EC 701: RF & MICROWAVE ENGINEERING
Contact: 3L
Credits: 3
Lectures: 41
Prepared by: Dr. Avali Banerjee

Course Objective:

1.Distinguish the RF & Microwave spectrum, Planar transmission lines and High frequency circuit elements.
2.Determine the Microwave passive components and Scattering matrix representation.
3.Illustrate the Microwave tubes, Semiconductor Microwave Devices.
4.Justify the microwave applications and typical microwave test bench.

Module II:

High frequency Circuit Elements Difference in High frequency and relatively low frequency behavior of Lumped circuit components. Miniaturization and Design of Lumped components at High RF. Realization of reactive elements as Waveguide and Planar Circuit components.

[4]

Waveguide Passive Components and their S-matrix Representation N-port networks-Properties of S matrix, Transmission matrix & their relationships; Microwave passive components and their S matrix representation: Attenuators, Phase shifter, Directional coupler, Bethe-hole coupler, Magic tee, hybrid ring, Circulators, Isolators; Design procedure of filter (maximally flat and equal ripple) using insertion loss method-specification, lowpass prototype design, scaling and conversion, implementation. **[8]**

Module II:

High frequency Circuit Elements Difference in High frequency and relatively low frequency behavior of Lumped circuit components.

Frequency-dependent behavior of passive components is one of the key concepts of RF, microwave, high-speed and all other types of high frequency design. An old saying among engineers goes, “At high frequencies, all components are R, L and C.” As application frequencies increase further, we can add that all those components are also transmission lines and antennas! This tutorial is a review of the factors that combine to create the effects referred to in these descriptive phrases. We will not try to rigorously define all the relationships—there are excellent textbooks for that—but we will try to instill an overall understanding of the types and relative magnitude of major frequency-dependent behaviors. While this discussion is limited to passive components, the general concepts certainly apply to active components as well, since their active regions are surrounded by the real-world environment that includes those “R, L, C, transmission line and antenna” elements.

Frequency-Dependence in Ideal Lumped Components

Lumped-element passive components include resistors, capacitors and inductors. AC circuit theory tells us that the latter two are reactive, where the relationships between voltage and current vary with the value of inductance or capacitance, and with frequency. The familiar XC and XL reactance formulas are shown in Figure 1.

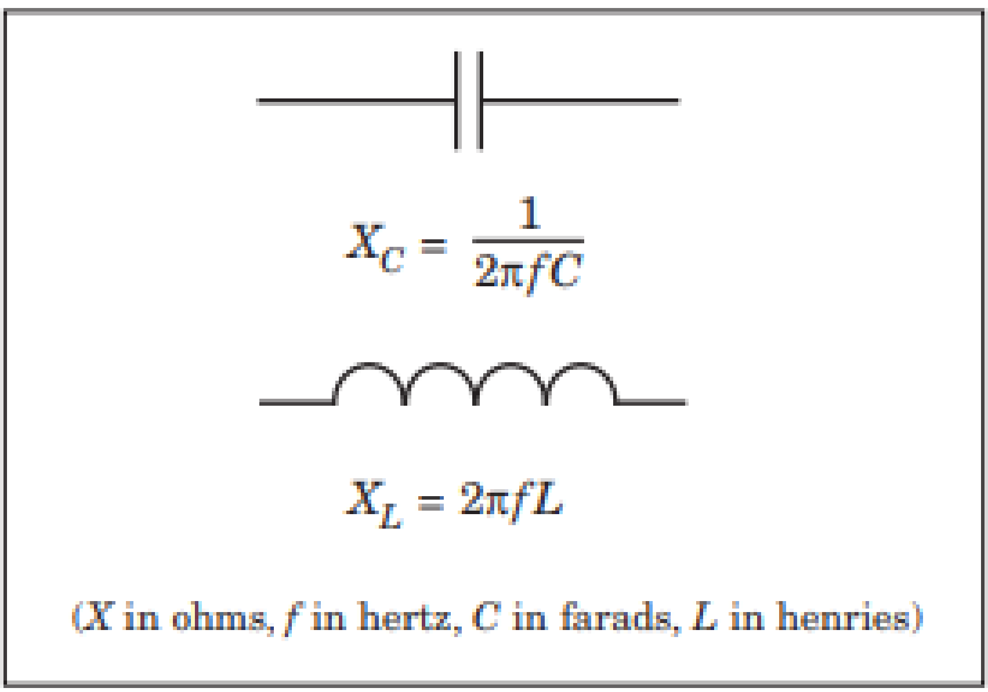


Figure 1 · The simplest frequency-dependent behaviors are capacitive and inductive reactance.

First-Order Parasitic Effects

Of course, the components we use do not behave according to simple mathematical formulas; they have size, shape, and are constructed using non-ideal materials. What I call first-order effects are the additional resistance, capacitance and inductance the result from these real-world factors. In actual resistors, capacitors and inductors, the physical length, width and height of the device, the properties of the conductors and dielectric, plus the electrodes for attachment to an external circuit are all part of the "component." A simple resistor has an inductance associated with its length and capacitance created by the two "plates" that are the end electrodes. Similarly, a capacitor has inductance associated with its length and a resistance due to losses in the dielectric that separates the plates. The additional capacitance due to the attachment electrodes is normally included in the nominal value.

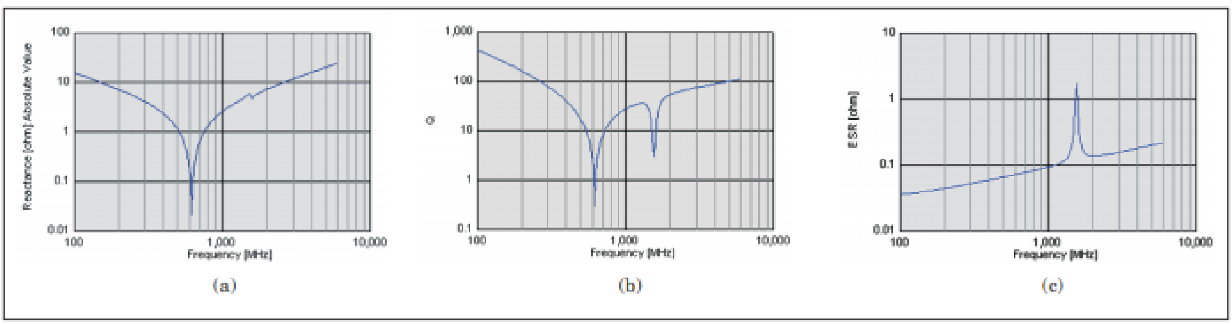


Figure 2 · Frequency-dependent behavior of a 100 pF chip capacitor: (a) reactance, (b) Q, and (c) effective series resistance (ESR).

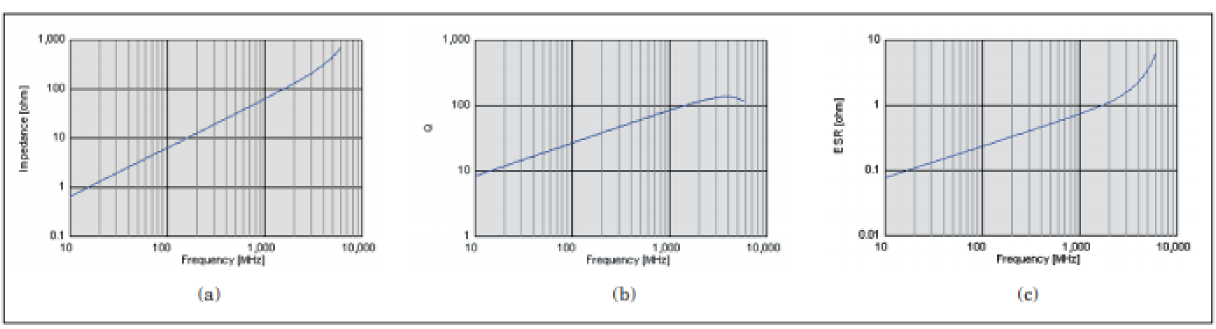


Figure 3 · Frequency-dependent behavior of a 10 nH chip inductor: (a) impedance, (b) Q, and (c) effective series resistance (ESR).

Likewise, an inductor will include the effects of its interconnecting terminals, but it has capacitance between those terminals, as well as capacitance between conductors of the winding. These first-order effects are usually the largest of all parasitic resistances and reactances, although the mounting of devices on a printed circuit board can introduce significant additional capacitance to the body of the component. This capacitance, and the added inductance of the pads and traces used to attach a component are normally considered part of the circuit, to be discussed separately from the individual components. How do these additional resistances and reactances affect the behavior of the components? In a resistor, the largest parasitic reactance is inductance. Inductive reactance is near zero at low frequencies, increasing linearly with frequency. Thus, a resistor will see an increasing inductive reactance that is in series with its resistance. Rather that try to compensate for this reactance, the best approach is to make resistors with very low inductance, which is practical to do. Usually, “non-inductive” resistors are made with bulk or thick film resistive materials rather than commonly used thin resistive films. The two main factors in a capacitor—dielectric loss resistance and series inductance—can be described in several ways. Resistance affects Q, which is a function of reactance and resistance. The series inductance creates a resonance at some high frequency where the increasing inductive reactance equals the decreasing capacitive reactance. The typical figure of merit for a capacitor at high frequencies combines these two effects as effective series resistance (ESR). Figure 2 shows how the values of reactance, Q and ESR vary with frequency. This data is for a Murata 100 pF chip capacitor in an 0805 package. Manufacturer’s data shows that this device has a series inductance of 0.66 nH. The plots of Figure 2 clearly show how capacitors become series resonant, as well having a parallel resonant point at a frequency above the series resonant frequency. When capacitors are selected for coupling or bypassing applications, it is essential to know these resonant points and be certain that the desired low reactance characteristics are maintained within the desired range of operating frequencies. Inductors are slightly more complicated, having both inter-winding capacitance and an end-to-end capacitance. Also, the wire of an inductor has more loss than the dielectric of a capacitor, so the Q values are significantly lower for this component. Figure 3 shows plots for impedance, Q and ESR for a Murata chip inductor. This component has a nominal value of 10 nH in a 0603 package, with a specified parallel capacitance of 0.04 pF. Figure 3 demonstrates the increasing reactance with frequency, and eventual parallel resonance, of inductors. Although these effects seem small in the example given, remember that large-value inductors are often used as RF chokes, and if improperly selected, may exhibit self resonance within the desired frequency range.

Transmission Line Effects

The next high frequency behavior to note is that of a transmission line. Often, components are sufficiently small that these effects are minimal, and can be ignored. But, the use of lumped element components at microwave frequencies may create significant transmission line effects.

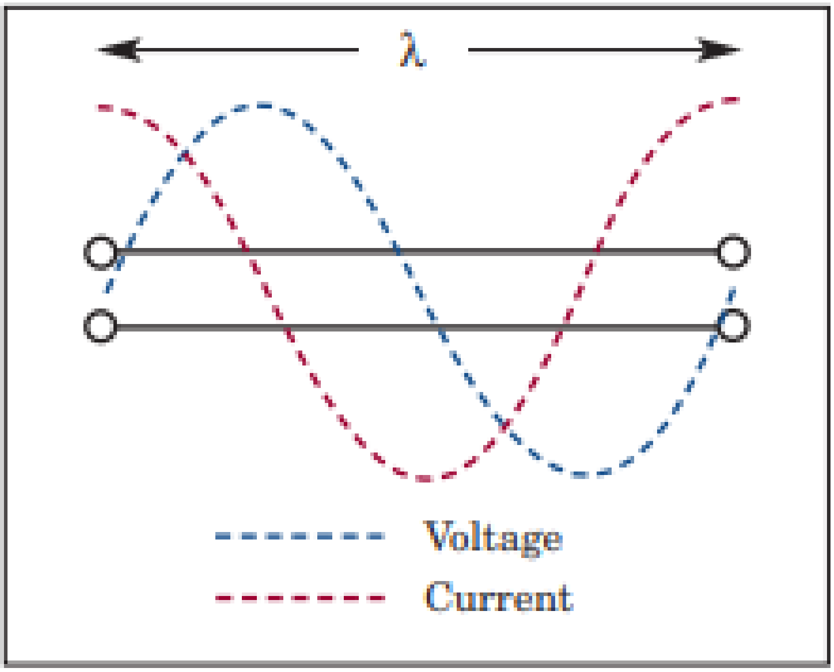


Figure 4 · Voltage and current on a one-wavelength long transmission line.

Figure 4 is a review of the voltage and phase of a signal as it travels one wavelength on a transmission line. Each goes through a full cycle, ranging from a maximum to minimum at various points along the line. Since impedance = V/I (complex values), we can see that this example has a low impedance (minimum voltage, maximum current) at each end, and a high impedance (maximum voltage, minimum current) at the points 1/4 wavelength from each end. At the center point (1/2 wavelength), the magnitude of the impedance is same as at either end, but the sign is opposite—after all, a half-wavelength delay is identical to a 180-degree “phase inversion.” At any point other than the ends and center of this line section, the impedance will be transformed to a new value. What physical dimensions will cause transmission line effects? Almost any—a 0.01 wavelength (3.6 degrees) transmission line with a characteristic impedance that is twice (or one-half) that of the system will change the magnitude of the impedance by 10 percent, and shift the reactive part in the direction of capacitance. Characteristic impedances that have a greater difference from the system impedance will have a greater effect. At 100 MHz, 0.01 wavelength is more than an inch, so common SMT components are much less than this size. Note, however, that at 1 GHz, 0.01 wavelength is nearly 1/8 inch, or about 3 mm. At this frequency, even a tiny 0201 component may have sufficient physical size to introduce an impedance change due to transmission line effect. Of special interest are inductors, with their additional conductor length in the coiled winding. This has the effect of increasing the effective length of the component as a transmission line element. Whenever transmission line effects become significant at the component level, they are certainly going to be substantial at the board level. The area surrounding the components now takes on greater importance, as pads and traces become distributed micro strip line circuit elements (transmission lines), often with poorly-defined characteristic impedances. Most lumped element components are characterized and tested in fixtures that minimize transmission line effects by maintaining low characteristic impedance and de-embedding the test fixture effects to ensure that the measured impedance is of the component alone. These test methods are not intended to hide transmission line effects; they are simply intended to provide a consistent, repeatable test methodology. The component manufacturer does not know the circuit impedance or the layout that their component will be installed in. Thus, it is up to the engineer to determine the magnitude of these effects in a circuit. Such decisions may be a subtle as deciding between an 0201 or an 0805 component based on a comparison of factors other than the nominal value. For example, is the reduction in circuit length with the smaller 0201 part most important, or does the lower characteristic transmission line impedance of the larger 0805 part result in better performance?

Electromagnetic Coupling— Components as Antennas

Our discussion has introduced effects that become significant as the frequency increases. The final characteristic in this article is electromagnetic radiation and coupling. Radiation causes losses and creates interference within, and external to, the circuit. Coupling between components may be included in the category of internal interference, or it can also be an integral part of the circuit design. Or, it can be both if not understood and controlled. A basic tutorial can barely introduce a topic as complex as electromagnetics, so we will emphasize the nature of its effects and the design issues that must be dealt with. These effects are present at low frequencies, mainly involving inductive components. The magnetic fields surrounding an inductor can couple to other inductors, or more weakly, to the parasitic inductances of other components and circuit board traces. Of course, these fields are also used intentionally to implement transformers, often using magnetic materials to enhance the coupling and help contain magnetic fields. As the frequency increases, coupling via radiation will also increase, as a result of both larger size relative to wavelength, and the fact that electromagnetic waves become more energetic with increasing frequency. For a more rigorous discussion of EM theory, I'll refer you to your favorite textbook, but here are some notes on the way EM radiation can occur in practical circuits: At frequencies below actual resonance, the efficiency of a radiator is related to its size—the length of an end-fed radiator or the area enclosed by a loop. The magnitude of the radiation is also related to the impedance at that point in the circuit. High impedances will radiate better with an end-fed (voltage-fed) "antenna," while low impedances will radiate better with a loop (current-fed) structure. Both are problematic in high frequency design, but with the common use of low impedances, loops are of special concern. A loop may consist of a component mounted above a p.c. board, with a ground plane on the back side. The area of the loop is established by the gap between the component and board, plus the thickness of the dielectric. Large components, and those that are mounted higher off the board will have larger loop area than their smaller, lower profile counterparts. The effects seem small until you try to keep signals from coupling to one another on a tightly-packed wireless handset board! Then, attention to the fine details of layout may mean the difference between acceptable performance and excessive LO radiation or CPU clock coupling into the radio. As low-cost consumer equipment has reached the 5 GHz frequency range, the use of electromagnetic simulation is imperative for successful design. These circuits typically combine traditional lumped components with distributed circuits. These two types cannot be considered separately, so accurate physical representation and high-performance EM modeling is required.

Frequency-dependent behavior of passive components, transmission line concepts and electromagnetics are the factors that have traditionally set high frequency engineering apart from other electronic specialties. Today, engineers working on nearly all electronics must consider high frequency design characteristics, whether within the equipment or in its interaction with the many high frequency/high speed devices they must work with. [1]

Miniaturization and Design of Lumped components at High RF.

In order to offer worldwide coverage and meet marketplace demand, smart phones, tablets, and other mobile devices must support more RF bands than ever and feature a growing number of wireless functions. Despite the fact that their RF circuits are becoming more complex, these devices must remain as compact—if not more compact—with each new model. With that in mind, designers are cognizant that one of the benefits of higher frequency is that components can get smaller. Specifically, antennas and inductors can shrink to surface-mount and PCB sizes. Integrated circuits, too, can contribute and have contributed to this shrink-fest as process geometries allow the integration of complete RF sections that seamlessly blend digital functionality with the analog realm of RF modulation, demodulation, antenna matching, and wave propagation.

This article looks at the miniaturization of some key components and systems that will allow designers of next-generation wireless links (hello, Internet of Things!) to provide smaller and more efficient radios and radio subsystems. In this Part 1 installment, we look at the latest passive components that integrate several discrete parts into smaller, surface-mount versions, saving space and cost and improving performance.

These components are useful with modern, highly integrated standard radio link transceiver chips that already offer small-size solutions. Since these established standards are on their third or fourth generation, integrated passives are highly optimized and include miniature front ends, filters, baluns, and other assorted parts that can serve single-standard solutions like Wi-Fi, or multi standard protocols like Wi-Fi Bluetooth combinations.

In the next context, we will look at active components such as RF transistors, mixers, modulators, and amplifiers for next-generation radios. These pieces of a radio may be useful for prototyping and even production of special function radio links that are not highly standardized. Here, SoC-style multichip modules or carrier boards can be used to more accurately preserve impedances and antenna characteristics. It should be noted that if high-volume manufacturing is to take place, more integrated custom silicon may be the best solution to reduce cost and size until the IC makers catch up. Even in this case, the components discussed here will still be useful. All parts, data sheets, tutorials, and development kits referenced here are available online at Digi-Key's website.

	Lumped L/C	**Dielectric**	**SAW**
Loss	Fair (3-5 dB)	Best (2-3 dB)	Good (2-4 dB)
Attenuation	Good	Good	Best (sharp rolloff)
Size (WxL)	Fair (200-400 mm^2)	Good (20-50 mm^2)	Best (5-14 mm^2)
Power	Watts	Several Watts	0.01-1 W
Frequency	<0.5 GHz	0.7-5 GHz	0.7-5 GHz

Table: *Monolithic filter type characteristics.*

ICs do not make the best filters. The typical inductor Q that can be fabricated using CMOS process technology is less than 10. With copper and metallization techniques, this can be doubled to around 20. The problem is that to minimize insertion loss, Q values in the hundreds are needed. Presently, only discrete filters can do this.

Filters are used throughout a radio inside and outside the silicon barrier (Refer to Figure). Here, external filters are needed in the antenna matching and signal paths to allow receive data to be stripped off and transmit data to pass with little or no attenuation. This is especially important when one antenna is used for transmit and receive, or is used with more than one protocol.

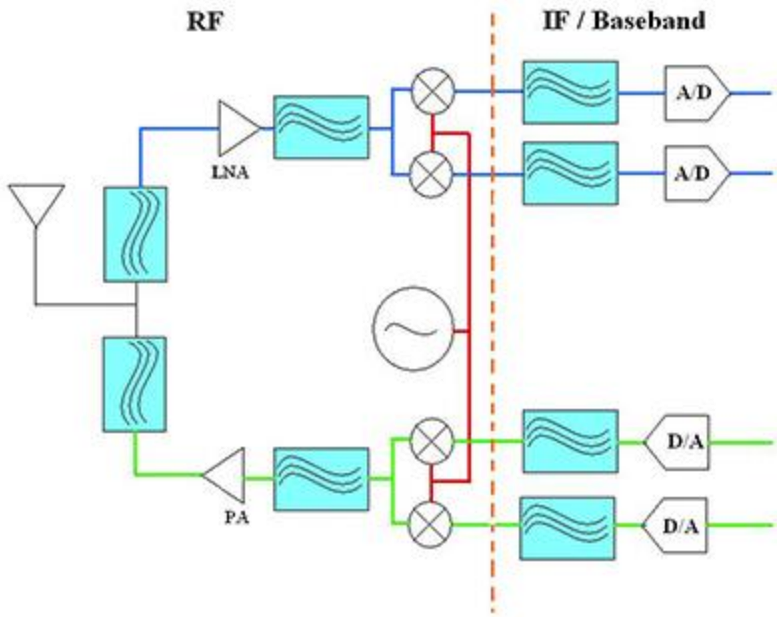


Figure: *Filters are needed extensively inside and outside the RF transceivers. Better performance of the external discretes allows internal IF and base-band filters to share a common path among different protocols and bands.[2]*

Realization of reactive elements as Waveguide and Planar Circuit components.

Planar Resonators

Modern systems require high performance filters with very low losses, small size, sharp cut-off, and high rejection at the stopband. Thus, different resonators and techniques have been introduced for RF and microwave filters to achieve different filters with this performance using the planar transmission line. The main used resonators and techniques are:

Figure 4: Two possible shapes of open-loop resonator.

1. Folded Transmission Line Resonator : It is the simplest resonator, it is basically a transmission line section that resonates when its length corresponds to half a wavelength .

2. Stepped Impedance Resonators: This type of resonators consists of high and low impedance sections, or in other words wide and narrow width transmission lines. This resonator resonates when its length corresponds to half a wavelength .

3. Open-Loop Resonators: It is a modified version of folded transmission line resonator, it is also known as U-shaped resonator and hairpin resonator, it looks like a loop which open from one side, see Fig. 4. This type of resonators has great advantages in reducing filter size, in addition, it has different coupling nature depending on the coupling sides.

The main disadvantage of these resonators is that the higher order harmonic does not allow for wide stopband.

Waveguide Passive Components and their S-matrix Representation N-port networks-Properties of S matrix, Transmission matrix & their relationships

Microwave systems consists of many Microwave components, mainly with source at one end and load at the other, which are all connected with waveguides or coaxial cable or transmission line systems.

Following are the properties of waveguides.

- High SNR
- Low attenuation
- Lower insertion loss

Waveguide Microwave Functions

Consider a waveguide having 4 ports. If the power is applied to one port, it goes through all the 3 ports in some proportions where some of it might reflect back from the same port. This concept is clearly depicted in the following figure.

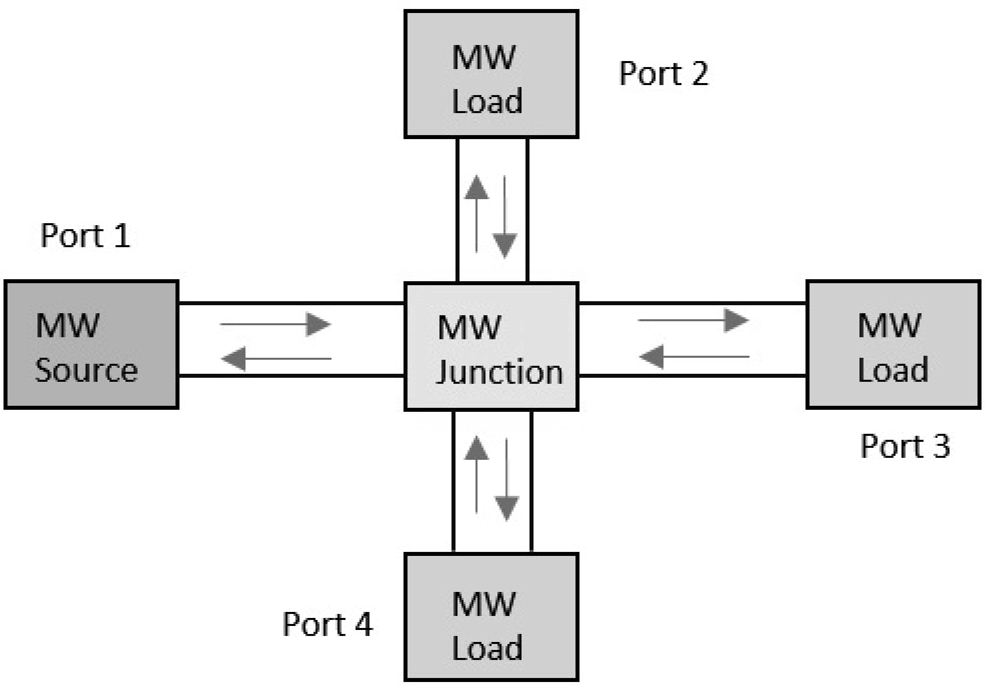


Figure :Waveguide Microwave Junction

Scattering Parameters [3]

For a two-port network, as shown in the following figure, if the power is applied at one port, as we just discussed, most of the power escapes from the other port, while some of it reflects back to the same port. In the following figure, if $\mathbf{V_1}$ or $\mathbf{V_2}$ is applied, then $\mathbf{I_1}$ or $\mathbf{I_2}$ current flows respectively.

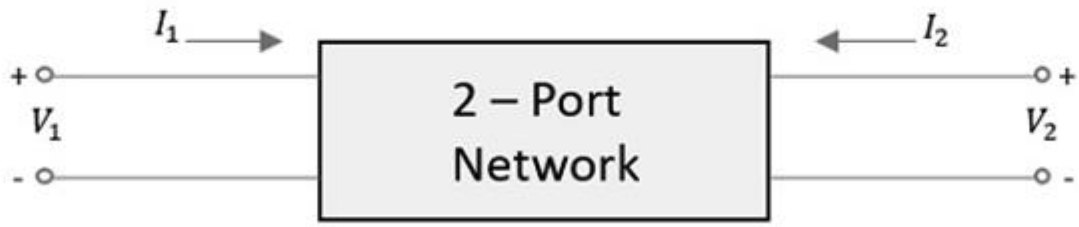


Figure: Structure of a Two Port Network

If the source is applied to the opposite port, another two combinations are to be considered. So, for a two-port network, 2 × 2 = 4 combinations are likely to occur.

The travelling waves with associated powers when scatter out through the ports, the Microwave junction can be defined by S-Parameters or Scattering Parameters, which are represented in a matrix form, called as "Scattering Matrix".

Scattering Matrix

The scattering matrix of an m-port junction is a square matrix of a set of elements which relate incident and reflected waves at the port of the junction. The diagonal elements of the s-matrix represents reflection coefficients and off diagonal elements represent transmission coefficients.

Characteristics of s-matrix:

- It describes any passive microwave component.
- It exists for linear passive and time invariant networks.
- It gives complete information on reflection and transmission coefficients.
-

It is a square matrix which gives all the combinations of power relationships between the various input and output ports of a Microwave junction. The elements of this matrix are called "Scattering Coefficients" or "Scattering (S) Parameters".

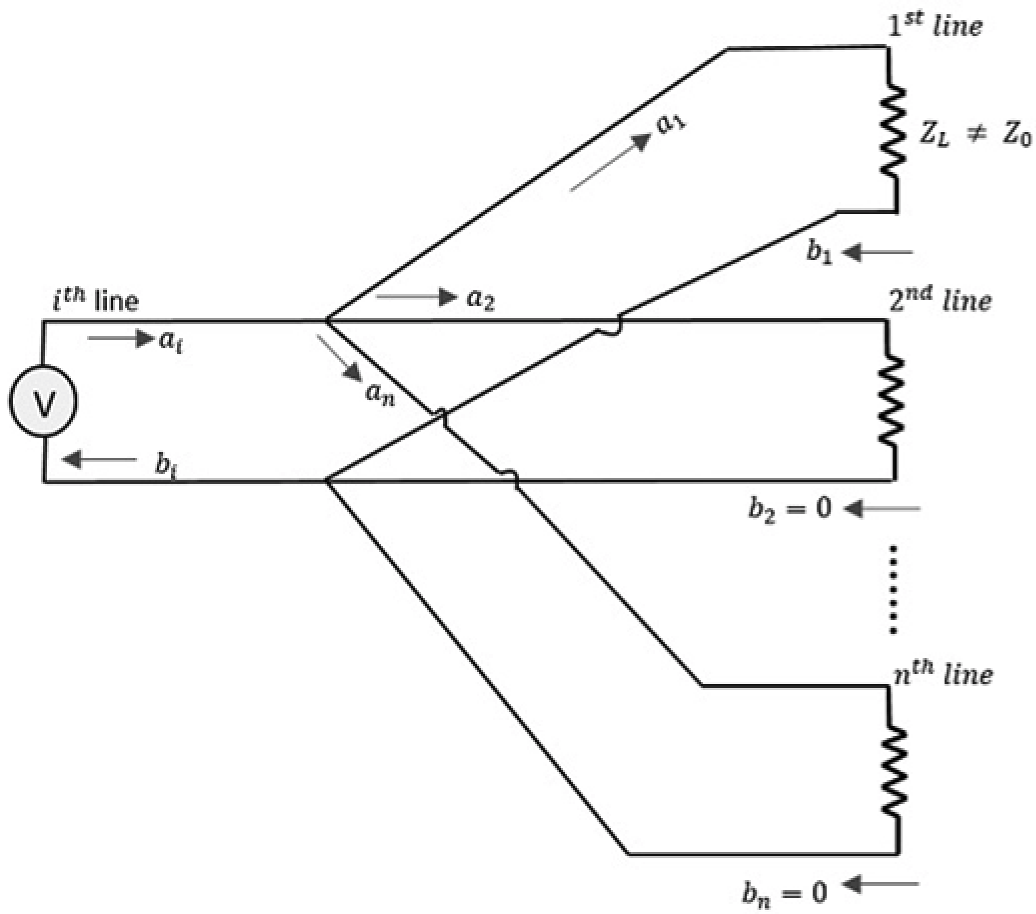


Figure:

Here, the source is connected through ithith line while a1a1 is the incident wave and b1b1 is the reflected wave.
If a relation is given between b1 and a1,

$$b_1 = (reflection\ coefficient)a_1 = S_{1i}a_1$$

Where

- S1i = Reflection coefficient of 1st line (where ii is the input port and 1is the output port)
- 1 = Reflection from 1st line
- ii = Source connected at i^{th} line

If the impedance matches, then the power gets transferred to the load. Unlikely, if the load impedance doesn't match with the characteristic impedance. Then, the reflection occurs. That means, reflection occurs if

$$Z_l \neq Z_o$$

However, if this mismatch is there for more than one port, example 'n' ports, then i=1 to n (since i can be any line from 1 to n).
Therefore, we have

$$b_1 = S_{11}a_1 + S_{12}a_2 + S_{13}a_3 + \ldots\ldots\ldots\ldots\ldots + S_{1n}a_n$$

$$b_2 = S_{21}a_1 + S_{22}a_2 + S_{23}a_3 + \ldots\ldots\ldots\ldots\ldots + S_{2n}a_n$$

$$\vdots$$

$$b_n = S_{n1}a_1 + S_{n2}a_2 + S_{n3}a_3 + \ldots\ldots\ldots\ldots\ldots + S_{nn}a_n$$

When this whole thing is kept in a matrix form,

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ . \\ . \\ . \\ b_n \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} & S_{13} & \dots & S_{1n} \\ S_{21} & S_{22} & S_{23} & \dots & S_{2n} \\ . & . & . & \dots & . \\ . & . & . & \dots & . \\ . & . & . & \dots & . \\ S_{n1} & S_{n2} & S_{n3} & \dots & S_{nn} \end{bmatrix} \times \begin{bmatrix} a_1 \\ a_2 \\ a_3 \\ . \\ . \\ . \\ a_n \end{bmatrix}$$

Column matrix $[b]$ **Scattering matrix** $[S]$ **Matrix** $[a]$

The column matrix [b] corresponds to the reflected waves or the output, while the matrix [a] corresponds to the incident waves or the input. The scattering column matrix [s] which is of the order of n×n contains the reflection coefficients and transmission coefficients. Therefore,

[b] = [S] [a]

Properties of [S] Matrix

The scattering matrix is indicated as [S] matrix. There are few standard properties for [S] matrix. They are −

- [S] is always a square matrix of order (nxn)
  $[S]_{n\times n}$
- [S] is a symmetric matrix
  i.e., $S_{ij}=S_{ji}$
- [S] is a unitary matrix
  i.e., $[S][S]^*=I$
- The sum of the products of each term of any row or column multiplied by the complex conjugate of the corresponding terms of any other row or column is zero. i.e.,

$$\sum_{i=j}^{n} S_{ik} S_{ik}^{*} = 0 \; for \; k \neq j$$

$$(k = 1, 2, 3, \ldots \; n) \; and \; (j = 1, 2, 3, \ldots \; n)$$

- If the electrical distance between some kth port and the junction is $\beta_k l_k$, then the coefficients of S_{ij} involving k, will be multiplied by the factor $e^{-j\beta_k l_k}$

Transmission Parameters:

Scattering transfer parameters, like scattering parameters, are defined in terms of incident and reflected waves. The difference is that *T*-parameters relate the waves at port 1 to the waves at port 2 whereas *S*-parameters relate the reflected waves to the incident waves. In this respect *T*-parameters fill the same role as *ABCD* parameters and allow the *T*-parameters of cascaded networks to be calculated by matrix multiplication of the component networks. *T*-parameters, like *ABCD* parameters, can also be called transmission parameters.

T-parameters are not so easy to measure directly unlike *S*-parameters. However, *S*-parameters are easily converted to *T*-parameters,

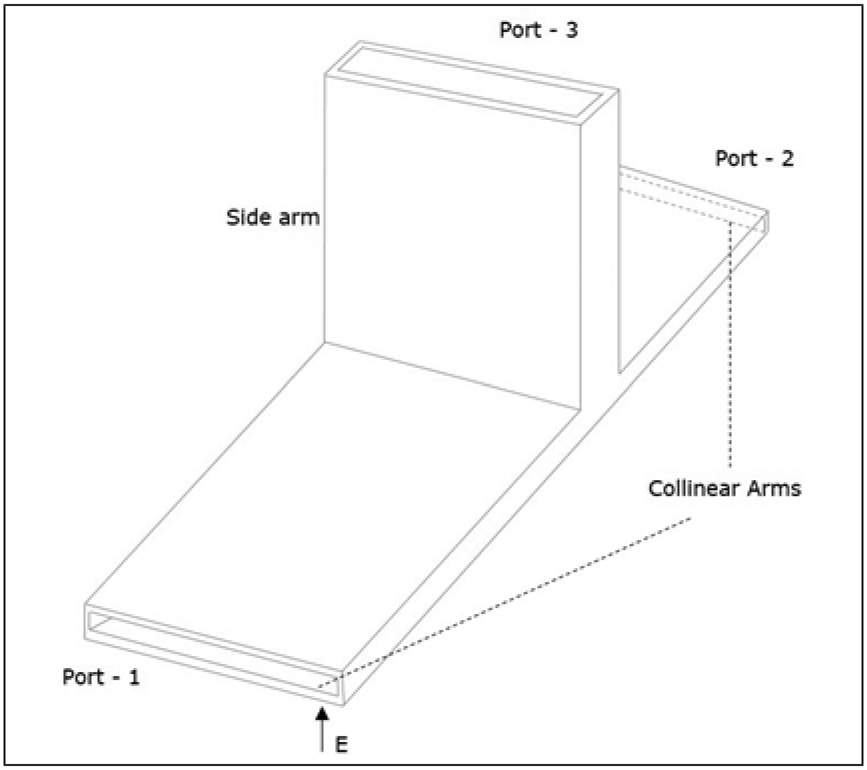
$$\begin{bmatrix} a_1 \\ b_1 \end{bmatrix} = \begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} \begin{bmatrix} b_2 \\ a_2 \end{bmatrix}$$

E-Plane Tee

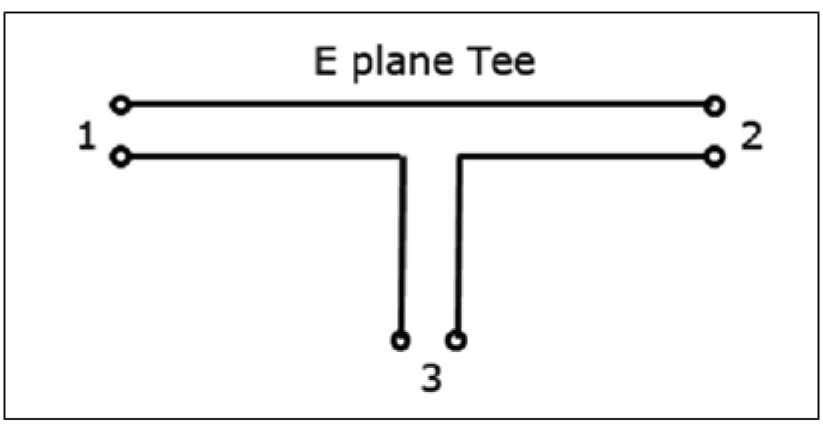
An E-Plane Tee junction is formed by attaching a simple waveguide to the broader dimension of a rectangular waveguide, which already has two ports. The arms of rectangular waveguides make two ports called collinear portsi.e., Port1 and Port2, while the new one, Port3 is called as Side arm or E-arm. T his E-plane Tee is also called as Series Tee.

As the axis of the side arm is parallel to the electric field, this junction is called E-Plane Tee junction. This is also called as Voltage or Series junction.

The ports 1 and 2 are 180° out of phase with each other. The cross-sectional details of E-plane tee can be understood by the following figure.



The following figure shows the connection made by the sidearm to the bi-directional waveguide to form the parallel port.



Properties of E-Plane Tee

The properties of E-Plane Tee can be defined by its [S]3x3 matrix.
It is a 3×3 matrix as there are 3 possible inputs and 3 possible outputs.

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{21} & S_{22} & S_{23} \\ S_{31} & S_{32} & S_{33} \end{bmatrix}$$ ……………………………………………………………(1)

Scattering coefficients S13S13 and S23S23 are out of phase by 180° with an input at port 3.

$S_{23} = -S_{13}$ ……………………………………………………………………………………(2)

The port is perfectly matched to the junction.

$S_{33} = 0$ ……………………………………………………………………………………(3)

From the symmetric property,

$S_{ij} = S_{ji}$

$S_{12} = S_{21}$; $S_{23} = S_{32}$; $S_{13} = S_{31}$ ……………………………(4)

Considering equations 3 & 4, the [S] matrix can be written as,

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{12} & S_{22} & -S_{13} \\ S_{13} & -S_{13} & 0 \end{bmatrix}$$ ……………………………………(5)

We can say that we have four unknowns, considering the symmetry property.

From the Unitary property

$$[S][S]* = [I]$$

$$\begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{12} & S_{22} & -S_{13} \\ S_{13} & -S_{13} & 0 \end{bmatrix} \begin{bmatrix} S_{11}^* & S_{12}^* & S_{13}^* \\ S_{12}^* & S_{22}^* & -S_{13}^* \\ S_{13}^* & -S_{13}^* & 0 \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$$

Multiplying we get,

(Noting R as row and C as column)

$$R_1C_1 : S_{11}S_{11}^* + S_{12}S_{12}^* + S_{13}S_{13}^* = 1$$
$$|S_{11}|^2 + |S_{11}|^2 + |S_{11}|^2 = 1$$
$$R_2C_2 : |S_{12}|^2 + |S_{22}|^2 + |S_{13}|^2 = 1$$
$$R_3C_3 : |S_{13}|^2 + |S_{13}|^2 = 1$$
$$R_3C_1 : S_{13}S_{11}^* - S_{13}S_{12}^* = 1$$

....................................(6)

From this we get

$S_{11} = S_{22}$

Also we get

$$2|S_{13}|^2 \quad or \quad S_{13} = \frac{1}{\sqrt{2}}$$

$$S_{13}\left(S_{11}^* - S_{12}^*\right)$$

Or $S_{11} = S_{12} = S_{22}$

Using all these we again get

$$|S_{11}|^2 + |S_{11}|^2 + \frac{1}{2} = 1$$
$$2|S_{11}|^2 = \frac{1}{2}$$
Or $S_{11} = \frac{1}{2}$

Substituting the values from the above equations in [S] matrix,
We get,

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

We know that [b] = [S][a]

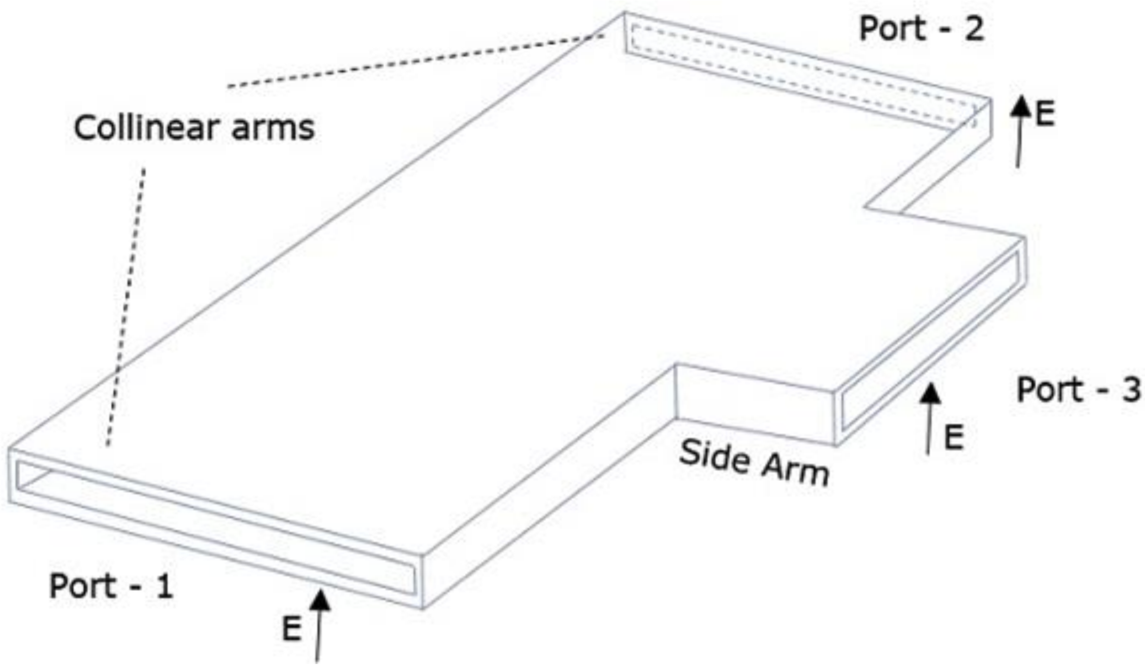
$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \end{bmatrix} = \begin{bmatrix} \frac{1}{2} & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ \frac{1}{2} & \frac{1}{2} & -\frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & -\frac{1}{\sqrt{2}} & 0 \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ a_3 \end{bmatrix}$$

This is the scattering matrix for E-Plane Tee, which explains its scattering properties.[4]

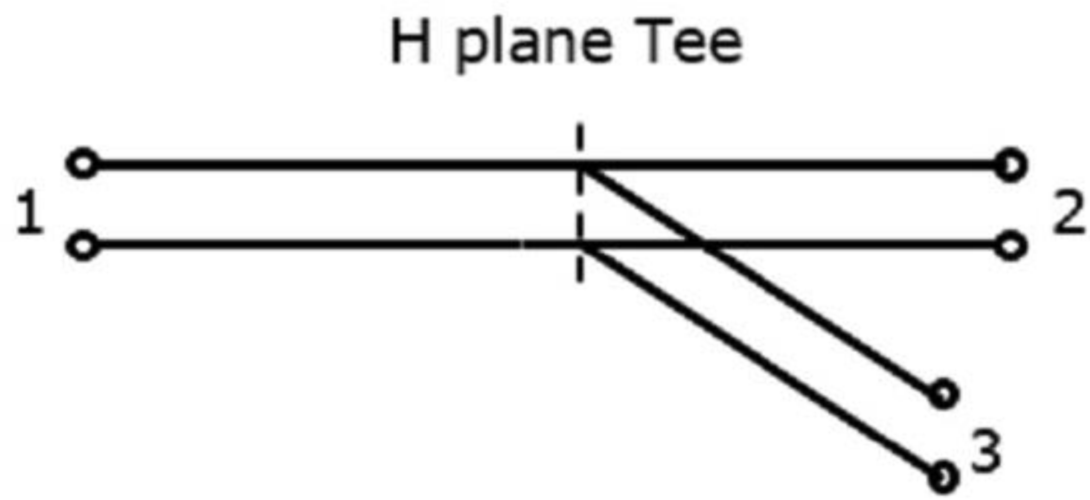
H-Plane Tee

An H-Plane Tee junction is formed by attaching a simple waveguide to a rectangular waveguide which already has two ports. The arms of rectangular waveguides make two ports called collinear ports i.e., Port1 and Port2, while the new one, Port3 is called as Side arm or H-arm. This H-plane Tee is also called as Shunt Tee.

As the axis of the side arm is parallel to the magnetic field, this junction is called H-Plane Tee junction. This is also called as Current junction, as the magnetic field divides itself into arms. The cross-sectional details of H-plane tee can be understood by the following figure.



The following figure shows the connection made by the sidearm to the bi-directional waveguide to form the serial port.



Properties of H-Plane Tee

The properties of H-Plane Tee can be defined by its $[S]_{3\times3}$ matrix.

It is a 3×3 matrix as there are 3 possible inputs and 3 possible outputs.

$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{21} & S_{22} & S_{23} \\ S_{31} & S_{32} & S_{33} \end{bmatrix}$$

Scattering coefficients S13S13 and S23S23 are equal here as the junction is symmetrical in plane.

From the symmetric property,

Sij=Sji

S12=S21 ; S23=S32=S13 ; S13=S31

The port is perfectly matched

S33=0

Now, the [S] matrix can be written as,

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

We can say that we have four unknowns, considering the symmetry property.

From the Unitary property

$$[S][S]* = [I]$$

$$\begin{bmatrix} S_{11} & S_{12} & S_{13} \\ S_{12} & S_{22} & S_{13} \\ S_{13} & S_{13} & 0 \end{bmatrix} \begin{bmatrix} S_{11}^* & S_{12}^* & S_{13}^* \\ S_{12}^* & S_{22}^* & S_{13}^* \\ S_{13}^* & S_{13}^* & 0 \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$$

Multiplying we get,

(Noting R as row and C as column)

$$R_1C_1 : S_{11}S_{11}^* + S_{12}S_{12}^* + S_{13}S_{13}^* = 1$$
$$|S_{11}|^2 + |S_{12}|^2 + |S_{13}|^2 = 1$$
$$R_2C_2 : |S_{12}|^2 + |S_{22}|^2 + |S_{13}|^2 = 1$$
$$R_3C_3 : |S_{13}|^2 + |S_{13}|^2 = 1$$
$$R_3C_1 : S_{13}S_{11}^* - S_{13}S_{12}^* = 0$$
$$2|S_{13}|^2 = 1 \quad or \quad S_{13} = \frac{1}{\sqrt{2}}$$
$$|S_{11}|^2 = |S_{22}|^2$$
$$S_{11} = S_{22}$$
$$S_{13}\left(S_{11}^* + S_{12}^*\right) = 0$$
$$S_{13} \neq 0, S_{11}^* + S_{12}^* = 0, \ or \ S_{11}^* = -S_{12}^*$$
$$S_{11} = -S_{12} \ or \ S_{12} = -S_{11}$$

Since,

$$S_{13} \neq 0, S_{11}^* + S_{12}^* = 0, \ or \ S_{11}^* = -S_{12}^*$$
$$|S_{11}|^2 + |S_{11}|^2 + \frac{1}{2} = 1 \quad or \quad 2|S_{11}|^2 = \frac{1}{2} \quad or \quad S_{11} = \frac{1}{2}$$

Now, we get

$$S_{12} = -\frac{1}{2}$$
$$S_{22} = \frac{1}{2}$$

Substituting for S13, S11, S12 and S22

We get,

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

We know that

$$[b] = [s][a]$$

$$\begin{bmatrix} b_1 \\ b_2 \\ b_3 \end{bmatrix} = \begin{bmatrix} \frac{1}{2} & -\frac{1}{2} & \frac{1}{\sqrt{2}} \\ -\frac{1}{2} & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & 0 \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ a_3 \end{bmatrix}$$

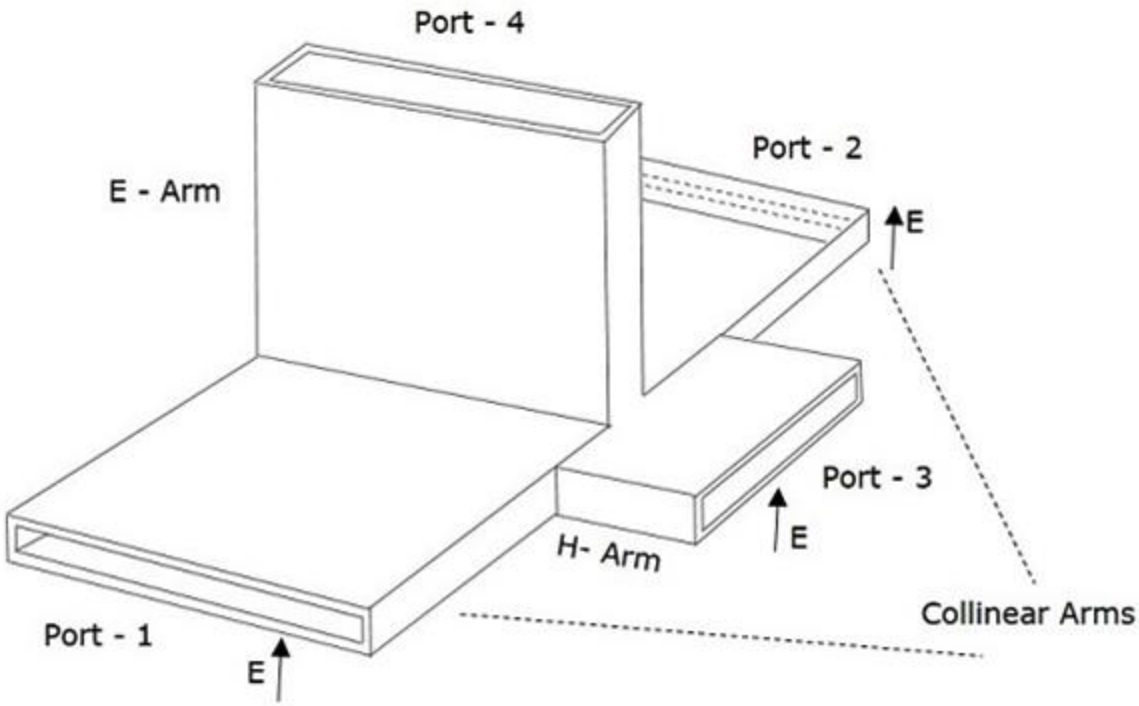
This is the scattering matrix for H-Plane Tee, which explains its scattering properties.[5]

Magic Tee

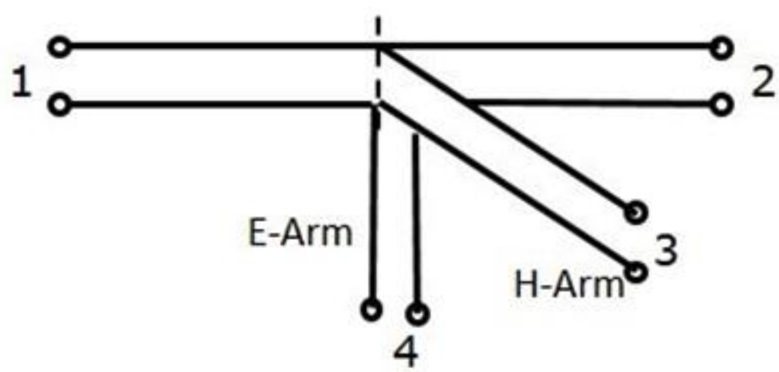
An E-H Plane Tee junction is formed by attaching two simple waveguides one parallel and the other series, to a rectangular waveguide which already has two ports. This is also called as Magic Tee, or Hybrid or 3dB coupler.

The arms of rectangular waveguides make two ports called collinear portsi.e., Port 1 and Port 2, while the Port 3 is called as H-Arm or Sum port or Parallel port. Port 4 is called as E-Arm or Difference port or Series port.

The cross-sectional details of Magic Tee can be understood by the following figure.



The following figure shows the connection made by the side arms to the bi-directional waveguide to form both parallel and serial ports.



Characteristics of E-H Plane Tee

- If a signal of equal phase and magnitude is sent to port 1 and port 2, then the output at port 4 is zero and the output at port 3 will be the additive of both the ports 1 and 2.
- If a signal is sent to port 4, (E-arm) then the power is divided between port 1 and 2 equally but in opposite phase, while there would be no output at port 3. Hence, S34 = 0.

- If a signal is fed at port 3, then the power is divided between port 1 and 2 equally, while there would be no output at port 4. Hence, S43 = 0.
- If a signal is fed at one of the collinear ports, then there appears no output at the other collinear port, as the E-arm produces a phase delay and the H-arm produces a phase advance. So, S12 = S21 = 0.

Properties of E-H Plane Tee

- The properties of E-H Plane Tee can be defined by its [S]4×4matrix.
- It is a 4×4 matrix as there are 4 possible inputs and 4 possible outputs.

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

From Unitary property, [S][S]∗=[I]

$$\begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{12} & S_{22} & S_{13} & -S_{14} \\ S_{13} & S_{13} & 0 & 0 \\ S_{14} & -S_{14} & 0 & 0 \end{bmatrix} \begin{bmatrix} S_{11}^* & S_{12}^* & S_{13}^* & S_{14}^* \\ S_{12}^* & S_{22}^* & S_{13}^* & -S_{14}^* \\ S_{13} & S_{13} & 0 & 0 \\ S_{14} & -S_{14} & 0 & 0 \end{bmatrix}$$
$$= \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix}$$

The scattering matrix of Magic Tee

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

We already know that, [b] = [S][a]

Rewriting the above, we get

$$\begin{vmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \end{vmatrix} = \begin{bmatrix} 0 & 0 & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ 0 & 0 & \frac{1}{2} & -\frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & 0 & 0 \\ \frac{1}{\sqrt{2}} & -\frac{1}{\sqrt{2}} & 0 & 0 \end{bmatrix} \begin{vmatrix} a_1 \\ a_2 \\ a_3 \\ a_4 \end{vmatrix}$$

Applications of E-H Plane Tee

Some of the most common applications of E-H Plane Tee are as follows −

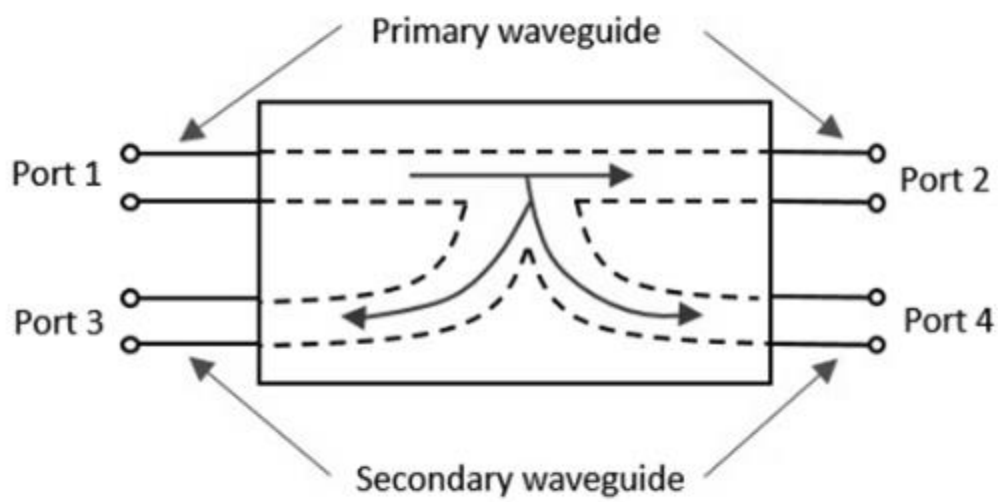
- E-H Plane junction is used to measure the impedance − A null detector is connected to E-Arm port while the Microwave source is connected to H-Arm port. The collinear ports together with these ports make a bridge and the impedance measurement is done by balancing the bridge.
- E-H Plane Tee is used as a duplexer − A duplexer is a circuit which works as both the transmitter and the receiver, using a single antenna for both purposes. Port 1 and 2 are used as receiver and transmitter where they are isolated and hence will not interfere. Antenna is connected to E-Arm port. A matched load is connected to H-Arm port, which provides no reflections. Now, there exists transmission or reception without any problem.
- E-H Plane Tee is used as a mixer − E-Arm port is connected with antenna and the H-Arm port is connected with local oscillator. Port 2 has a matched load which has no reflections and port 1 has the mixer circuit, which gets half of the signal power and half of the oscillator power to produce IF frequency.

In addition to the above applications, an E-H Plane Tee junction is also used as Microwave bridge, Microwave discriminator, etc.[6]

Directional Couplers

A Directional coupler is a device that samples a small amount of Microwave power for measurement purposes. The power measurements include incident power, reflected power, VSWR values, etc.

Directional Coupler is a 4-port waveguide junction consisting of a primary main waveguide and a secondary auxiliary waveguide. The following figure shows the image of a directional coupler.



Directional coupler is used to couple the Microwave power which may be unidirectional or bi-directional.

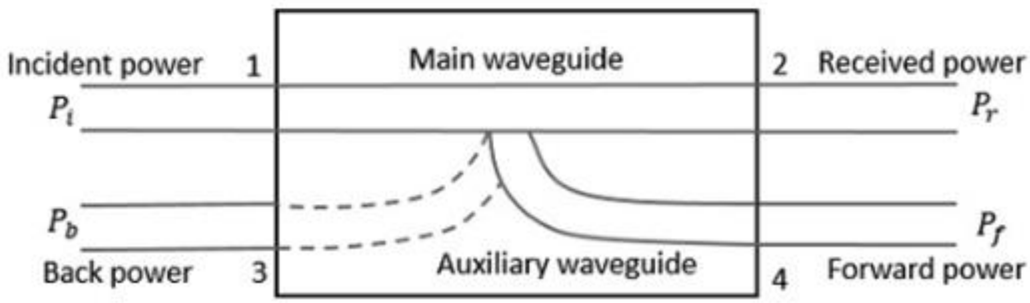
Properties of Directional Couplers

The properties of an ideal directional coupler are as follows.

- All the terminations are matched to the ports.
- When the power travels from Port 1 to Port 2, some portion of it gets coupled to Port 4 but not to Port 3.
- As it is also a bi-directional coupler, when the power travels from Port 2 to Port 1, some portion of it gets coupled to Port 3 but not to Port 4.
- If the power is incident through Port 3, a portion of it is coupled to Port 2, but not to Port 1.

- If the power is incident through Port 4, a portion of it is coupled to Port 1, but not to Port 2.
- Port 1 and 3 are decoupled as are Port 2 and Port 4.

Ideally, the output of Port 3 should be zero. However, practically, a small amount of power called back power is observed at Port 3. The following figure indicates the power flow in a directional coupler.



Where

- Pi = Incident power at Port 1
- Pr = Received power at Port 2
- Pf = Forward coupled power at Port 4
- Pb = Back power at Port 3

Following are the parameters used to define the performance of a directional coupler.

Coupling Factor (C)

The Coupling factor of a directional coupler is the ratio of incident power to the forward power, measured in dB.

$$C = 10\,log_{10}\frac{P_i}{P_f}dB$$

Directivity (D)

The Directivity of a directional coupler is the ratio of forward power to the back power, measured in dB.

$$D = 10\,log_{10}\frac{P_f}{P_b}dB$$

Isolation

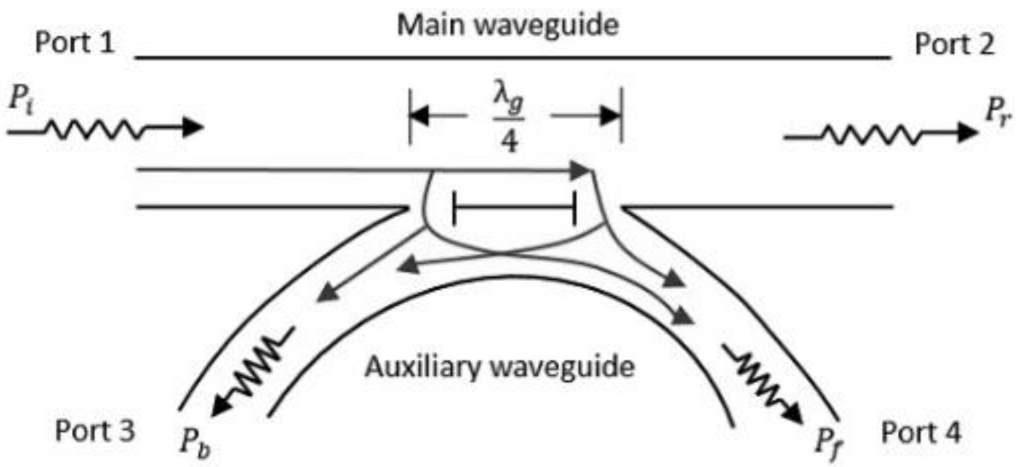
It defines the directive properties of a directional coupler. It is the ratio of incident power to the back power, measured in dB.

$$I = 10\,log_{10}\frac{P_i}{P_b}dB$$

Isolation in dB = Coupling factor + Directivity

Two-Hole Directional Coupler

This is a directional coupler with same main and auxiliary waveguides, but with two small holes that are common between them. These holes are λg/4distance apart where λg is the guide wavelength. The following figure shows the image of a two-hole directional coupler.



Two-hole directional coupler

A two-hole directional coupler is designed to meet the ideal requirement of directional coupler, which is to avoid back power. Some of the power while travelling between Port 1 and Port 2, escapes through the holes 1 and 2.

The magnitude of the power depends upon the dimensions of the holes. This leakage power at both the holes are in phase at hole 2, adding up the power contributing to the forward power P_f. However, it is out of phase at hole 1, cancelling each other and preventing the back power to occur.

Hence, the directivity of a directional coupler improves.

Bethe-hole directional coupler

One of the most common, and simplest, waveguide directional couplers is the Bethe-hole directional coupler. This consists of two parallel waveguides, one stacked on top of the other, with a hole between them. Some of the power from one guide is launched through the hole into the other. The Bethe-hole coupler is another example of a backward coupler.

The concept of the Bethe-hole coupler can be extended by providing multiple holes. The holes are spaced $\lambda/4$ apart. The design of such couplers has parallels with the multiple section coupled transmission lines. Using multiple holes allows the bandwidth to be extended by designing the sections as a Butterworth, Chebyshev, or some other filter class. The hole size is chosen to give the desired coupling for each section of the filter. Design criteria are to achieve a substantially flat coupling together with high directivity over the desired band.

Waveguide Joints

As a waveguide system cannot be built in a single piece always, sometimes it is necessary to join different waveguides. This joining must be carefully done to prevent problems such as − Reflection effects, creation of standing waves, and increasing the attenuation, etc.

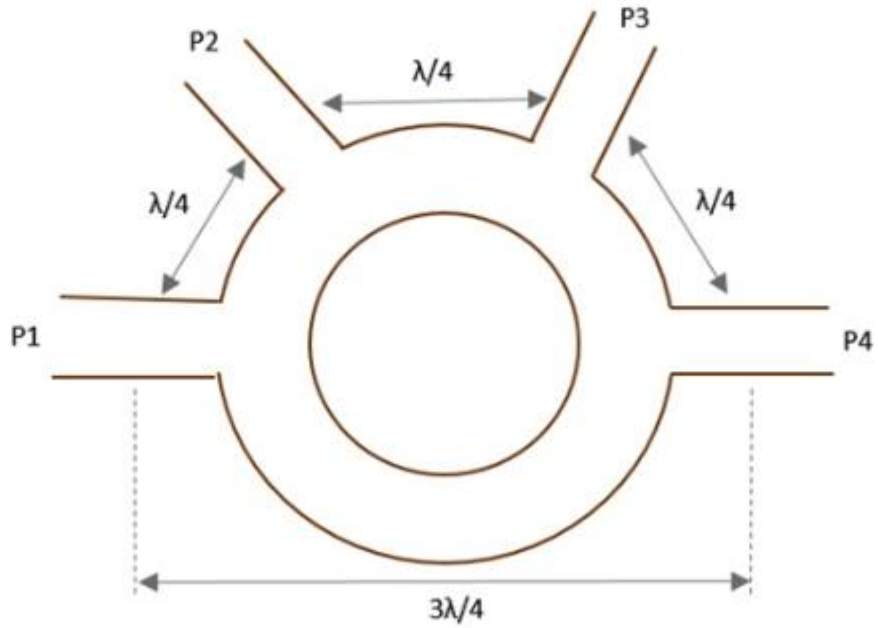
The waveguide joints besides avoiding irregularities, should also take care of E and H field patterns by not affecting them. There are many types of waveguide joints such as bolted flange, flange joint, choke joint, etc.[7]

Hybrid Ring/ Rat-race Junction

This microwave device is used when there is a need to combine two signals with no phase difference and to avoid the signals with a path difference.

A normal three-port Tee junction is taken and a fourth port is added to it, to make it a rat race junction. All of these ports are connected in angular ring forms at equal intervals using series or parallel junctions.

The mean circumference of total race is 1.5λ and each of the four ports are separated by a distance of λ/4. The following figure shows the image of a Rat-race junction.



Rat-race Coupler

Let us consider a few cases to understand the operation of a Rat-race junction.

Case 1

If the input power is applied at port 1, it gets equally split into two ports, but in clockwise direction for port 2 and anti-clockwise direction for port 4. Port 3 has absolutely no output.

The reason being, at ports 2 and 4, the powers combine in phase, whereas at port 3, cancellation occurs due to λ/2 path difference.

Case 2

If the input power is applied at port 3, the power gets equally divided between port 2 and port 4. But there will be no output at port 1.

Case 3

If two unequal signals are applied at port 1 itself, then the output will be proportional to the sum of the two input signals, which is divided between port 2 and 4. Now at port 3, the differential output appears.

The Scattering Matrix for Rat-race junction is represented as

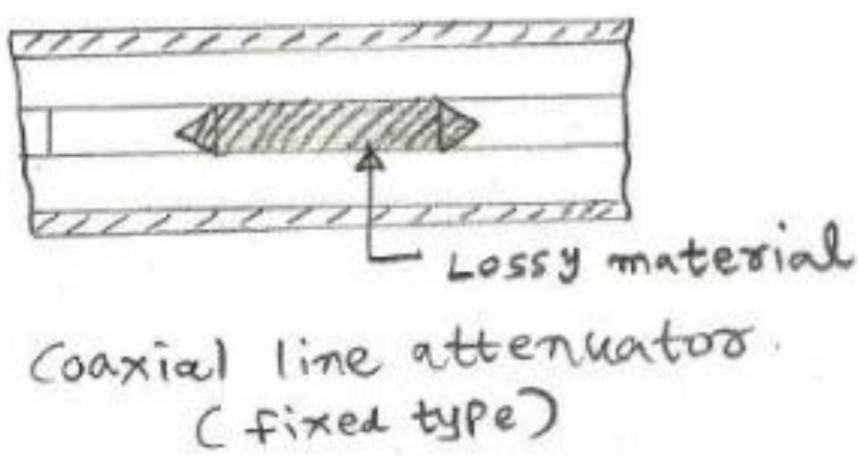
$$[S] = \begin{bmatrix} 0 & S_{12} & 0 & S_{14} \\ S_{21} & 0 & S_{23} & 0 \\ 0 & S_{32} & 0 & S_{34} \\ S_{41} & 0 & S_{43} & 0 \end{bmatrix}$$

Applications

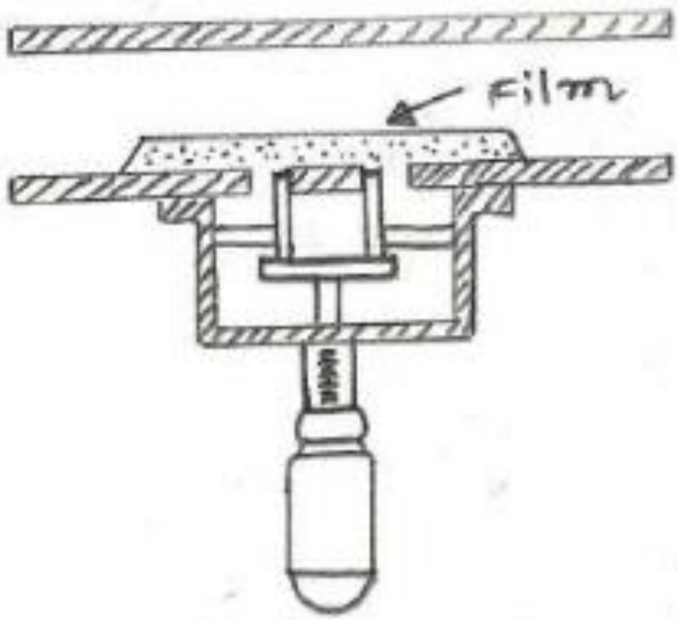
Rat-race junction is used for combining two signals and dividing a signal into two halves.[8]

Attenuator:

The attenuators are basically passive devices which control power levels in microwave system by absorpsion of the signal. Attenuator which attenuates the RF signal in a waveguide system is referred as waveguide attenuator. There are two main types fixed and variable. They are achieved by insertion of resistive films. [9]



This figure depicts coaxial line based fixed type of attenuator. Here resistive film is fixed at the centre conductor which absorb the power and as a result of power loss and hence microwave signal gets attenuated. This is referred as coaxial line attenuator.



This figure depicts waveguide based fixed type of attenuator. Here thin dielectric strip with coated resistive film is placed at the centre of waveguide. Film is placed in the waveguide parallel to the maximum E field.

In a variable type of waveguide attenuator, resistive vane is moved from one side of the wall to the centre by using screw where E field is considered to be maximum. This resistive film is shaped to give linear attenuation variation.[10]

Phase Shifter:

Phase shifters can be used to translate the frequency of an RF carrier by subjecting it to a linear time varying phase shift. An IQ Vector Modulator is an RF or microwave circuit which has the ability to control both the amplitude and phase of the transmitted signal simultaneously.

The phase shifter routes the microwave signal that is supplied to each radiating element through cables of varying length. The cables delay the wave, thereby shifting the relative phase of the output. The illustration shows the three basic delays each phase shifter can introduce.

Circulator:

A circulator is a passive, non-reciprocal three- or four-port device, in which a microwave or radio-frequency signal entering any port is transmitted to the next port in rotation (only). A *port* in this context is a point where an external waveguide or transmission line (such as a microstrip line or a coaxial cable), connects to the device. For a three-port circulator, a signal applied to port 1 only comes out of port 2; a signal applied to port 2 only comes out of port 3; a signal applied to port 3 only comes out of port 1, so to up to a phase-factor, the scattering matrix for an ideal three-port circulator is

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

Isolator:

An isolator is a two-port device that transmits microwave or radio frequency power in one direction only. It is used to shield equipment on its input side, from the effects of conditions on its output side; for example, to prevent a microwave source being detuned by a mismatched load.

An isolator is a non-reciprocal device, with a non-symmetric scattering matrix. An ideal isolator transmits all the power entering port 1 to port 2, while absorbing all the power entering port 2, so that to within a phase-factor its S-matrix is

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

To achieve non-reciprocity, an isolator must necessarily incorporate a non-reciprocal material. At microwave frequencies this material is invariably a ferrite which is biased by a static magnetic field. The ferrite is positioned within the isolator such that the microwave signal presents it with a rotating magnetic field, with the rotation axis aligned with the direction of the static bias field. The behaviour of the ferrite depends on the sense of rotation with respect to the bias field, and hence is different for microwave signals travelling in opposite directions. Depending on the exact operating conditions, the signal travelling in one direction may either be phase-shifted, displaced from the ferrite or absorbed.

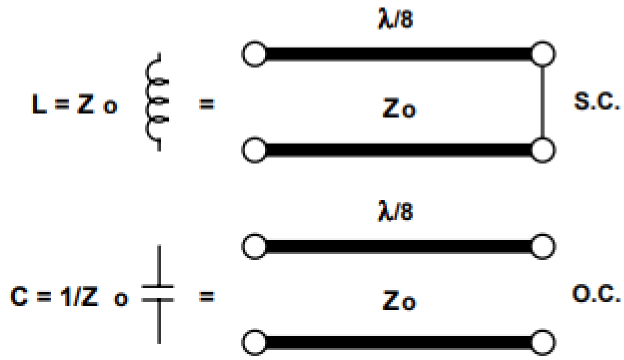
Design procedure of filter (maximally flat and equal ripple) using insertion loss method-specification, lowpass prototype design, scaling and conversion, implementation.

Passbands and Stopbands in Periodic Structures:

Periodic structures generally exhibit passband and stopband characteristics in various bands of wave number determined by the nature of the structure. This was originally studied in the case of waves in crystalline lattice structures, but the results are more general. The presence or absence of propagating wave can be determined by inspection of the k-β or ω-β diagram. For our purposes it's enough to know the generality that periodic structures give rise to bands that are passed and bands that are stopped. To construct specific filters, we'd like to be able to relate the desired frequency characteristics to the parameters of the filter structure. The general synthesis of filters proceeds from tabulated low-pass prototypes. Ideally, we can relate the distributed parameters to the corresponding parameters of lumped element prototypes. As we will see, the various forms of filter passband and stopband can be realized in distributed filters as well as in lumped element filters. Over the years, lumped element filters have been developed that are non-minimum phase; that is, the phase characteristics are not uniquely determined by the amplitude characteristics. This technique permits the design of filters for communications systems that could not be constructed using only minimum phase filter concepts. This generally requires coupling between multiple sections, and can be extended to distributed filters.

Richard's Transformation and Kuroda's Identities (λ/8 Lines):

Richard's Transformation and Kuroda's Identities focus on uses of λ/8 lines, for which X = jZo. Richard's idea is to use variable Zo (width of microstrip, for example) to create lumped elements from transmission lines. A lumped low-pass prototype filter can be implemented using λ/8 lines of appropriate Zo to replace lumped L and C elements. So if we need an inductance of L for a prototype filter normalized to cutoff frequency ωc = 1 and admittance go = 1, we can substitute a λ/8 transmission line stub that has Zo = L. The last step of the filter design will be to scale the design to the desired ωc and Zo (typically 50Ω).



The λ/8 transmission line sections are called commensurate lines, since they are all the same length in a given filter. Kuroda's idea is use the λ/8 line of appropriate Zo to transform awkward or unrealizable elements to those with more tractable values and geometry. As an example, the series inductive stub in the diagram here can be replaced by a shunt capacitive stub on the other end of the λ/8 line, with different values of characteristic impedance determined by

$k = n^2 = 1+[Z1/ Z2]$

Low Pass Filter Using Stubs:

The prototype lowpass LC structure employs series inductors, so a direct conversion to transmission line stubs by Richard's transformation would result in series stubs. However, we can use the Kuroda identity for series inductors to create a structure that has only series transmission line sections and shunt open stubs. In order to do this we must be aware that we should begin by adding unit elements (λ/8 transmission lines of Zo = 1) at each end of the filter, so that there will be structures that are of the form of the Kuroda identities.

The filter is designed by the following steps:

• Lumped element low pass prototype (from tables, typically)

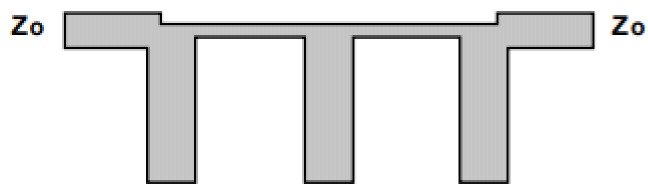
• Convert series inductors to series stubs, shunt capacitors to shunt stubs

• Add l/8 lines of Zo = 1 at input and output

• Apply Kuroda identity for series inductors to obtain equivalent with shunt open stubs with λ/8 lines between them

• Transform design to 50Ω and fc to obtain physical dimensions (all elements are λ/8).

The completed filter in microstrip form looks like this:



Although the lumped element filter has only lowpass response, the periodic nature of distributed elements results in harmonic responses because of the periodic nature of the structure. In this case, there are responses centered on 4nfc, where n can take any positive integral value. This is a general characteristic of distributed-element filters, and can be a problem if unanticipated in system designs. This general characteristic of periodic structures can be a serious problem in the application of distributed-element filters in microwave systems. It is often an overlooked artifact of filters constructed entirely from transmission line elements.[11]

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. Of course, such filters do not exist in practice, so compromises must be made; herein lies the art of filter design. The image parameter method of the previous section may yield a usable filter response for some applications, but there is no methodical way of improving the design. The insertion loss method, however, allows a high degree of control over the passband and stopband amplitude and phase characteristics, with a systematic way to synthesize a desired response. The necessary design tradeoffs can be evaluated to best meet the application requirements. If, for example, a minimum insertion loss is most important, a binomial response could be used; a Chebyshev response would satisfy a requirement for the sharpest cutoff. If it is possible to sacrifice the attenuation rate, a better phase response can be obtained by using a linear phase filter design. In addition, in all cases, the insertion loss method allows filter performance to be improved in a straightforward manner, at the expense of a higher order filter. For the filter prototypes to be discussed below, the order of the filter is equal to the number of reactive elements.

In the insertion loss method a filter response is defined by its insertion loss, or power loss ratio, P_{LR}:

$$P_{\mathrm{LR}} = \frac{\text{Power available from source}}{\text{Power delivered to load}} = \frac{P_{\mathrm{inc}}}{P_{\mathrm{load}}} = \frac{1}{1 - |\Gamma(\omega)|^2}.$$

Observe that this quantity is the reciprocal of $|S_{12}|^2$ if both load and source are matched. The insertion loss (IL) in dB is

IL = 10 log P_{LR}

In general insertion loss is defined as the ratio of the power delivered to load when connected directly to the generator to the power delivered when the filter is inserted. It has been seen that a realizable filter has a power loss ratio of the form given by:

$$PLR = 1 + \left[\frac{M(\omega^2)}{N(\omega^2)} \right]$$

where M (ω2) and N (ω2) are polynomials in the frequency domain.. There are many different types of polynomials that result in good filter response and each type has its own set of characteristics. In this work Chebyshev response has been considered to demonstrate the different design technique. Design of a filter using the Insertion-Loss approach usually begins by designing a normalized low-pass prototype (LPP). The LPP is a low-pass filter with source and load resistance of 1Ω and cut-off frequency of 1 rad/sec. Impedance transformation and frequency scaling are then applied to renormalize the LPP and synthesized. To demonstrate the conventional technique the following technical specifications are considered:

Order of filter N = 5

RS = RL = 50Ω

Cut-off frequency = 2.5 GHz

Pass band ripple = 0.01dB

The element values for Chebyshev filters for desired pass band ripple for LPP circuit has been given as

$g_0 = 1$ $g_1 = 0.7563$ $g_2 = 1.3049$ $g_3 = 1.5773$

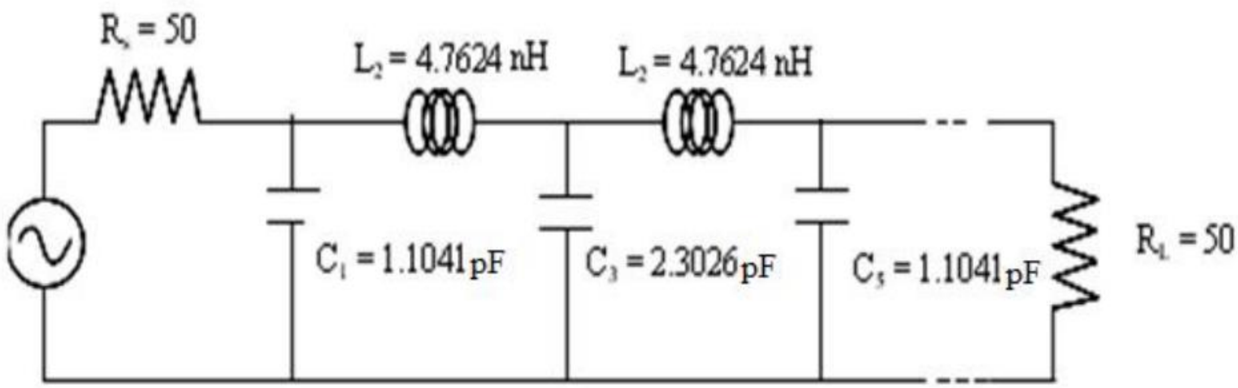
$g_4 = 1.3049$ $g_5 = 0.7563$ $g_6 = 1.$

By applying the frequency transformation along with the impedance transformation, the desired L-C ladder network is obtained. The transformed values from the prototype of corresponding inductance 'L' and capacitance 'C' can be calculated for the desired filter from following expressions given in as below :

$$L = \left(\frac{\Omega_c}{\omega_c}\right) Z_0 g \quad \text{; g representing the inductance}$$

$$C = \left(\frac{\Omega_c}{\omega_c}\right) \frac{g}{Z_0} \quad \text{; g representing the capacitance}$$

Where Ωc is cut off frequency of LPP, ωc desired cut off frequency, and Z0 is load and source impedance. The transformed filter network with its elemental values has been shown in figure



Rs = 50Ω

C1 = 1.1041 pf

L2 = 4.7624 nh

C3 = 2.3026 pf

L4 = 4.7624 nh

C5 = 1.1041 pf

RL = 50Ω

The above results represent the lumped parameters which have to be transformed into a distributed network through transformation methods.

Microstrip stub filters implemented through Richards transformation is one of the ways of designing low pass microwave filters.

Microwave Measurement Devices:

Among the Microwave measurement devices, a setup of Microwave bench, which consists of Microwave devices has a prominent place. This whole setup, with few alternations, is able to measure many values like guide wavelength, free space wavelength, cut-off wavelength, impedance, frequency, VSWR, Klystron characteristics, Gunn diode characteristics, power measurements, etc.

The output produced by microwaves, in determining power is generally of a little value. They vary with the position in a transmission line. There should be an equipment to measure the Microwave power, which in general will be a Microwave bench setup.

Microwave Test Bench General Measurement Setup

This setup is a combination of different parts which can be observed in detail. The following figure clearly explains the setup.

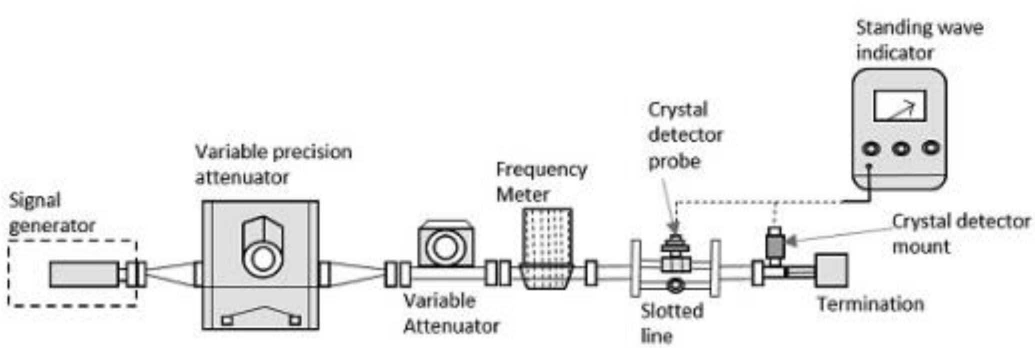


Figure: Microwave Test bench

Signal Generator

As the name implies, it generates a microwave signal, in the order of a few milliwatts. This uses velocity modulation technique to transfer continuous wave beam into milliwatt power.

A Gunn diode oscillator or a Reflex Klystron tube could be an example for this microwave signal generator.

Precision Attenuator

This is the attenuator which selects the desired frequency and confines the output around 0 to 50db. This is variable and can be adjusted according to the requirement.

Variable Attenuator

This attenuator sets the amount of attenuation. It can be understood as a fine adjustment of values, where the readings are checked against the values of Precision Attenuator.

Isolator

This removes the signal that is not required to reach the detector mount. Isolator allows the signal to pass through the waveguide only in one direction.

Frequency Meter

This is the device which measures the frequency of the signal. With this frequency meter, the signal can be adjusted to its resonance frequency. It also gives provision to couple the signal to waveguide.

Crystal Detector

A crystal detector probe and crystal detector mount are indicated in the above figure, where the detector is connected through a probe to the mount. This is used to demodulate the signals.

Standing Wave Indicator

The standing wave voltmeter provides the reading of standing wave ratio in dB. The waveguide is slotted by some gap to adjust the clock cycles of the signal. Signals transmitted by waveguide are forwarded through BNC cable to VSWR or CRO to measure its characteristics.

A microwave bench set up in real-time application would look as follows −

Figure: Microwave Test bench(Practical Example)

Now, let us take a look at the important part of this microwave bench, the slotted line.

Slotted Line

In a microwave transmission line or waveguide, the electromagnetic field is considered as the sum of incident wave from the generator and the reflected wave to the generator. The reflections indicate a mismatch or a discontinuity. The magnitude and phase of the reflected wave depends upon the amplitude and phase of the reflecting impedance.

The standing waves obtained are measured to know the transmission line imperfections which is necessary to have a knowledge on impedance mismatch for effective transmission. This slotted line helps in measuring the standing wave ratio of a microwave device.

Construction

The slotted line consists of a slotted section of a transmission line, where the measurement has to be done. It has a travelling probe carriage, to let the probe get connected wherever necessary, and the facility for attaching and detecting the instrument.

In a waveguide, a slot is made at the center of the broad side, axially. A movable probe connected to a crystal detector is inserted into the slot of the waveguide.

Operation

The output of the crystal detector is proportional to the square of the input voltage applied. The movable probe permits convenient and accurate measurement at its position. But, as the probe is moved along, its output is proportional to the standing wave pattern, which is formed inside the waveguide. A variable attenuator is employed here to obtain accurate results.

The output VSWR can be obtained by

$$VSWR = \sqrt{\frac{V_{max}}{V_{min}}}$$

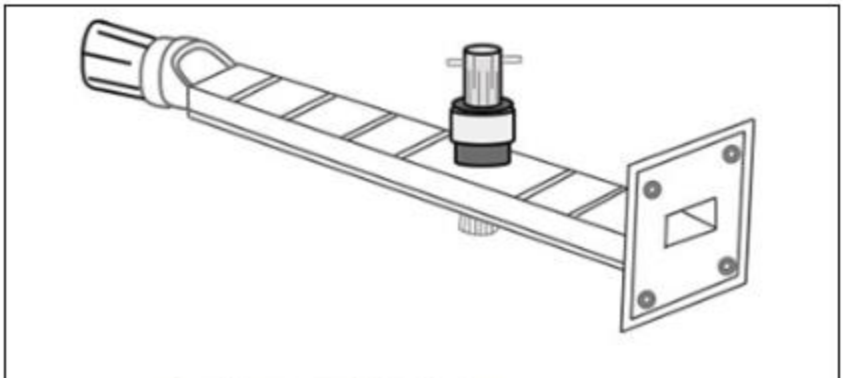
Where, V is the output voltage.
The following figure shows the different parts of a slotted line labelled.

In order to obtain a low frequency modulated signal on an oscilloscope, a slotted line with a tunable detector is employed. A slotted line carriage with a tunable detector can be used to measure the following.

- VSWR (Voltage Standing Wave Ratio)
- Standing wave pattern
- Impedance
- Reflection coefficient
- Return loss
- Frequency of the generator used

Tunable Detector

The tunable detector is a detector mount which is used to detect the low frequency square wave modulated microwave signals. The following figure gives an idea of a tunable detector mount.



Tunable waveguide detector

The following image represents the practical application of this device. It is terminated at the end and has an opening at the other end just as the above one.

To provide a match between the Microwave transmission system and the detector mount, a tunable stub is often used. There are three different types of tunable stubs.

- Tunable waveguide detector
- Tunable co-axial detector
- Tunable probe detector

Also, there are fixed stubs like −

- Fixed broad band tuned probe

- Fixed waveguide matched detector mount

The detector mount is the final stage on a Microwave bench which is terminated at the end. [13]

Reference

[1] http://www.highfreqelec.summittechmedia.com/Jun06/HFE0606_Tutorial.pdf

[2] https://www.digikey.in/en/articles/techzone/2015/feb/smaller-is-better-miniaturizing-rf-part-1

[3] https://www.tutorialspoint.com/microwave_engineering/microwave_engineering_microwave_devices.htm

[4] https://www.tutorialspoint.com/microwave_engineering/microwave_engineering_eplane_tee.htm

[5] https://www.tutorialspoint.com/microwave_engineering/microwave_engineering_hplane_tee.htm

[6] https://www.tutorialspoint.com/microwave_engineering/microwave_engineering_eh_plane_tee.htm

[7] https://www.tutorialspoint.com/microwave_engineering/microwave_engineering_directional_couplers.htm

[8] https://www.tutorialspoint.com/microwave_engineering/microwave_engineering_ratrace_junction.htm

[9] https://www.google.com/search?q=Microwave+attenuator&rlz=1C1CHBD_enIN753IN753&oq=Microwave+attenuator&aqs=chrome..69i57.5488j0j8&sourceid=chrome&ie=UTF-8

[10] http://www.rfwireless-world.com/Terminology/waveguide-microwave-attenuator.html

[11] http://home.sandiego.edu/~ekim/e194rfs01/filterek.pdf

[12] https://shodhganga.inflibnet.ac.in/bitstream/10603/28302/9/09_chapter%202.pdf

[13]https://www.tutorialspoint.com/microwave_engineering/microwave_engineering_measurement_devices.htm