h}{W}\right)^{-0.5} + 0.04\left(1 - \frac{W}{h}\right)^2\right\}$$

$$Z_c = \frac{\eta}{2\pi\sqrt{\varepsilon_{re}}}\ln\left(\frac{8h}{W} + 0.25\frac{W}{h}\right) \qquad \text{(g 1.1)}$$

$\eta = 120\pi$ is the impedance of the free space

<u>For $W/h \geq 1$</u>

$$\varepsilon_{re} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left(1 + 12\frac{h}{w}\right)^{-0.5}$$

$$Z_c = \frac{\eta}{\sqrt{\varepsilon_{re}}}\left\{\frac{W}{h} + 1.393 + 0.677\ln\left(\frac{W}{h} + 1.444\right)\right\}^{-1} \qquad \text{(g 1.2)}$$

Synthesis of W/h:

Expressions for W/h in terms of Z_c and ε_r , derived by Wheeler and Hammerstad which provide accuracy better than 1%.

<u>For $W/h \leq 2$</u>

$$\frac{W}{h} = \frac{8\exp(A)}{\exp(2A) - 2} \tag{g 2.1}$$

Here

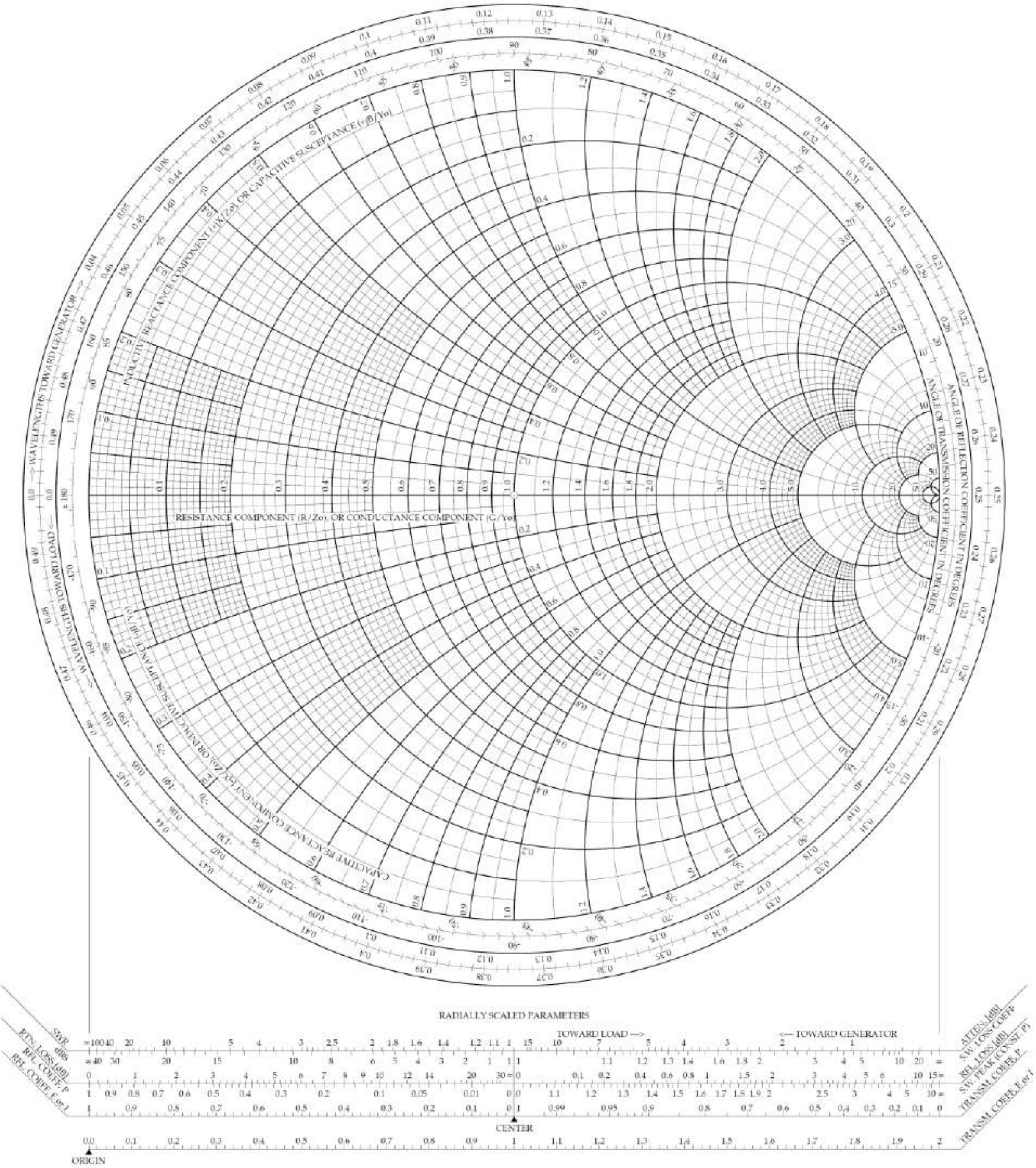
$$A = \frac{Z_c}{60}\left\{\frac{\varepsilon_r + 1}{2}\right\}^{0.5} + \frac{\varepsilon_r - 1}{\varepsilon_r - 1}\left\{0.23 + \frac{0.11}{\varepsilon_r}\right\}$$

<u>For $W/h \geq 2$</u>

$$\frac{W}{h} = \frac{2}{\pi}\left\{(B-1) - \ln(2B-1) + \frac{\varepsilon_r - 1}{2\varepsilon_r}\left[\ln(B-1) + 0.39 - \frac{0.61}{\varepsilon_r}\right]\right\} \tag{g 2.2}$$

Here,

$$B = \frac{60\pi^2}{Z_c\sqrt{\varepsilon_r}}$$


RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE (-jB/Yo)
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
ANGLE OF TRANSMISSION COEFFICIENT IN DEGREES
WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
RADIALLY SCALED PARAMETERS
TOWARD LOAD
TOWARD GENERATOR
SWR
dBS
RTN. LOSS [dB]
RFL. COEFF. P
RFL. COEFF. E or I
ATTEN. [dB]
S.W. LOSS COEFF
RFL. LOSS [dB]
S.W. PEAK (CONST. P)
TRANSM. COEFF. P
TRANSM. COEFF. E or I
CENTER
ORIGIN

TP-LINK®

2.4GHz 24dBi Grid Parabolic Antenna

TL-ANT2424B

⊙ Features:

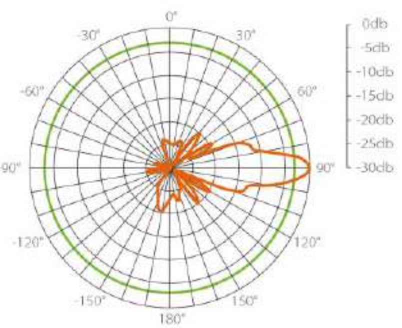
- 24dBi directional operation, ideal for extraordinary long distance point to point connection
- Weather proof design, suitable for all weather conditions
- N Female connector, applicable in most outdoor solutions

⊙ Description:

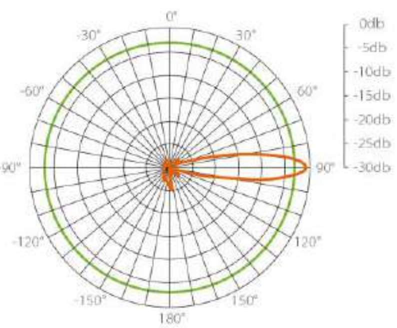
The TL-ANT2424B 24dBi directional antenna is ideal for outdoor use by attaching to your wireless access points/routers. It could be used for long distance point-to-point connection, providing your stable wireless links. It is very easy to use, no configuration or software installation required.

⊙ Radiation Patterns:

V-Plane Co-Polarization Pattern



H-Plane Co-Polarization Pattern



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2.4GHz 24dBi Grid Parabolic Antenna **TL-ANT2424B**

Specifications:

Frequency Range	2.4~2.4835GHz
Gain (Exclude Cable Loss)	24dBi
VSWR	≤1.5
HPBW/H(°)	10
HPBW/V(°)	14
F/B Ratio	>30dB
Impedance	50 Ohms
Admitted Power	100W
Interfaces	N Female (Jack)
Polarization	Vertical or Horizontal
Mounting Mast Diameter	Ø30~Ø50 mm
Mounting	Wall Mount / Pole Mount
Survival Wind Speed	216Km/hr (134Miles/hr)
Standards	RoHS, WEEE
Operating Temp.	-40℃~60℃(-40℉~140℉)
Storage Temp.	-40℃~60℃(-40℉~140℉)

Diagram:

Outdoor Point to Point Application

Simple and Convenient Installation with Excellent Transmission Range

Package:

- 24dBi Grid Parabolic Antenna
- Installation mounting kits
- User Guide

Optional Accessories:

- 54Mbps High Power Wireless Access Point TL-WA5110G
- 150Mbps Wireless N Access Point TL-WA701ND

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