CHAPTER 4
Microwave System

This chapter covers various aspects of the microwave system. This includes the design, development of: i) a power supply for energizing the coaxial magnetron source, which is used as a source of up to 2 kW microwave power at 2.45 GHz frequency in continuous wave mode, ii) the design, development and characterization of the microwave transfer line which consists of an isolator (which allows the microwave propagation in forward direction while the reflected power is dumped), iii) a directional coupler for power measurement, iv) a tunable three-stub tuner for impedance matching, v) a microwave vacuum window for vacuum isolation between the plasma chamber and microwave system, vi) a glass-water load for the characterization of the microwave line and the source, and vii) a microwave launcher to feed the microwave power to the plasma chamber.

The microwave system consists of a magnetron as a microwave generator, its power supply, waveguide components (an isolator with dummy load, a directional coupler, a three-stub tuner, a high voltage break, a microwave vacuum window, and a microwave launcher) for transferring microwave power from the microwave source to plasma chamber. This type of microwave systems at 2.45 GHz, 2 kW continuous wave have been widely used in various plasma applications to deliver the microwave power to the plasma chamber [79,80,81]. The microwave system at frequency 2.45 GHz and power up to 2 kW continuous wave have been designed, and developed using rectangular waveguide WR-284. A schematic diagram of the microwave system for RRCAT-ECRIS is shown in Fig. 4.1. A 3D-view of the microwave system test setup for RRCAT-ECRIS is shown in Fig. 4.2.
4.1 Magnetron as a Source of Microwave Power

As a source of microwave power, we have used standard coaxial magnetron (Model: NL10250L, Make: M/s Richardson Electronics Pvt. Ltd., USA) suitably water-cooled for operating in continuous wave. It is a low cost and efficient device for producing microwave radiation. It is widely used in microwave ovens (frequency: 2.45 GHz) for cooking, and in plasma production for various applications. The technical specifications of the coaxial magnetron are given in Table 4.1.
The magnetron requires a low voltage, high current AC power supply, floated at negative high voltage for the filament (thoriated tungsten coil, directly heated by AC or DC) and a high voltage negative and low current DC power supply. The body of the magnetron is at ground potential. A magnetron can be operated in both continuous wave, and pulsed mode using a suitable pulse-forming network [82,83,84,85,86,87]. To energize this magnetron, a high voltage DC power supply of rating negative (-) 5 kV, 1 A for cathode, and an AC power supply of rating 5 V, 20 A for filament (floating), were developed in-house. A block diagram of the magnetron power supply is shown in Fig. 4.3. The power supply employs a voltage doubler scheme to achieve full voltage 5 kV DC with a 2.5 kV transformer. The power supply was tested independently, and characterized with a DC resistive load up to full rating. The output waveform of the magnetron power supply was measured using a storage oscilloscope, and a high voltage probe. The measured waveform and ripple of the magnetron power supply for a) DC output for cathode, and b) AC output for filament are shown in Fig. 4.4. The measured output voltage was (-) 4.7 kV DC and ripple peak to peak was 0.1 %.

### Table 4.1: The technical specifications of the coaxial magnetron.

<table>
<thead>
<tr>
<th>Description</th>
<th>Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency</td>
<td>2.45 ± 0.025 GHz</td>
</tr>
<tr>
<td>Filament voltage</td>
<td>4.6 V AC</td>
</tr>
<tr>
<td>Filament current</td>
<td>19 A</td>
</tr>
<tr>
<td>Cathode voltage</td>
<td>4 kV DC</td>
</tr>
<tr>
<td>Cathode current</td>
<td>725 mA</td>
</tr>
<tr>
<td>Output power</td>
<td>2 kW CW</td>
</tr>
</tbody>
</table>

The output voltage was stable for the required period of time. A photograph of the magnetron power supply is shown in Fig. 4.5.
Figure 4.3: A block diagram of the magnetron power supply.

Figure 4.4: The measured waveform and ripple of the magnetron power supply for a) DC output for cathode, and b) AC output for filament.
4.2 Waveguide Components

Various devices have been commonly used to transfer the microwave power, viz., coaxial line, and waveguide, depending on power or frequency used. A coaxial line with N-type connector was used where microwave power was slightly less than 200 W. With a specially designed 50 Ω coaxial line, a microwave power of ~ 300 W was coupled to the ECR plasma source and an electron density of $5 \times 10^{10}$ cm$^{-3}$ was achieved [88]. Considering the requirement of microwave power of about 2 kW continuous wave for our application, we have chosen a waveguide based microwave system. Although, the designated rectangular waveguide for 2.45 GHz is WR-340, WR-284 was used since its cut-off frequency is 2.078 GHz and it is suitable to operate at average power level of up to 6 kW. The other reason behind using WR-284 was that, this waveguide has also been used in some other on-going projects at RRCAT.

The cross-section of rectangular waveguide WR-284 was 72.14 mm (broader side) × 34.04 mm (narrow side) with typical wall thickness 2 mm, made up of oxygen free high conductivity copper. Standard plane, and choke type flanges were used for joining of the waveguide components [73]. The choke joint also keeps the electromagnetic field confined in...
the waveguide and helps to reduce the transverse microwave leakage. There have been various commercial software’s like CST-MWS [89], Ansoft-HFSS [90], Mafia [89] for the electromagnetic analysis / modeling of complex RF and microwave components. We chose the CST-MWS software for the electromagnetic analysis of waveguide components to reduce the microwave losses and with an adequate matching of the waves to the plasma chamber operating at frequency 2.45 GHz. It is 3D-electromagnetic simulation software. It has three different simulations techniques based on the finite integration method. Its transient solver technique was found to be suitable for analysis of microwave components. In the simulation, the outer boundaries were chosen as perfect electric conductor. The electromagnetic field pattern a) standard transverse electric TE mode, b) dominant TE\textsubscript{10} mode, and c) microwave studio simulated inside the WR-284 rectangular waveguide are shown in Fig. 4.6. The design and constructional details of the microwave components are presented in the following sections.

**Figure 4.6**: The electromagnetic field pattern a) standard transverse electric TE mode, b) dominant TE\textsubscript{10} mode, and c) microwave studio simulated, inside the WR-284 rectangular waveguide.

1) **Isolator with water load**: An isolator allows un-attenuated transmission of microwave power in the forward direction whereas power incident from other direction (reflected power) directed towards the load, is completely absorbed. In this way, it is an
important two-port device, which protect the microwave source from any reflected power damage due to mismatch of the waveguide impedance to the plasma impedance. It is a standard device (Model : 2722-163-02004, Make : M/s Richardson Electronics Pvt. Ltd., USA). The operating power of the isolator is 3 kW continuous wave, and it is water-cooled. It provides the isolation of 20 dB to the magnetron.

2) Directional coupler: The directional coupler [73,91,92] is a device which is commonly used for the measurement of the forward, and reflected power simultaneously. It is a four-port network (input port, transmitted port, coupled port and isolated port), and a known fraction of the microwave power flowing in a particular direction is sampled and used for evaluating the forward and reflected power and knowing its coupling factor. A block diagram of the directional coupler is shown in Fig. 4.7.

![Figure 4.7: A block diagram of the directional coupler.](image)

The working principle of a directional coupler is simple to understand. When a wave travels from port 1 to port 2, a fixed fraction of its power appears at port 3 (i.e. waves are in phase), and there is no power appears at port 4 (i.e. wave is 180° out of phase). Conversely, if the wave is travelling from port 2 to port 1, a fraction of this power appears at port 4, and there is no power appears at port 3. When the transmission takes place from port 1 (input port) to port 2 (transmitted port), the port 3 is called coupled port, and the port 4 is called isolated port. We have designed, and developed a directional coupler for 45 dB as its coupling factor using the microwave studio software. A schematic diagram of this directional coupler is shown in Fig. 4.8. It has waveguide as a main line, two coupling holes for power
coupling, a loop plate, a loop holder, and a power sampling connector (N-Type). The plane of the loop is kept parallel to the axis of the main line to which the loop is coupled. The length of the waveguide is 200 mm. Based on the schematic diagram of the directional coupler, the model has been created in *microwave studio* for simulations, for its scattering parameters. The directional couplers were designed to sample the power propagating in one direction. A schematic diagram of the orientation of loop plate with respect to direction of the propagation axis for the directional coupler is shown in Fig. 4.9. A model prepared in *microwave studio* for the directional coupler is shown in Fig. 4.10.

![Figure 4.8: A schematic diagram of the directional coupler.](image)

![Figure 4.9: A schematic diagram of the orientation of loop plate with respect to the direction of the propagation axis for the directional coupler.](image)
For *microwave studio* simulations, the materials for all the parts were chosen to be vacuum, except the inner conductor of the ports, loops, and supporting holder. These were simulated as perfect electric conductors. The distance between the holes for microwave coupling was kept one-fourth of the waveguide wavelength from the centre of the waveguide, to keep the frequency sensitivity to a minimum. A number of iterations were carried out by changing the loop, and the mid plate dimensions like, the thickness ($T : 0.5$ to $2$ mm) of the loop plate, the width of the loop plate ($W : 3$ to $5$ mm), the length of the mid plate ($X : 5$ mm), the width of the mid plate ($Y : 6$ to $10$ mm), the height of the loop ($H : 0.5$ to $2$ mm) from the main waveguide, and the theta rotations (clockwise) for the loop plate ($\theta_1$) as well as the mid plate ($\theta_2$). The coupling hole diameter ($\phi$) was also varied. The S-parameters optimized with frequency for the directional coupler are shown in Fig. 4.11. The S-parameters were optimized at 2.45 GHz frequency. From the figure, one can see that it has a very low insertion loss ($S_{21}$) of 0 dB, and the return loss ($S_{11}$) is about 100 dB, the coupling factor ($S_{31}$) is 45 dB, isolation ($S_{41}$) is 70 dB and the directivity ($S_{31}-S_{41}$) is 25 dB. The coupling factor and the directivity together decide the performance of the directional coupler. Variation of the coupling factor ($S_{31}$), and the directivity ($S_{31}-S_{41}$) with the hole diameter for the directional coupler at 2.45 GHz frequency is shown in Fig. 4.12. It was observed in

*Figure 4.10*: A model prepared in microwave studio for the directional coupler.
simulation that the coupling factor, and the isolation depends on the coupling hole diameter ($\phi$). The directionality of the coupling of the power remains unchanged, since the coupling and the isolation increases in symmetrical manner as expected, with increasing the hole diameter.

Figure 4.11: The S-parameters optimized with frequency for the directional coupler.

Figure 4.12: Variation of the coupling factor (S31), and the directivity (S31- S41) with the hole diameter for the directional coupler at 2.45 GHz frequency.
It was observed that when the height of the loop (H) of the main waveguide, and the
thickness of the loop plate (T) were increased, the coupling (45 dB) was not much affected.
The directivity was 30 dB for T = 0.5, W= 1 mm, X = 5 mm, Y = 6 mm, θ₁ = 54°, θ₂ = 45°.
The loop was placed close to the waveguide (H = 1 mm) and the effect of the orientation of
the loop plate, and the mid plate was studied, keeping T = 1 mm, W = 3 mm, X = 5 mm, Y =
6 mm, and φ = 13.7 mm. The results were as follows:

**Case 1**: When θ₁ and θ₂ are equal (θ₁ = θ₂), there is no directivity except at 90° and
270°. In the case of 90°, the coupling and directivity were 43 and 20 dB respectively. In the
270° case, the results remained the same except that the coupling port and the isolated port
interchange.

**Case 2**: When θ₁ = 90° and θ₂ was changed from 0 to 90°, the coupling factor and
the directivity were 43 and 20 dB respectively.

**Case 3**: When θ₁ was changed from 0 to 90° and θ₂ = 90°, the coupling remained
almost same (-) 43 dB, but the directivity changed from 25 to 33 dB.

**Case 4**: When θ₁ and θ₂ are not equal (θ₁ ≠ θ₂), some of the results for return loss,
coupling, isolation and directivity calculated when θ₁ ≠ θ₂ are given in Table 4.2.

<table>
<thead>
<tr>
<th>θ₁ (°)</th>
<th>θ₂ (°)</th>
<th>Return Loss (dB)</th>
<th>Coupling (dB)</th>
<th>Isolation (dB)</th>
<th>Directivity (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>54</td>
<td>45</td>
<td>100.00</td>
<td>45.00</td>
<td>70.00</td>
<td>25.00</td>
</tr>
<tr>
<td>63</td>
<td>54</td>
<td>100.80</td>
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<td>77.81</td>
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<tr>
<td>72</td>
<td>18</td>
<td>91.78</td>
<td>43.46</td>
<td>82.65</td>
<td>39.19</td>
</tr>
<tr>
<td>63</td>
<td>18</td>
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<td>81</td>
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<td>43.65</td>
<td>66.49</td>
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<td>45</td>
<td>103.20</td>
<td>43.59</td>
<td>69.86</td>
<td>26.27</td>
</tr>
<tr>
<td>81</td>
<td>36</td>
<td>99.07</td>
<td>43.59</td>
<td>69.26</td>
<td>25.67</td>
</tr>
</tbody>
</table>

*Table 4.2*: Return loss, Coupling, Isolation, and Directivity calculated when θ₁ ≠ θ₂.
Due to the 4-fold symmetry, the loop plate and the mid plate could be rotated from 0° to 90°. The coupling factor was not much affected with any orientation other than longitudinal, while the directivity was good when the loop plate and the mid plate were oriented at different angles. It was insensitive to the coupling hole diameter. The coupling becomes minimum when the loop is in the transverse direction. For the fabrication of the directional coupler (waveguide length 200 mm), brass rings of φ35 mm and height 30 mm were fixed over both the coupling holes. Care was taken at the time of brazing that coupling holes remain in the centre of the brass rings. The power sensing loops were fabricated using oxygen free high conductivity copper plate. The loops for sensing the radio frequency signal were mounted on a brass holder. A 50 Ω, 10 W chip resistance (for termination) and N-type connector (for sampling the power) were fixed on ends of the loop. The brass holder is an independent device with respect to the main waveguide, and could be freely rotated up to 360° for changing the orientation of the loop for coupling and directivity adjustment. The mechanical dimensions of the coupling loop play an important role in coupling and directivity performance because the electric and magnetic fields get actually coupled through it. All the sharp edges were removed to avoid electrical arcing due to the microwave power. A photograph of the directional coupler is shown in Fig. 4.13.

![Figure 4.13](image.png)

*Figure 4.13 : A photograph of the directional coupler.*
3) **Three-stub tuner**: Three-stub tuner [73,74] is a device which is widely used for matching of the impedance. It has three capacitive stubs, which are in parallel, and provides the variable shunt susceptance for matching the impedance. Combination of two neighbouring stubs (1 + 2 or 2 + 3) is used at a time for matching. Each pair covers about half of the complete matchable area in the Smith chart. The matchable area in the complex reflection coefficient plane is shown in Fig. 4.14. The regions A and B are prone to electrical breakdown.

![Figure 4.14: The matchable area in the complex reflection coefficient plane.](image)

The impedance of the plasma is inhomogeneous in nature because of the magnetic field gradients and recombination near the walls of the plasma chamber. It is dynamical in nature, and depends on the gas pressure, magnetic field, microwave power, operating conditions etc. Hence, it is required to match the impedance of the microwave transmission line to the impedance of the plasma, in order to minimize the microwave reflections. This is necessary to prevent the magnetron as well as the microwave vacuum window from damage. We have designed and developed a three-stub tuner as a variable impedance matching device. It matches the modulus and the phase of the incoming wave, in order to match the plasma impedance by adjusting the depth of the stubs (small sections of transmission lines which are
connected as shunt or series to the main line for impedance matching). A schematic diagram of the shunt connected the stub as a three-stub tuner is shown in Fig. 4.15.

**Figure 4.15**: A schematic diagram of the shunt connected the stub as a three-stub tuner.

For understanding the behaviour of the three-stub tuner, it was simulated in microwave studio. Vacuum was chosen as the medium for the waveguide, and perfect electric conductor for the shunt stubs. Three-stubs are mounted on the broad wall of the waveguide at equal intervals (three-eighth of the guide wavelength) to keep the frequency sensitivity to minimum. They are manually guided in or out. The length of the waveguide for three-stub tuner was 300 mm. The effect of shunt stubs was checked with microwave studio for its tuning range by varying the depth of penetration of the shunt stubs. It was observed that shunt capacitive susceptance changes for a wide range which helps to minimize the reflected power, and matches the impedance of the plasma as a load. A model prepared in microwave studio for the three-stub tuner is shown in Fig. 4.16.

**Figure 4.16**: A model prepared in microwave studio for the three-stub tuner.
4) **High voltage break**: A high voltage break [55] is a crucial component which is commonly used for the high voltage isolation between the microwave system and the plasma chamber. The plasma chamber remains floating at a high voltage for extracting the ion beam of required energy. It is essential to keep the microwave side at ground potential during the operation of the source for tuning the reflected power. We have designed and developed a high voltage break to withstand a voltage of 75 kV DC. A high voltage break should have good mechanical rigidity, low microwave loss, and low microwave radiation leakage to the environment. It was simulated using *microwave studio* from microwave point of view (low microwave loss), with the proper thickness to model the insulator to withstand 75 kV DC voltage. A model prepared in *microwave studio* for the high voltage break is shown in Fig. 4.17.

*Figure 4.17*: A model prepared in microwave studio for the high voltage break.

It has two waveguides of length 150 mm on both sides. Vacuum was chosen as inside material, and perfect electric conductor for the thickness of the waveguide. An insulator was sandwiched in between for finding its scattering parameters. The complete assembly was enclosed in a rectangular box with material as air to see the real effects under actual
environment. Teflon, high-density polyethylene, and polypropylene, which are commercially available, are normally used to get low loss for microwave, good mechanical strength and high voltage isolation. We have used Teflon as the high voltage isolation material. Using microwave studio, a numbers of iteration were carried out to get the proper thickness of Teflon to achieve the minimum return loss and insertion loss. The S-parameters optimized with frequency for the high voltage break: Teflon of thickness 7.35 mm are shown in Fig. 4.18. It is noted from the figure, that the return loss is 40 dB and the insertion loss is 0.08 dB for 7.35 mm Teflon thickness at 2.45 GHz frequency. The observed S-parameters show that the device is perfectly matched with input impedance with low microwave loss. The device was backed at 120 °C for two hours, for the settlement of high voltage epoxy under vacuum environment. The high voltage insulation was tested using high voltage Megger (Model: 220123-47, Make: M/s Megger, Biddle, England). It works satisfactorily up to 75 kV DC isolation without any leakage current.

Figure 4.18: The S-parameters optimized with frequency for the high voltage break: Teflon of thickness 7.35 mm.
5) Microwave vacuum window: Microwave vacuum window [10,11,12,93,94,95] is a device which is widely used for the isolation of microwave system from the plasma chamber which is under vacuum. A microwave vacuum window commonly uses a single / double / triple layer of mica, quartz, Teflon, alumina, aluminium nitride, boron nitride or combinations of these. The properties of these materials are such that they allow microwave propagation with low loss, and have good mechanical strength (except quartz). The thickness of the window material should be chosen in such a way that it can sustain the force of backstreaming of high energetic electrons, and stop the plasma flow back to the microwave source (i.e. magnetron). Sometimes, it may get punctured due to the excessive heating, leading to the failure of the ion source. The use of multilayered window increases the cost as well as complexity of fabrication. We have designed and built a single layer window at 2.45 GHz frequency. For constructing the geometry in microwave studio for simulation, the length of the waveguide was chosen to be 100 mm. Very thin capacitive obstacles were used for impedance matching on the both sides of the flange. Quartz was chosen as the window material, since it is cheap and easily available. A model prepared in microwave studio for the microwave vacuum window is shown in Fig. 4.19.

For simulation purpose, vacuum was chosen to represent all the components. In the simulations, the thickness of the window material was varied from 0.1 to 10 mm. To check the validity of the design and model, it was first run with zero thickness to make sure that same results are obtained as achieved with only standard flanges (return loss S11, insertion loss S21). Variation of the return loss (S11), the insertion loss (S21) with the thickness of quartz plate for the microwave vacuum window at 2.45 GHz frequency is shown in Fig. 4.20. The S-parameters optimized with frequency for a microwave vacuum window with quartz thickness of 6 mm are shown in Fig. 4.21. The simulations show that for the 6 mm thickness of quartz plate, the return loss is 75 dB and the insertion loss is 2.5 µdB, which is close to
zero. A rectangular quartz plate was fixed using conductive glue. The window was helium leak tested using a helium leak detector and the leak rate was found to be less than $10^{-10}$ mbar l/s. The outer periphery was covered with grounded copper strip to avoid any transverse leakage of the microwave power.

**Figure 4.19**: A model prepared in microwave studio for the microwave vacuum window.

**Figure 4.20**: Variation of the return loss ($S_{11}$), the insertion loss ($S_{21}$) with the thickness of quartz plate for the microwave vacuum window at 2.45 GHz frequency.
Figure 4.21: The S-parameters optimized with frequency for a microwave vacuum window with quartz thickness of 6 mm.

6) Microwave launcher: Microwave launcher is a device which is used to couple the microwave power to the plasma chamber. There are various devices used as a microwave launcher viz. coaxial line [96], open ended waveguide [97], horn, slotted and helical antenna [12,80,98], ridged and tapered waveguide [64,99,100,101,102]. The performance of a plasma source (viz. electron density, electron temperature produced) and hence the ion current can be improved significantly by proper selection and design of the microwave launcher. Here, we have studied using microwave studio a tapered waveguide as well as a ridged waveguide. The latter is widely used now days as a microwave launcher because of its ease of fabrication, low cost, and wide bandwidth. Both launchers have a special property that they gradually matches the impedance of the microwave transmission to the impedance of the plasma, which leads to a low reflection and a higher order mode coupling. Other devices were ignored because of space limitations. The details of the two launchers are described in the following sections.
a) **Tapered launcher**: We have designed and developed a tapered launcher to feed the microwave power to the plasma chamber. It gradually matches the plasma impedance and increases the intensity of the electric field at the mouth of the waveguide. The intensity of the electric field is a function of narrow dimensions of the waveguide. This was simulated using the *microwave studio*. A model prepared in *microwave studio* for a tapered launcher is shown in Fig. 4.22.

*Figure 4.22*: A model prepared in microwave studio for a tapered launcher.

For simulations of the tapered launcher in *microwave studio*, two bricks of length 50 mm of dimensions 72.14 mm × 34.04 mm (matching with waveguide dimension) and 72.14 mm × 10 mm (achieved with *microwave studio* simulation), with a wall thickness 2 mm, at a distance of one guide waveguide length. Thus by changing the narrow dimensions of the waveguide, the intensity of the electric field was simulated using the *microwave studio* software. It has been observed that the intensity of the electric field becomes almost doubled when the narrow dimension reaches 10 mm. Variation of electric field with the height of waveguide for a tapered launcher is shown in Fig. 4.23.
This electric field enhances the electron density, since the power is directly proportional to the square of the intensity of the electric field. In *microwave studio* simulation, vacuum was chosen as the inner material and a perfect electric conductor for the wall thickness. The S-parameters optimized with frequency for a tapered launcher are shown in Fig. 4.24. It is noted from the figure, that the return loss is 23 dB and the insertion loss is close to 0.02 dB at 2.45 GHz frequency. The performance of microwave power coupling into the plasma was found to be very good, and most of the incoming power was coupled to the plasma chamber with very low reflected power. The electric field distribution in a tapered launcher is shown in Fig. 4.25.
**Figure 4.24**: The S-parameters optimized with frequency for a tapered launcher.

**Figure 4.25**: The electric field distribution in a tapered launcher.

**b) Ridged Waveguide**: A ridged waveguide with three / four sections has been widely used as a microwave launcher in different laboratories like Chalk River Laboratory, Trips, Silhi, MIDAS, VIS, ALISES, and Spiral2. Like a tapered waveguide, a ridged
waveguide also matches the impedance (progressively reducing with ridged gap) from the source waveguide impedance to the plasma impedance which is of the order of 150-100 Ω [103]. A ridged waveguide concentrates the electric field near the source axis, while in a standard waveguide it is distributed in the sinusoidal form along the larger dimensions. It has a wide bandwidth [104,105]. This reduces the necessity of high power level of the microwave for the extraction of higher ion beam current and also reduces the requirement of water-cooling required for the plasma chamber, thereby enabling operation of the source in continuous mode for several hours. The magnitude of the electric field intensity depends on the number of ridged sections.

Analytical approach: There are various devices like coaxial line [96], open ended waveguide [97], E-plane horn antenna, E-plane horn antenna with helical and slotted antenna [12,79,98] etc. have been used to feed the microwave power to the plasma chamber for plasma heating by means of ECR. The use of above devices is limited due to space restrictions, as they make the ECR source bulky in size. With the demand of compact ECR source, advancement of research in science and technology, it has been observed that with the use of ridged waveguide, the performance of the plasma production (electron density and temperature) has improved leading to higher ion beam current extraction. Another advantage of using ridged waveguide is that, it can be operated at a lower frequency and has lower impedance and a wider mode separation between the cut-off numbers of its dominant mode and the first higher mode than a standard waveguide. Due to these advantages, ridge waveguides have been extensively used in microwave active and passive components, and ECR based high intensity, low emittance proton source development at 2.45 GHz frequency. A schematic diagram of a four sections ridged waveguide coupled to the plasma chamber is shown in Fig. 4.26.
Figure 4.26 : A schematic diagram of a four sections ridged waveguide coupled to the plasma chamber.

Here, $Z_{vi}$, $Z_1$, $Z_2$, $Z_3$, $Z_4$, and $Z_p$ are the impedances of the four sections ridged waveguide, where $Z_{vi}$ is the voltage-current impedance of a standard waveguide, and $Z_p$ is the plasma impedance. The voltage-current impedance of a standard rectangular waveguide is given by,

$$Z_{vi} = \left[ \frac{\pi \eta}{2} \left( \frac{2b}{a} \right) \right]^{1/2} \left[ 1 - \left( \frac{\lambda_c}{\lambda_0} \right)^2 \right]$$ .................................................................(4.1)

where $\eta$ is the free space impedance (377 $\Omega$), $2b$ is the height of the waveguide, $a$ is the width of the waveguide, $\lambda_0$ is the free space wavelength, and $\lambda_c$ is the cut-off wavelength, is about $Z_{vi} = 527 \Omega$ for standard waveguide WR-284 (width : 72.14 mm, height : 34.04 mm) operating in a dominant mode TE$_{10}$ at operating frequency of 2.45 GHz. The dimensions of the ridged waveguide have been evaluated using standard characteristics impedance chart for the binomial matching transformers [75], to match the voltage-current impedance of a standard waveguide to plasma impedance, and each sections of the waveguides was one quarter wavelength ($\lambda_g/4$) long at operating frequency, where $\lambda_g$ is the guide wavelength. The impedances of the ridged section are evaluated by the relation,

$$\frac{Z_{n+1}}{Z_n} = \exp \left[ 2^{-N} C_n^N \ln \frac{Z_p}{Z_{vi}} \right]$$ .................................................................(4.2)

where $n$ is the number of ridged section, $N$ is the total number of the ridged section, $C_n^N$ is the binomial coefficients i.e. $N! / (N-n)! \ n!$. Variation of impedance with the length for a standard and a ridged waveguide is shown in Fig. 4.27.
Figure 4.27: Variation of impedance with the length for a standard waveguide and a ridged waveguide. (The solid blue curve is for visual aid only and dot indicates the calculated impedance in that section, ref. Table 4.3)

The figure shows that the impedance of a ridged waveguide decreases exponentially with increasing the number of ridged sections and progressively approaches to plasma impedance at the fourth section of the ridged waveguide. On the other hand, the voltage-current impedance of a standard waveguide remains constant. The dimensions and impedances of the four sections of the ridged waveguide are shown in Table 4.3.

<table>
<thead>
<tr>
<th>Number of Ridge Sections (N)</th>
<th>Impedance of the Ridge Section ($Z_n$) (Ω)</th>
<th>Gap of the Ridge Section ($2b_2$) (mm)</th>
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<tr>
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<td>32</td>
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<tr>
<td>2</td>
<td>341</td>
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<td>4</td>
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</table>

Table 4.3: The dimensions and impedances of the four sections of the ridged waveguide.
Electromagnetic analysis of a ridged waveguide: The dimensions obtained with the analytical approach, as shown in Table 4.3, are used for generating a model structure of a ridged waveguide. An electromagnetic model of a ridged waveguide coupled with a plasma chamber is shown in Fig. 4.28.

![Diagram of an electromagnetic model of a ridged waveguide coupled with a plasma chamber.](image)

**Figure 4.28**: An electromagnetic model of a ridged waveguide coupled with a plasma chamber.

It has a standard waveguide part, an upper and a lower ridge section and a plasma chamber. For modelling purpose, the material for standard waveguide and ridged sections were chosen as vacuum and perfect electrical conductor respectively. The transient solver module of the *microwave studio* was used for modelling the waveguide. The transient signal was fed from a standard waveguide, as input port 1, and ridged waveguide as output port 2. The structure was discretized at 10 lines per wavelength mesh density. The complete geometry had 28510 mesh cells. The dimension of the ridge gap I to IV was varied in the steps of 1 mm and corresponding scattering parameters were recorded at 2.45 GHz frequency. The simulation results in terms of scattering parameters at 2.45 GHz frequency for ridge gap I, II, III, and IV are presented in Figs. 4.29, 4.30, 4.31, and 4.32.
Figure 4.29: The simulated variations of the return loss (S11), the insertion loss (S21) for a ridge gap I at 2.45 GHz frequency for the ridged waveguide.

Figure 4.30: The simulated variations of the return loss (S11), the insertion loss (S21) for a ridge gap II at 2.45 GHz frequency for the ridged waveguide.
Figure 4.31: The simulated variations of the return loss (S11), the insertion loss (S21) for a ridge gap III at 2.45 GHz frequency for the ridged waveguide.

Figure 4.32: The simulated variations of the return loss (S11), the insertion loss (S21) for a ridge gap IV at 2.45 GHz frequency for the ridged waveguide.

In the above figures, the dashed lines show the actual gap of each ridge. It is observed from these figures that the scattering parameters S11 (return loss) is sensitive to the gap of the ridge, whereas the other parameter S21 (insertion loss) is not so sensitive. Figure 4.29, shows
that the return loss decreases with increasing gap of ridge gap I. The behaviour seen in Figure 4.30, 4.31, 4.32 is the return loss is 15 dB at an analytical ridged gap values corresponding to the 2.45 GHz frequency and the insertion loss is close to zero. The figure 4.33 show that, optimized ridge gaps (31 mm, 20 mm, 15.5 mm, and 10 mm) are closed to the analytical ridged gap values. The optimized variations of the return loss (S11), the insertion loss (S21) for the ridged waveguide with frequency range between 2.4 to 2.5 GHz is shown in Fig. 4.33.

![Graph](image)

*Figure 4.33 : The optimized variations of the return loss (S11), the insertion loss (S21) for the ridged waveguide with frequency range between 2.2 to 2.7 GHz.*

The return loss of the integrated geometry of the ridged waveguide is 38 dB (i.e. power transmission ~ 100 %) and corresponding insertion loss is 0 dB at 2.45 GHz frequency. The variations of the return loss (S11), the insertion loss (S21) with ridge width at 2.45 GHz frequency for the ridged waveguide is shown in Fig. 4.34. It shows that the return loss is first decreasing with increasing width of ridge sections from 10 to 11 mm, due to establishing the uniform electric field distribution between the ridge gap. Beyond this, it
increases and reaches a saturation level, may be due to distortions of the electric field intensity between the ridged gap. The insertion loss does not change much; it remains close to zero with increasing the ridged width.

![Graph](image)

**Figure 4.34**: The variations of the return loss (S11), the insertion loss (S21) with ridge width at 2.45 GHz frequency for the ridged waveguide. Dashed lines show the actual width of ridge.

A vector plot of the distribution of the electric field for a standard waveguide is shown in Fig. 4.35. The maximum electric field amplitude along Z-axis is about 757 V/m at port 1. A vector plot of the distribution of the electric field for a ridged waveguide is shown in Fig. 4.36. The maximum electric field amplitude along Z-axis is about 1574 V/m, which is about two times that of the standard waveguide as evaluated with analytical approach and the distribution of electric field intensity is also uniform in the ridge gap.

The electric field intensity distribution in the horizontal mid-plane for a standard waveguide and a ridged waveguide of width 12.5 mm is shown in Fig. 4.37. The electric field
Figure 4.35: A vector plot of the distribution of the electric field for a standard waveguide.

Figure 4.36: A vector plot of the distribution of the electric field for a ridged waveguide.

Figure 4.37: The electric field intensity distribution in the horizontal mid-plane for a standard waveguide and a ridged waveguide of width 12.5 mm.
intensity for a standard waveguide varies in the form of an arc of sinusoidal pattern along the width of waveguide, while that for a ridged waveguide has a Gaussian distribution, with maximum electric field intensity localized within the ridged gap.

**Microwave coupling study**: For obtaining the best microwave coupling [106,107,108] between the optimized ridged waveguide and the plasma chamber, the optimized ridged waveguide was coupled to the plasma chamber, to check the good match between the ridged waveguide and the plasma and to produce the electric field intensity maximum at the centre of the plasma chamber. The electric field intensity in a horizontal plane of the ridged waveguide coupled with the plasma chamber is shown in Fig. 4.38. The electric field intensity in a transverse plane of the ridged waveguide coupled with the plasma chamber is shown in Figure 4.39. It shows that the absolute electric field intensity in the horizontal plane and in the transverse plane is about 2750 V/m at 2.45 GHz, at the centre of the plasma chamber. It also shows that the electric field intensity is maximum at the centre of

![Image](image.png)

**Figure 4.38**: The electric field intensity in the horizontal plane of the ridged waveguide coupled with the plasma chamber.
the plasma chamber, suitable to produce high density plasma and further supports the extraction of higher ion beam current. Thus, it is observed that the microwave energy is well coupled to the plasma chamber using a ridged waveguide.

7) **Glass-water load**: There are various materials like graphite, cement with graphite, wood, quartz, water etc. which are used as a microwave load depending on the level of microwave power to be dissipated [109,110]. We have used water as an absorbing material, since it is a good absorber of the microwave power. One can absorb a good amount of microwave power easily with continuous flow of water. The heat dissipated is removed by continuous flow of water. We have designed and fabricated a pyrex glass-water load for absorbing the 3 kW microwave power at 2.45 GHz frequency, for the characterization of the microwave components at high power. This glass-water load was connected at the end of the microwave-integrated line for initial high power characterization of the microwave components. The glass-water load was fabricated using Pyrex glass having tube diameter of 12 mm and length equal to one guide wavelength. The Pyrex glass-water load has kovar...
fittings at both ends of the tube. The glass-water load was calibrated using standard network analyzer at the operating frequency. A photograph of the glass-water load is shown in Fig. 4.40.

![Glass Water Load](image)

**Figure 4.40**: A photograph of the glass-water load.

### 4.3 Characterization of the Microwave Components

The low power characterization [111,112] of the microwave components was carried out using a network analyzer (Model : E5071 ENA series with electronic-cal kit, N4431-60004, Make : M/s Agilent Technologies Pvt. Ltd., USA) for validating the design procedure. Scattering parameters of the microwave components were measured using network analyzer. The test setup for low power characterization of the microwave components using network analyzer is shown in Fig. 4.41.

The full term characterization (two port scattering parameters viz. S11, S21, S12 and S22) was carried out with frequency 2.4 to 2.5 GHz (marker at centre frequency 2.45 GHz, bandwidth 100 MHz), with various standards of network analyzers. The measurement inaccuracy in characterization was 10 mDB. The measurement accuracy in the results could be maintained by routine characterization of the microwave components. The measured results for the directional coupler are shown in Fig. 4.42.
Figure 4.41: The test setup for low power characterization of microwave components using network analyzer.

![Test Setup Diagram]

Figure 4.42: The measured results for the directional coupler. The solid curve is for visual aid only.

The mechanical error of fabrication etc. could not be taken into account in the microwave studio simulation. Hence, some differences in simulations and actual results were expected. The results are summarized as: the insertion loss for the device is close to 0.05 dB,
the return loss is 37 dB (measured) and 100 dB {Fig. 4.11, software, assumes ideal conditions}, the coupling factor 44.5 ± 0.5 dB, the directivity is 25.5 ± 0.5 dB at 2.45 GHz frequency, except the return loss, the measured results are close to optimized results. This validates the design procedure of the directional coupler. For the low power measurements of the three-stub tuner, two cases were studied i.e. all stubs were FULL IN and FULL OUT to see the impedance variations on Smith chart. The measured results for the three-stub tuner when stubs are FULL IN are shown in Fig. 4.43. One can see from the figure that, when all stubs are FULL IN, the insertion loss is 32 dB and the return loss is 0.3 dB. In this case, the microwave power was severely reflected which causes more insertion loss. The measured results for the three-stub tuner when stubs are FULL OUT are shown in Fig. 4.44. It is noted from this figure that the measured insertion loss is 0.3 dB and remains unchanged up to 2.46 GHz, whereas the return loss is close to 26 dB up to 2.46 GHz, and beyond this it increases with increasing frequency. The behaviour of the figure is as expected, since the obstacles of the stubs produce standing wave patterns during the propagation of the microwaves which cause the losses of microwave power.

**Figure 4.43**: The measured results for three-stub tuner when stubs are FULL IN. The solid curve is for visual aid only.
Figure 4.44: The measured results for the three-stub tuner when stubs are FULL OUT. The solid curve is for visual aid only.

Effectively it covers the wide range of insertion loss 3 to 30 dB (due to obstacles of stubs, simply comprise for impedance matching) and the return loss is 0 to 25 dB. The insertion loss was brought down to less than 1 dB with the multiple combinations of stubs during the operation of the source (matched condition i.e. low reflected power). The measured results for the high voltage break are shown in Fig. 4.45. It is noted from the figure, that the return loss is 25 dB {Fig. 4.18, software, 40 dB} and the insertion loss is 0.1 dB {Fig. 4.18, software, 0.08 dB} at 2.45 GHz frequency.

Figure 4.45: The measured results for the high voltage break. The solid curve is for visual aid only.
The measured results for the microwave vacuum window are shown in Fig. 4.46. It is noted from the figure, that the return loss is 21 dB (Fig. 4.21, software, 75 dB) and the insertion loss is 0.3 dB (Fig. 4.21, software, 0.0 dB) at 2.45 GHz frequency. The discrepancies in the measured and software results may be due to fabrication, misalignment error, test port match, waveguide to coaxial adaptor etc. and software assumes always ideal matched conditions.

**Figure 4.46**: The measured results for the microwave vacuum window. The solid curve is for visual aid only.

Finally, all the microwave components were integrated with the glass-water load at the end, for the high power characterization of the microwave components. This microwave line was energized and tested for 2 kW of microwave power at 2.45 GHz frequency. The performance of the microwave line was quite satisfactory. A radio frequency and microwave survey using radiation survey meter (Model: HI 1501, Make: M/s Richardson Electronics Pvt. Ltd., USA) was carried out. A maximum radiation leakage of 2-3 mW/cm² was observed around the microwave vacuum window and the high voltage break. The radiation level was found to be well within the permissible limit. Further, to avoid any radiation leakage from the dielectric portion, a metallic cover was wrapped externally at the joint. An integrated
schematic diagram of the microwave components with glass-water load for high power characterization is shown in Fig. 4.47.

**Figure 4.47**: An integrated schematic diagram of the microwave components with glass-water load for high power characterization.

Variation of microwave output power with the cathode current of the magnetron is shown in Fig. 4.48. The figure shows that microwave output power increasing linearly with increase in the cathode current of the magnetron, with a slope of 2.73 W/mA. The reflected power was minimized during the course of experiments with the tuning stubs of the three-stub tuner.

**Figure 4.48**: Variation of the microwave output power with the cathode current of the magnetron. The solid curve is for visual aid only.