



Research Article

Automatic Impedance Double-Stub Tuner for High-Power S-Band Microwave Networks

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A B S T R A C T

This paper presents a systematic development of an impedance tuner for use in S-band microwave networks which specifically targets the 2.45GHz ISM band to provide efficient automatic impedance matching under high-power operating conditions. Designed as a double-stub structure within the WR-340 waveguide, the proposed tuner is intended to work alongside a reflectometer in a closed-loop feedback system, utilizing reflectometer measurements to dynamically adjust impedance. This method facilitates stable and temperature-independent performance, making the tuner well-suited for online monitoring and control in industrial waveguide application. To achieve this, a combination of simulations and preliminary measurements using Agilent E5071c VNA and coax-to-waveguide (C2W) transitions were employed to validate the proposed tuner's stability and efficiency in real-time network parameter adjustment. Key findings in this work revealed capability of the backshort reflecting strong incident power and delivering about -78dB transmission at 2.45GHz highlighting the tuner's robustness and operational consistency. This provides a practical solution for precise impedance matching in industrial environment for reliable high-power microwave control while maintaining optimal performance.

INTRODUCTION

Microwave tuning has experienced immense development over the years. Its contribution to ensure good impedance matching and fine tuning for different microwave applications makes it an indispensable tool. In recent times, they are employed in microwave heating systems for sterilization and pasteurization applications [1], sintering of ceramics [2], drying of agricultural and forestry products [3] and in wireless communication systems [4-7] among other applications.

In industrial processes, microwave tuners play a major role in defining the quality of the product. Hence, it is an integral element in determining the economic outcome of the industry. Microwave tuners are primarily waveguide elements used to match the load impedance with the source impedance. They provide the most efficient coupling of power to the load to reduce the extent of reflected power. There are varieties of tuners depending on the type of microwave application. In microwave heating mechanisms, a lot of tuners are based on stub tuning technique. Single, double, triple and multi stub designs configurations exist. Due to the cons associated with single and double stubs, triple-stub tuners have proven efficient since they solve challenges such

as high energy leakages and matching problems of different load impedances. Thus, it is mostly employed in high power applications. Impedance matching in microwave applications with the aid of the tuner can be achieved either manually or automatically. In typical systems, impedance matching is achieved by changing the depth of the tuner's post in a try-and-error manner to tune the network. This resulted in the development of electrical and auto tuners. They employ motorized tuning stubs that are controlled electronically by a microcontroller or microprocessor. These types of tuners are suitable for applications where there are fast and significant changes of the load impedance due to changes in the load. A typical case is the microwave plasma [8].

Several tuning design configurations exist for microwave guide tuners. Information and data collected from this field has prompted a few studies aimed at developing microwave tuning systems with a few prototypes developed. Most tuning systems are primarily based on stub tuning. In [9, 10], short-circuited guide sections are employed as shunt or series reactive elements. The nature of reactive load posed by the stub using this technique are determined by the depth and position of the stub in the waveguide. Single, double, triple, and multiple stub configurations essentially form the backbone of the design

process. The single-stub tuner (sliding screw) is composed of a screw or post with variable depth mounted in sliding carriage and employed to move freely about a distance along the longitude axis of the waveguide. On the other hand, the double-stub tuner uses two tuning stubs in fixed positions and depends on its variable depth for matching. Triple-stub tuner, the popular choice for high power applications, employs its closest stub to transform the load admittance to a new one. Subsequently, the remaining stubs match the new admittance in the manner of the double stub tuner. However, challenges such as high energy leakage, load matching impedance complications and arcing associated with single, double, and triple-stub tuners respectively make them less efficient.

A work conducted by [11] showed an automatic tuner unit operation for the microwave system of the ESS-Bilbao H⁺ ion source (ISHP). The system utilizes an automatic tuner unit (ATU) control system to ensure proper impedance matching and maximum power transfer during plasma formation. This is achieved when the reflection coefficient is measured at various points distributed along the waveguide and its results are used to calculate the optimum configuration of the ATU using a controller. The three capacitive posts of the ATU are moved with a motor to the desired positions to effect change in load impedance based on data received from the controller. Although this system eliminates the try-and-error fashion of varying the depth of the tuner's posts, rigorous programming is required. [12] present a tunable matching network (TMN) based on phase-switched impedance modulation (PSIM). The system is composed of a power stage and a gate-waveform generation module. The power stage has phase-switched capacitors, input and output filters that ensure load-to-source impedance match. The gate-waveform generation module was designed based on a phase locked loop (PLL) chip and 5-12 MHz voltage-controlled oscillator (VCO). This was to achieve sensing and drive of the PSIM circuit. This approach provided a much faster and accurate impedance matching as compared to the traditional techniques. Aside from its advantages, the development is very technical and protection against voltage surge is required.

The model showcased in [13, 14] are E-H tuner designs used for impedance matching. E-H tuner is a stub-based waveguide impedance tuner based on the Magic-T hybrid junction. In [13], it was designed using Ansys HFSS 3D electromagnetic simulation software to match a reflection factor of 0.001 using inductive irises in the E- and H- arms. Removable sliding shorting plungers with half wavelength range were connected to the E-H arms. Additionally, the choke plunger is motorized and positioned with the computer. The E-H tuner in [14] has a dielectric-based waveguide backshorts at the E- and H- arm. This offers more accurate tuning than a traditional backshort since the phase changes gradually with the movement of the micrometer spindle. Even though this system provides a higher degree of accuracy, it is difficult to implement them in applications requiring very small dimensions.

There are other tuning techniques that employ semiconductor devices to achieve a large tuning range for microwave applications [15, 16]. In relation to [15], the system employs a SiC Schottky diode varactor in dynamic load modulation (DLM) of power amplifiers (PA). The SiC provides a wide bandgap that

withstands high electric fields and has mobility for frequency devices. SiC varactor is fabricated using a custom designed doping profile by spreading C(V). Allowable voltage swings are achieved by making C(V) non-abrupt. This approach enables large tuning range and is mostly used in wireless communication.

In this paper, we present the development of a microwave tuner designed for impedance matching in industrial high-power S-band waveguide networks. This tuner, which is fitted with motorized actuators, forms part of an automatic tuning system. The device is cascaded with a reflectometer discussed in [17] and uses inline reflection measures to achieve self-executing tuning mechanism through a closed feedback loop. This paper is organized as follows: section 1 provides the introduction and review of literature; section 2 presents the methodology which covers the general theory about matching and considerations for developing the waveguide tuner. Section 3 describes the simulations, calibration, and evaluation of the developed system. Section 4 concludes the work.

METHOD

Impedance Matching and Tuning

The basic concept of impedance matching is illustrated in Figure 1, which shows a matching network placed between a load impedance and a transmission line. The ideally lossless matching network is usually designed such that characteristic impedance Z_0 is seen looking into its input. In this case, reflections on the transmission line to the left of the matching network will be eliminated, although there will usually be multiple reflections between the load and the matching network. Assuming the generator itself is matched, power loss in the feed line is minimized and maximum power is delivered when the load is matched to the line.

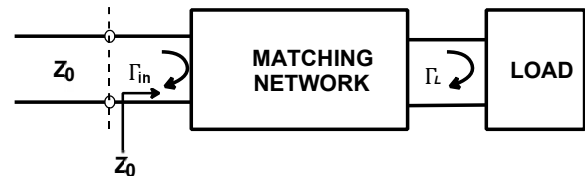


Figure 1. An arbitrary network matching load to transmission line of characteristic impedance Z_0 .

Generic Matching Network

Consider a generic matching network placed between a transmission line and an arbitrary load as shown in Figure 1. The load Z_L is characterized by a reflection coefficient Γ_L and connected to port 2 of the two-port network, whose role is to transform Z_L to the Z_0 at port 1. The matching network is by the 2×2 scattering matrix S as in [18, Eq. (1)]

$$[S] = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \quad (1)$$

If the matching network $[S]$ is reciprocal, $S_{12} = S_{21}$ and if symmetric in addition, $S_{11} = S_{22}$. Thus, in symmetric networks, the identical nature of the network observed from both the input and output implies that $S_{11} = S_{21}$. Furthermore, $[S]$ is unitary if the network is lossless.

$$[S][S]^T = [U] \Rightarrow \begin{bmatrix} |S_{11}|^2 + |S_{12}|^2 & S_{11}S_{21}^* + S_{12}S_{22}^* \\ S_{21}S_{11}^* + S_{22}S_{12}^* & |S_{21}|^2 + |S_{22}|^2 \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \quad (2)$$

The reciprocal nature of the network coupled with its lossless property reduces the network's characteristics to

$$|S_{11}|^2 + |S_{12}|^2 = 1 \quad (3)$$

$$S_{11}S_{21}^* + S_{12}S_{22}^* = 0 \quad (4)$$

The input reflection coefficient when port 2 is loaded with Z_L is

$$\Gamma_{in} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L} = S_{11} + \frac{S_{12}^2\Gamma_L}{1-S_{11}\Gamma_L} \quad (5)$$

Matching is fulfilled if the Z_L is transformed to a matched load at the input i.e., $\Gamma_{in} = 0$

$$\Gamma_{in} = S_{11} + \frac{S_{12}^2\Gamma_L}{1-S_{11}\Gamma_L} = 0 \Rightarrow \Gamma_L = \frac{S_{11}}{S_{11}^2 - S_{12}^2} \quad (6)$$

The matching condition becomes

$$\Gamma_L = \frac{S_{11}}{S_{11}^2 - S_{12}^2} = \frac{S_{11}S_{21}^*}{S_{11}S_{11}S_{21}^* - S_{21}S_{21}S_{11}^*} = \frac{|S_{21}|^2 S_{11}}{S_{21}|S_{11}|^2 + S_{21}|S_{21}|^2} = S_{11}^* \quad (7)$$

Waveguide Tuners

Typical tuning systems for waveguide networks are based on stub tuning, a technique where short-circuited guide sections are employed as shunt or series reactive elements. The depth and position of the stub in the waveguide determines the nature of reactive load posed by the stub. Such tuners come in the form of single, double, triple, or multiple stub configurations. The operation of the single stub tuner requires a new stub position for every load. An approximate equivalent of a single-stub matching network is the slide screw tuner.

This tuner consists of a screw or post with variable depth and mounted in sliding carriage. The screw is free to move about $\frac{\lambda_g}{2}$ along the longitudinal axis of the waveguide. The screw penetrates the guide through a centered narrow broad wall slot cut along the current flow lines. Such slots ensure that there is a negligible perturbing effect on the internal field. The sliding screw tuner is typically not preferred for high power applications as its broad wall slot introduce the risk of high energy leakage. The double-stub tuner uses two tuning stubs in fixed positions and rely on their variable depth for matching. The double stub tuner solves the slide screw problem of leakage as the stub positions can be adequately sealed. However, the double stub-tuner cannot match all load impedances. On the Smith chart, such loads are a range of admittances in the so-called forbidden region shown in Figure 2.

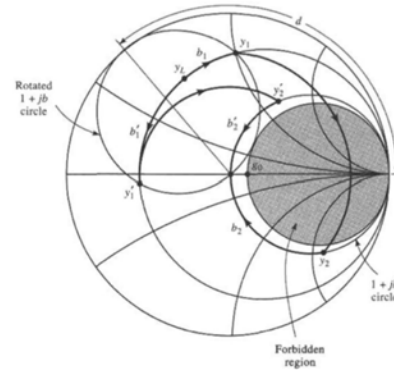


Figure 2. Smith chart diagram of the double-stub tuner showing the forbidden region [18].

The forbidden region is not eliminable but can be reduced. A simple way to shrink the region is to reduce the distance between the stubs as practically as possible [18]. This essentially rotates the unit admittance circle towards the infinite admittance point on the chart. In practice, the stub spacing are usually $\frac{\lambda_g}{8}$ or $\frac{3\lambda_g}{8}$. Theoretically, if the length of the transmission line between the stub is closer to the load and the load can be adjusted, the load admittance can always be moved out of the forbidden region. This cannot be achieved dynamically in waveguide systems.

The triple-stub tuner remedies the forbidden region problem of the double-stub tuner. The closest stub to the load transforms the load admittance into new admittance. Then, the two remaining stubs match this new admittance in typical double stub tuner fashion. As such the triple-stub can match all values of load admittances (except purely reactive loads) and is used in many high-power applications. However, the presence of the posts/ stub in the guide increases the risk of arcing. This is because the stubs reduce the "height" of the electric field causing the field strength to increase substantially. In such situations, the breakdown of air is a possibility.

E-H Tuner

The E-H tuner is stub-based waveguide impedance tuner based on the Magic-T hybrid junction. It combines series (E - plane) and shunt (H - plane) shorting stubs to achieve matching (Figure 3). The positions of the short circuits (backshorts) in the E - and H - plane arms are variable; thus, a wide range of load impedances can be matched. The Magic-T hybrid, which forms the main body of the E-H tuner is a waveguide directional coupler with an equal power division ratio (3 dB) between two of its outputs with zero coupling in the last port. The Magic-T junction is also anti-symmetric in its coupling, i.e., it produces a 180° phase shift between the output ports. The Magic-T is described by the S matrix in [19, Eq. (8)].

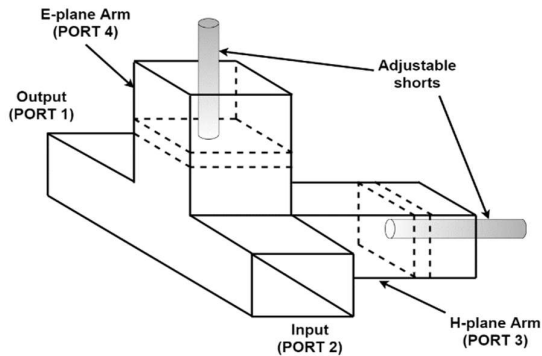


Figure 3. E-H Tuner adapted [19]

Tuning Law and Convergence (Theory Formulation)

The scattering matrix for an ideal Magic-T hybrid shown in Figure 3 is

$$[S_{\text{Magic-T}}] = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & 0 & 1 & 1 \\ 0 & 0 & 1 & -1 \\ 1 & 1 & 0 & 0 \\ 1 & -1 & 0 & 0 \end{bmatrix} \quad (8)$$

The in-arm stubs with length $l_{E,H} \in [0, \frac{\lambda_g}{2}]$ corresponds to loads characterized by the reflection coefficients on the E and H arms ports given by the [19, Eq. (9)].

$$\Gamma_{E,H} = e^{-j2\beta l_{E,H}} = e^{-j\phi_{E,H}} \quad (9)$$

where $\phi_{E,H} = 2\beta l_{E,H} \in [0, 2\pi]$. Therefore, the input and output power waves of the E - H tuner are as follows

$$[b] = [S_{\text{Magic-T}}][a] = \begin{bmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & 0 & 1 & 1 \\ 0 & 0 & 1 & -1 \\ 1 & 1 & 0 & 0 \\ 1 & -1 & 0 & 0 \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ \Gamma_H b_3 \\ \Gamma_E b_4 \end{bmatrix} \quad (10)$$

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \begin{bmatrix} \Gamma_H b_3 \\ \Gamma_E b_4 \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} \Gamma_H & \Gamma_E \\ \Gamma_H & -\Gamma_E \end{bmatrix} \begin{bmatrix} b_3 \\ b_4 \end{bmatrix} \quad (11)$$

$$\begin{bmatrix} b_3 \\ b_4 \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \quad (12)$$

Using Eq. (11) and (12), the E - H tuner can be reduced to a 2-port device

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} \Gamma_H & \Gamma_E \\ \Gamma_H & -\Gamma_E \end{bmatrix} \cdot \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} = \frac{1}{2} \begin{bmatrix} \Gamma_H + \Gamma_E & \Gamma_H - \Gamma_E \\ \Gamma_H - \Gamma_E & \Gamma_H + \Gamma_E \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \quad (13)$$

Therefore, the E - H tuner has a scattering matrix of form

$$[S_{E-H \text{ tuner}}] = \frac{1}{2} \begin{bmatrix} \Gamma_H + \Gamma_E & \Gamma_H - \Gamma_E \\ \Gamma_H - \Gamma_E & \Gamma_H + \Gamma_E \end{bmatrix} \quad (14)$$

It is apparent that the E - H tuner is reciprocal and lossless since its scattering matrix is both symmetric and unitary.

The E - H tuner largely satisfies the primary need for a matching device to be capable of correcting any mismatches. An additional requirement for the E - H tuner is that the two arm controls should be nearly independent of each other's effects. This point is of great practical importance as it ensures rapid convergence to a match.

However, the E - H tuner rarely satisfies this second need [19]. Suppose the tuner is employed to match an arbitrary load with reflection coefficient Γ_L , the control settings will be those required to produce a reflection coefficient Γ_L^* when looking into the tuner's output port, if port 1 were matched Eq. (7). If the tuner produced Γ_t under these conditions, an approximate residual reflection coefficient with the mismatched load will be $(\Gamma_t - \Gamma_L^*)$. This approximation improves as Γ_t approaches Γ_L^* . The goal is to move point Γ_t to Γ_L^* in the complex reflection coefficient plane with minimum adjustments as possible. This means that adjusting the two controls should produce orthogonal Γ_t increments.

Thus, it is required that

$$\angle \delta \Gamma_H - \angle \delta \Gamma_E = \pm \frac{\pi}{2} \quad (15)$$

To match the input port, the input reflection coefficient

$$\Gamma_{in} = S_{11} + \frac{S_{21}^2 \Gamma_L}{1 - S_{11} \Gamma_L} = \frac{\Gamma_H + \Gamma_E}{2} + \frac{(\Gamma_H - \Gamma_E)^2 \Gamma_L}{4 - 2(\Gamma_H + \Gamma_E) \Gamma_L} = 0 \quad (16)$$

With suitably chosen reference planes, the reflection coefficient

$$\Gamma_L = \frac{1}{2} \left(\frac{1}{\Gamma_H} + \frac{1}{\Gamma_E} \right) = \frac{e^{j\phi_H} + e^{j\phi_E}}{2} \quad (17)$$

Eq. (17) proves the assertion that the E - H tuner can match any passive load $|\Gamma_L| < 1$. It is also clear that

$$\left. \begin{aligned} \delta \Gamma_H &= \frac{\delta \Gamma}{\delta \phi_H} \delta \phi_H = \frac{1}{2} j e^{j\phi_H} \delta \phi_H \\ \delta \Gamma_E &= \frac{\delta \Gamma}{\delta \phi_E} \delta \phi_E = \frac{1}{2} j e^{j\phi_E} \delta \phi_E \end{aligned} \right\} \quad (18)$$

Therefore,

$$\angle \delta \Gamma_H - \angle \delta \Gamma_E = \phi_H - \phi_E \quad (19)$$

It is obvious that the condition in Eq. (15) is generally not satisfied. A comparison with Eq. (17) and (19) shows that it is satisfied if and only if $|\Gamma_t| = \frac{1}{\sqrt{2}}$; a situation which will only occur in the final correction stages of an initial $|\Gamma_L| = \frac{1}{\sqrt{2}}$ mismatch. On the other hand, for small initial mismatches ($|\Gamma_L| \approx 0$), the difference in adjustments phases ($\phi_H - \phi_E$) $\approx \pi$ and separate movement of the E and H arms produce similar effects on Γ_L . In such cases, the tuning process is laborious as matching converges very slowly. A remedy for this problem is immediately noticed if Eq. (17) is rewritten as

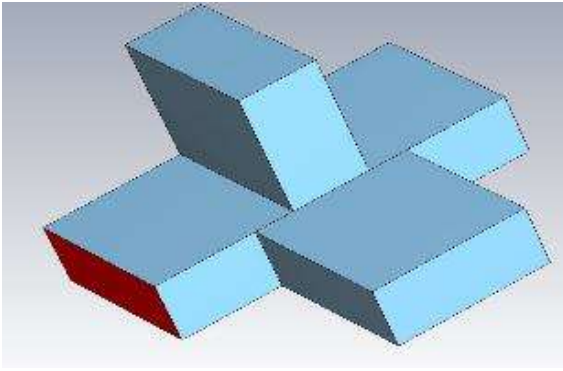


Figure 4. Simulated model of basic E-H Tuner

$$\Gamma_L = \cos(\phi^-) e^{j\phi} \quad (20)$$

where

$$\left. \begin{aligned} \phi^- &= \frac{\phi_H - \phi_E}{2} \\ \phi^+ &= \frac{\phi_H + \phi_E}{2} \end{aligned} \right\} \quad (21)$$

$$\left. \begin{aligned} \delta\Gamma_L(\phi^-) &= -\sin(\phi^-) e^{j\phi^+} d\phi^- \\ \delta\Gamma_L(\phi^+) &= j\cos(\phi^+) e^{j\phi^+} d\phi^+ \end{aligned} \right\} \quad (22)$$

Provided that ϕ^- and ϕ^+ are independent variables, Eq. (15) is always satisfied. Therefore, in practice, the two stubs of the $E-H$ tuner should always be adjusted simultaneously, either in opposite senses (ϕ^-) or in the same sense (ϕ^+) but never separately. In this way, a good match would be obtained in a few rapidly convergent adjustments, irrespective of the initial mismatch.

For a given load, there are two solutions, which are one symmetric to the other:

$$\Gamma_H = -e^{-\phi_H} = e^{-j[\angle\Gamma_L \pm \arccos(|\Gamma_L|)]} \quad (23)$$

$$\Gamma_E = -e^{-\phi_E} = e^{-j[\angle\Gamma_L \mp \arccos(|\Gamma_L|)]} \quad (24)$$

Therefore, to match the load, the shunt and series shorting stubs must be placed in one of two configurations by setting

$$l_{H,E} = \frac{\phi_{H,E}}{2\beta} = \lambda_g \frac{\angle[-\Gamma_{H,E}^{-1}]}{4\pi} = \lambda_g \left[\frac{\angle\Gamma_L \pm \arccos(|\Gamma_L|)}{4\pi} + \frac{1}{4} + n \right] \quad (25)$$

Simulation Study

The efficacy of the derived tuner control law was investigated using CST MWS and MATLAB. A model of a WR340 $E-H$ tuner was developed and simulated in CST (Figure 4 and plots in Figure 5a and 5b). Here, adjustable perfect electric walls were employed as the (ideal) E and H shorting stubs. The S parameters describing the tuner were extracted and studied.

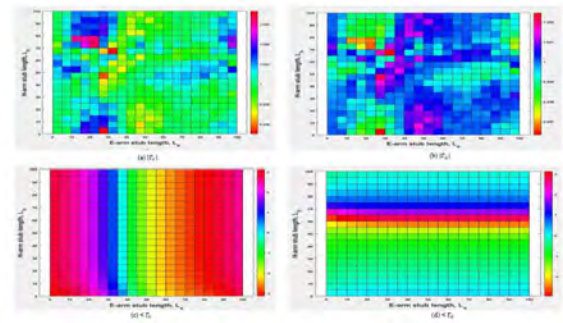
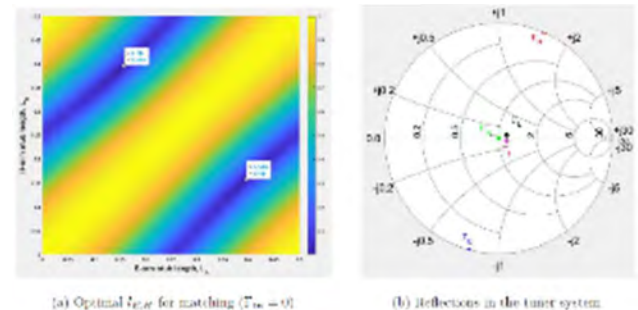

 Figure 5. Simulated study of the effect of the E-H tuners stub lengths on Γ_E and Γ_H .

Figure 5 shows the effect of the stub lengths of the $E-H$ tuner on Γ_E and Γ_H . It can be seen from the plots in Figures 5a and 5b that both arms produce near unity reflections with the fluctuations in the order of 10^{-3} . While these plots may suggest that the reflection in each arm is dependent on both stub lengths l_E and l_H , it is not so. This can be inferred by inspecting the phase plot of the generated reflections in Figures 5c and 5d.

Consider the E arm as an example (this holds true for the H arm as well). It can be seen that $\angle\Gamma_E$ varies solely with l_E and almost independently of l_H . This property is important for satisfying the matching requirement in Eq. (17). However, when the H backshort is near the main waveguide's inner wall i.e., $l_H \in [0, 25 \text{ mm}]$, $\angle\Gamma_E$ shows dependence on l_H . This can be attributed to the presence of higher order modes present at the junction. These fields decay with l_H increase. The $\angle\Gamma_E$ variation with l_E is uniform, almost flat with a $\pm 3^\circ$ range, and repeats with an approximately 85 mm ($\approx \frac{\lambda_g}{2}$) period. From the above, the following design remarks could be made:

- The minimum travel range of the backshort is $\frac{\lambda_g}{2}$.
- Each arm must be at least $(\frac{\lambda_g}{2} + 25 \text{ mm})$ to support backshort travel range and attenuation distance of higher order modes.
- A good backshort should produce near unity $|\Gamma|$ over the range of travel.
- A good backshort should also produce linear, predictable, and almost flat $\angle\Gamma$ variation over the range of travel.
- The two backshorts should produce reflections independently of each other.

Another study was conducted to examine the matching of different arbitrary loads using the discussed tuning law above. Consider a load characterized by $\Gamma_L = 0.0715 - j0.0252$.


 Figure 6. Simulated study of an arbitrary load characterized by $\Gamma_L = 0.0715 - j0.0252$.

It is seen from Figure 6b (and Figure 7b and 8b) that, to match the load reflection Γ_L (black dot), tuner must produce an internal reflection Γ_i (magenta dot) which is conjugated to Γ_L and is a combined effect of Γ_H (red dot) and Γ_E (blue dot). The pair of optimal $l_{E,H}$ was 0.158 mm and 0.395 mm, applied in any order. The plots in Figures 6, 7, and 8 confirmed the symmetry of the stub length solutions, as asserted by Eq. (24) and (25).

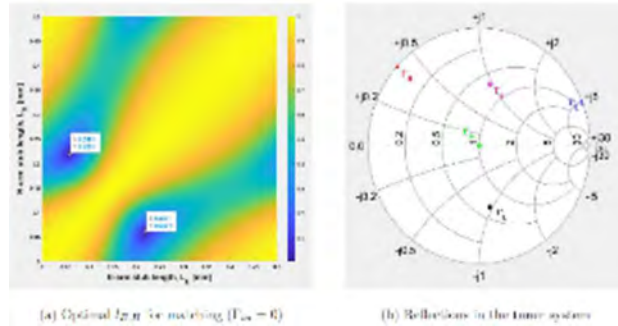


Figure 7. Simulated study of matching of an arbitrary load characterized by $\Gamma_L = 0.0967-j0.522$

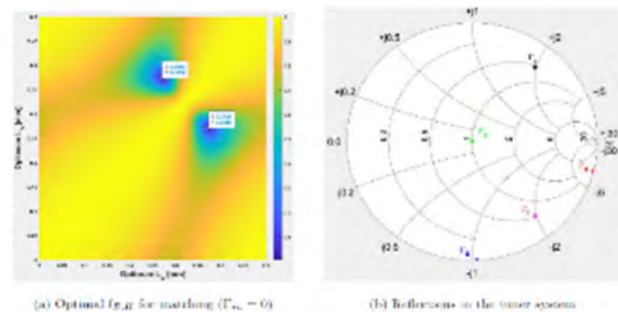


Figure 8. Simulated study of the matching of an arbitrary load characterized by $\Gamma_L = 0.5028-j0.6226$.

Furthermore, the study demonstrates the convergence issue in the $E-H$ tuner. With the very small initial mismatch, Figure 6a shows a diffused set of sub-optimal $l_{E,H}$ solutions. Therefore, the tuner would converge very slowly to the optimal solution. The tuner converges much more rapidly for a larger initial mismatch (Figure 8a).

Flowchart of Technical Setup and Tuning Processes

The flowchart illustrated in Figure 9 describes the process of automatic impedance tuning in the high-power S-band microwave network, starting with the signal generation. The microwave signal, at 2.45 GHz, originates from a high-power generator and is directed into a WR-340 waveguide containing a double-stub structure that functions as the impedance tuner. This proposed tuner is designed to automatically adjust impedance, crucial for efficient power transfer. As the signal travels through the tuner, a reflectometer continuously measures the reflection coefficient, identifying any impedance mismatches between the network and the load. These measurements are fed into a feedback control unit, which calculates necessary adjustments to minimize impedance mismatch. The closed-loop feedback system automatically aligns the tuner's stubs based on these calculations, achieving real-time adjustments that ensure stable and temperature-independent performance. Hence, once the

tuning process is optimized, the system can maintain impedance matching and adapt to any network changes through continuous monitoring.

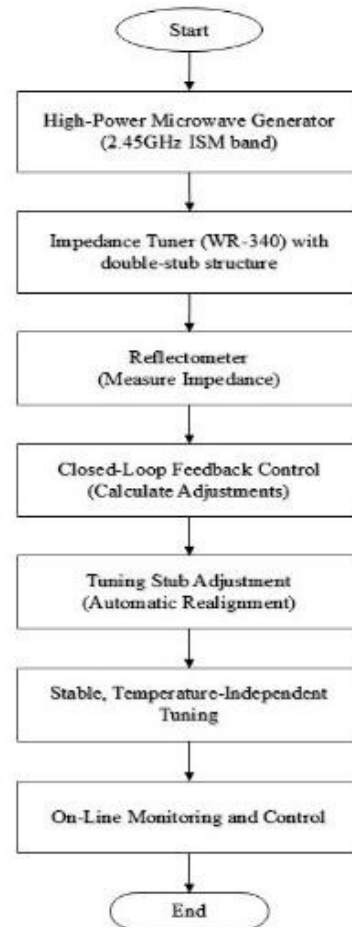


Figure 9. Flowchart of Setup and Tuning Processes

RESULTS AND DISCUSSION

Waveguide Backshort

In the $E-H$ tuner, movable backshorts are installed in the E and H arms to produce variable reactive loads which adjust the device's input or output impedances. An ideal backshort totally reflects incident power and modulates the reflected wave's phase by a desired amount when moved. Other desired characteristics include wide impedance bandwidths, non-distorted of local fields, smooth sliding motion, and coincidence of electrical and mechanical surfaces. Additionally, it must be easy to fabricate and highly repeatable. In practice, good backshorts are characterized by voltage standing-wave ratio (VSWR) and linear phase variation of reflected waves with frequency and movement [20, 21]. Proper and predictable operation of the $E-H$ tuner relies on having sliding shorts that offer a return loss of less than 0.3 dB across the full waveguide band. A poor backshort might produce abrupt phase variations due to in-band resonances. Therefore, designing backshorts that exhibit good RF performances is desired.

Backshort Loaded Waveguide

The simplest structure of a tunable rectangular waveguide backshort is a rectangular block made of conducting material, which slides in the guide. The block's position is varied by a push/pull or micrometer shaft. Movable short circuits are broadly categorized as either contacting or non-contacting. Contacting backshorts have galvanic contacts with the waveguide inner walls while non-contacting ones are suspended in the guide. Contacting shorts suffer inevitable wear due to repeated sliding friction, resulting in unreliable contacts and limited backshort lifetimes. Contact shorts also plagued with resistive losses at DC contact points. Careful rounding of contact areas together with heavy plating of backshort and waveguide walls have proven to be effective methods of mitigating issues with contacting shorts [21]. Non-contacting backshorts are generally immune to the drawbacks of the contacting ones.

In this work, only non-contacting short-circuits were considered as design candidates. The simple non-contacting sliding block, illustrated in Figure 10, exhibits poor electrical performances. First, there are losses due to power leakage into the gap beyond the short surface. The leakage is exacerbated in heavy use due to the wear of block surfaces and guide walls. This is because contact strips act as antennas to couple portions of the incident power into the small waveguides formed between matching of an arbitrary load characterized by $\Gamma_L = 0.5028 - j0.6226$ the backshort body and guide walls [21].

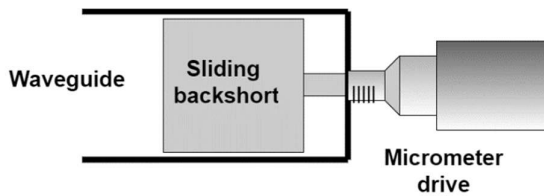


Figure 10. Basic non-contacting shorting block in waveguide

Additionally, erratic contact between the surfaces produces an equivalent short-circuit position which deviates randomly with respect to the physical short position i.e., front face of block. This causes instability of the reflected wave phase and deteriorates short performance. The phase instability corresponds to spurious resonances in the short's response in the operating band, which are caused by propagating electromagnetic wave modes within the short-filled guide section [22]. The resonant dips in the reflection coefficient are accompanied by transmission peaks, forcing a substantial portion of the microwave energy beyond the backshort. This could lead to device damage or environmental /health hazard if the leakage is not controlled. Excitation of these modes depends on the block wobble with respect to its ideal centered position. To guarantee good electrical performance, very tight machining tolerances are mandatory for backshort dimensions, especially at very high frequencies.

Modal Analysis

The starting point for the design of an effective non-contacting adjustable backshort is to investigate the nature of the wave

modes that can exist in the short-loaded waveguide. The analyzed structure is a perfectly conducting brick short of length L , placed in a rectangular waveguide section with dimensions a and b (Figure 11).

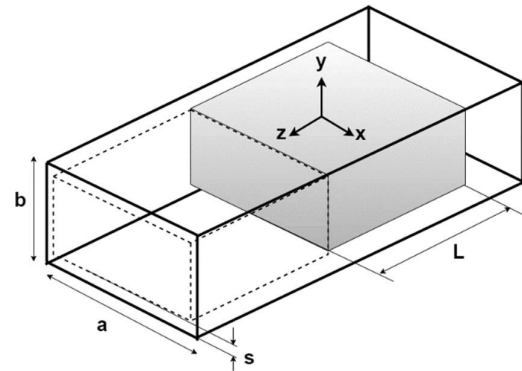


Figure 11. Analyzed plunger filled waveguide structure

In an ideal case, the plunger is centered within the waveguide cross-section such that its sides were parallel with and separated equally from the guide walls by a small gap s . The mean gap perimeter is $p = 2(a + b) - 4s$. Assuming a very small gap with respect to the guide perimeter ($s \ll 2(a+b)$), modes existing in the coaxial structure are categorized into the *TEM*, "low frequency" *TE*, "high frequency" *TE*, and *TM* modes. The *TEM* mode has a DC cutoff and can propagate in the structure once it is excited. Several low frequency *TE* modes also propagate in the S-band. Such modes are designated TE_{m0} as m free-space wavelengths fit in the gap at their cutoff frequencies f_c and radial fields are essentially homogeneous ($n = 0$). The TE_{m0} modes have cutoff wavelengths λ_c approximated by [22, Eq. (26)] and exist in degenerate pairs for each m (i.e., they have the very close f_c). The identical members of these mode couples which are designated TE_{am0} and TE_{bm0} have field patterns which appear axially rotated around the guide's axial axis by $90/m$ degrees i.e., $TE_{bm0} \approx TE_{a0m}$ [22].

On the other hand, the *TM* modes as well as high-frequency *TE* modes are oscillatory in the radial direction with half-wavelength periodicity. The cutoff wavelengths of these modes are approximated by [22, Eq. (27)]. Here, n free-space half-wavelength are packed in the gap at corresponding cutoff frequencies. Eq. (27) implies that, for snug fit non-contacting backshorts (e.g., s in the order of millimeters), these modes are excited at millimeter-wave frequencies. Therefore, in this analysis, only the *TEM* and the low frequency *TE* modes were considered significant.

$$\lambda_c = \frac{p}{m}, m = 1, 2, 3.. \quad (26)$$

$$\lambda_c = \frac{2s}{n}, n = 1, 2, 3.. \quad (27)$$

Mode properties such as cutoff frequency and field patterns were studied for the loaded WR340 waveguide section ($a = 86.36$ mm, $b = 43.18$ mm) with $s = 2$ mm and the short in its ideal position. Effects of the plunger waggle were also studied. The study was assisted by the CST MWS EM simulator. To simulate the deviations, the backshort position was perturbed by x -offset, and y -offset: $\Delta x = 0.5$ mm, $\Delta y = 0.5$ mm respectively. Conversion

coefficients, K , from the hollow guide's dominant TE_{10} mode at 2.45 GHz, were also obtained. Results indicated that the TEM mode and up to two degenerate TE mode pairs were present in the structure. The E-field patterns and transverse surface currents of the TE modes are illustrated in Figure 12.

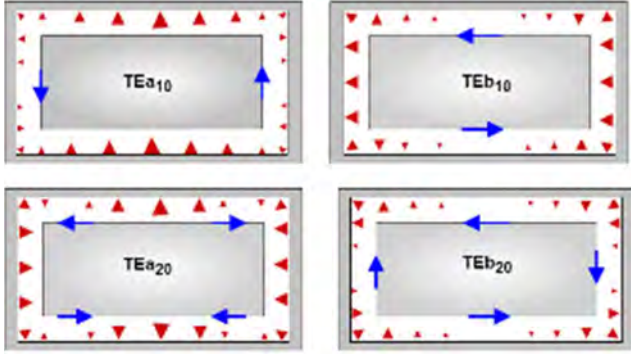


Figure 12. Propagating TE_{m0} in waveguide section with centered backshort.

The red arrows indicate the E-field of the modes while the blue ones show the transverse surface current maxima. The basic mode properties of the modes are summarized in Table 1. The parameters marked with subscripts *min* and *max* refer to recorded limits of the wobble effects in the S-band while *ideal* corresponds to the ideal backshort position. In the ideal case, only the TEa_{10} mode was strongly excited. However, backshort perturbations destroyed mode symmetries and excited all/most modes with their cutoff frequencies varying by at least 50 MHz (Figure 12, Table 1). Cutoff frequencies of the TE_{20} modes were within or close to the 2.4 - 2.5 GHz ISM band. This is undesirable for many applications.

Table 1. Properties of propagation modes.

Parameter	TEM	TEa_{10}	TEb_{10}	TEa_{20}	TEb_{20}
f_c (GHz), ideal	0	1.197	1.205	2.393	2.409
f_{cmin} (GHz)	0	1.140	1.187	2.338	2.397
f_{cmax} (GHz)	0	1.201	1.259	2.403	2.448
Δf_c (MHz)	0	61	72	65	51
K (dB)	-	-6.3	-	-	-

Backshort Resonances

Excited propagating modes that are reactively terminated at the backside of the backshort produce resonances trapped within the short-filled guide section. The spurious resonances manifest as narrow dips/peaks in the structure's transmission coefficient frequency response and are converted to the TE_{10} mode behind the backshort under certain conditions. The trapped modes are characterized by an approximately q of waveguide half-wavelengths along the backshort length where $q = 0, 1, 2$,

Hence, each resonance mode is marked by an additional subscript, q : TEM_q , TEa_{m0q} , TEb_{m0q} . The loaded guide can be viewed approximately as a transmission line that is open on both ends and having an effective $(L+\Delta L)$ length where the extension

ΔL is the length correction factor accounting for fringe evanescent fields in the empty waveguide [22]. The relation between the cutoff frequency f_c of a mode in an ideally loaded guide section and a related resonance frequency f_r can be approximated by Eq. (28).

$$f_r^2 = f_c^2 + \left(\frac{cq}{2(L+\Delta L)}\right)^2 \quad (28)$$

where c represents the speed of light.

To study these resonances, a short length $L = 50$ mm was simulated. In its ideal position, the backshort-filled section excites the TEa_{10} mode which exhibited a single broad $|S_{21}|$ peak near the top of the band (at 2.953 GHz) signifying the strong coupling to the empty guide (Figure 13).

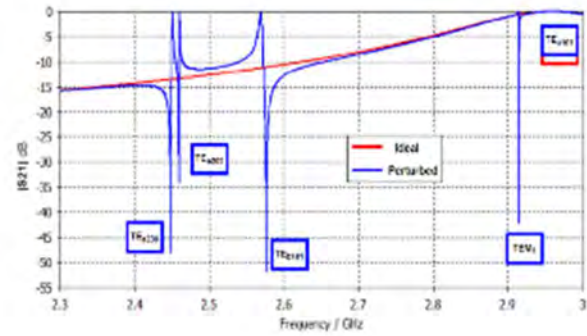


Figure 13. Transmission response of backshort-filled guide for ideal and perturbed short positions ($\Delta x = \Delta y = 0.5$ mm).

This mode is designated TEa_{101} as it resonated such that approximately one half of the standing wavelength was distributed along L . Offsetting the backshort preserved the broad $|S_{21}|$ peak response but introduced four extra peaks with near total transmission. The resonances at 2.57 GHz and 2.91 GHz corresponded to TEM_1 and TEb_{101} respectively. Varying L changed f_r of each of the above three modes (Figure 14b). Hence, by appropriately choosing L , the peaks could be shifted out of the operating band. The two remaining resonances corresponded to TEa_{200} and TEb_{200} and are seen as closely spaced peaks located in the ISM band at 2.446 GHz and 2.456 GHz respectively. Unlike the first three, the field strengths of modes for which $q = 0$ do not change along backshort length and can hardly be moved out of the band by varying L (see Eq. (28), Figure 14b). Their standing wave patterns could be viewed as setup from two axially uniform TEM waves travelling around a backshort that is terminated on both ends by non-ideal open circuits [22]. Thus, in reality, f_r of TEa_{200} and TEb_{200} are very loosely dependent on L and are approximated by Eq. (29).

$$f_r^2 = f_c^2 + \left(\frac{c\gamma}{2L}\right)^2 \quad (29)$$

where $\gamma < 1$ is an adjusting coefficient compensating for effects of the fringing fields. For $L \geq 30$ mm in WR-340 guides, γ values are in the order of 10^{-2} , hence $f_r \approx f_c$ [23]. The two resonances persist with the simple block backshort since the only remaining parameter available is the gap thickness whose variation has limited effect on f_r .

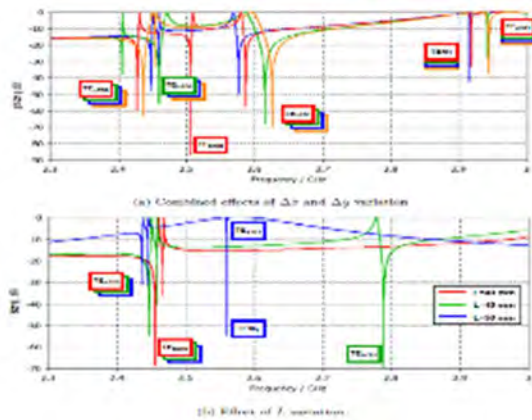


Figure 14. Parametric study of effect of perturbation and L variation. Constant study values were ($\Delta x = \Delta y = 0.5\text{mm}$, $L = 50\text{mm}$)

Standard Microwave Choke

The microwave choke is a frequently used backshort design which is implemented as a folded two section quarter-wave transformer as illustrated in Figure 15.

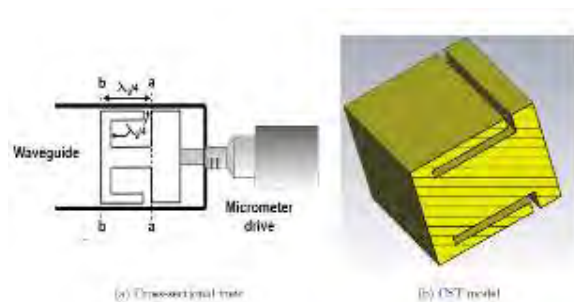


Figure 15. The standard microwave choke.

The choke is made up of pairs of low and high impedance sections and behaves as a band-stop filter. The inner quarter-wavelength long section transforms the terminating short to an ideally infinite impedance point at the plane $a-a$, preventing any flow of current in the gap. The outer section then transforms the open-circuit impedance at $a-a$ into a short-circuit plane $b-b$ at the front of the backshort. The *wavelength* mentioned here refers to the choke inner λ_g of the mode to be suppressed. Therefore, the standard microwave choke can be designed to suppress either one mode or a pair of degenerate modes. Assuming operation in the ideal position, only the strong coupled TE_{a10} mode is excited.

Therefore, the dimensions of choke sections are designed towards suppressing the TE_{a10} . In the event of a waggle, such a choke will also attenuate the TE_{b10} hybrid twin. The TEM is not suppressed but its resonance can be tuned out by properly choosing L leaving an often tolerable off-resonance remnant in the operating band. However, the resonances related to the TE_{a20} and TE_{b20} modes are practically fixed at f_c and cannot be attenuated by this choke type. This is because a mode's λ_g approaches infinity near its f_c . From Table 1, the $f_c \approx f_r$ of the pair are in the band of interest. Hence, a practical S-band choke to suppress such modes is electrically non-existent. Furthermore, as discussed in the above

section, this mode can neither be tuned out by adjusting L . The transmission performance of an optimized choke design, studied in CST MWS, is presented in Figure 16.

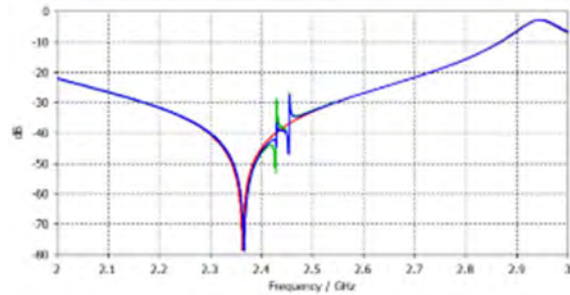


Figure 16. Transmission coefficient performances of an ideally positioned and perturbed standard microwave choke.

Slotted Microwave Choke

A possible solution to remedying the lingering resonance modes in the microwave choke is by forcing the modes to behave like TEM modes and then tuning them out. This idea is because TE modes produce non-zero transverse surface currents on the backshort surface as shown in Figure 11. Introducing longitudinal slots at current maxima limits transverse current flow resulting in the modulation of the TE modes into near- TEM waves and consequently, the grouping of their wavelengths [23, 24]. The choke is then designed targeting TEM -like modes, which can be suppressed by tuning L . Although Figure 11 indicates that three broadface slots and one in the end-wall suffices for blocking transverse currents of all four TE modes considered, more slots are in fact required as backshort waggle may cause current maxima to shift.

With the aid of CST MWS, a series of computer simulations were performed to understand the isolation properties of the slotted backshort and maximize its resonance-free bandwidth. The slotted choke design had five slots along the broadside and three on the end-wall (Figure 17a).

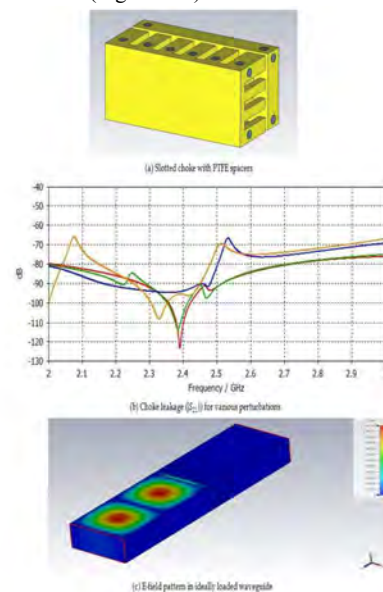


Figure 17. Design of the slotted choke and its electrical performance.

Each slot was 5mm wide and span the entire choke length l_c . The resulting backshort was $L = 49.3$ mm long and had a choke of length $l_c = 25.2$ mm with an input situated 30.2 mm from the backshort's front face. PTFE spacers meant to secure the backshort in the ideal position were incorporated in the design.

Anticipating a good fabrication process, the slotted backshort was redesigned and optimized for an air gap, $s = 0.1$ mm. Key dimensions were $L = 42.5$ mm and $l_c = 22.5$ mm. Thin cylindrical PTFE space were added to design which serves to secure the choke in the ideal position. Simulated isolation performances in the ideal position as well as with some selected perturbation combinations, are shown in Figure 17b. The figure shows that the backshort produced excellent RF isolation, supporting the design concept. The leakage is less than -80 dB in the 2.4 - 2.5 GHz ISM band and -65dB over the whole S-band. Simulation thus supports the suggested concept. Inspection of field patterns showed that the maximum E-field strength, E_{max} in the ideally loaded guide was approximately 214 kV/m for 6 kW (Figure 17). An empty guide, $E_{max} = 71$ kV/m, giving a reference power handling capacity of 4 MW [23]. For all simulated backshort positions, the highest recorded E_{max} in the structure was 463 kV/m ($\Delta x = 0$, $\Delta y = 0.1$). Thus, the reference power handling capability is expected to be reduced by about four hundred-fold (~ 94 kW), which suffices for the target application.

E-H Tuner

The complete design of the *E-H* tuner is illustrated in Figure 18.

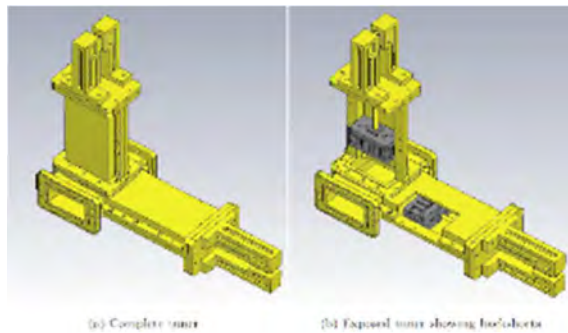


Figure 18. Design of the slotted choke and its electrical performance.

It consists of a magic-T with all four of its arms terminated in CPR340F groove flanges. The main waveguide was $\lambda_g \approx 173$ mm long and both the *E* and *H* plane arms measured 230mm when referenced to the inner waveguide wall. The length of the tuning arms was designed to support the full travel range of the tuner arms, $\frac{\lambda_g}{2}$, the backshort length, L and support clearance for both the backshort-actuator connection and higher-order mode zone near main guide wall. The tuner also had a backshort in each arm which was adjusted by a linear actuator. The backshort-actuator assembly is suspended within the magic-T through a flange cover by a thin cylindrical rod and a pair of braces. The actuator employed in this work is the Actuonix L12-I micro linear actuator.

This actuator is very compact and features an onboard controller which makes it highly configurable. The 100 mm stroke L12-I actuator possesses a powerful 100:1 gear drive train producing a maximum force of about 42 N at a speed of about 13 mm/s. This allows a smooth and accurate positional tuning over the travel range. Additional information on L12-I can be found in [25]. The magic-T was split four ways for easy realization. All parts of the tuner except the actuator are being milled from aluminum.

An experimental sliding backshort based on the slotted choke design was fabricated and tested. The short block was manufactured in two parts which were mill-machined from stock aluminum and fastened together with screws and dowels. The backshort also had a receptacle for holding the head of the linear actuator (Figure 19).

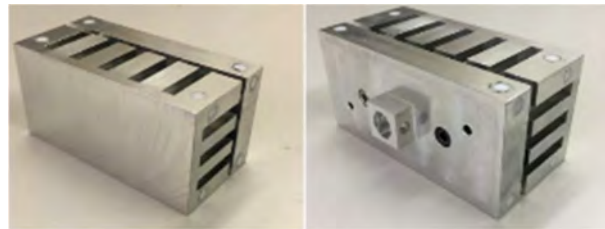


Figure 19. Fabricated waveguide non-contacting backshort with PTFE spacers.

The PTFE spacers were reduced to four per side to minimize friction and keep force required to push/pull backshort within the 42 N max force rating of the actuator. As shown in Figure 20, the short was inserted into a section of WR-340 waveguide and the magnitude of its reflection coefficient was measured in the 2.2 - 2.7 GHz range using an Agilent E5071c VNA and coax-to-waveguide (C2W) transitions. Figure 21 compares the measured response and the corresponding simulated S_{21} . Although there was a discrepancy between the two responses, the agreement was reasonably good and the similarity in the two reflection behaviors over the band is evident.

The differences between the simulated and measured results can be attributed to manufacturing defects in the short, minor leakages in the gaps, and mode conversions in the waveguide. Nonetheless, the backshort performed very well in the 2.4 - 2.5 GHz band recording less than -60 dB transmissions across the operational frequency range. The mid-band response was -78 dB. The Magic-T junction for the tuner, at the time writing is under fabrication. The complete system shall be presented in future work.



Figure 20. Test setup for measurement of the isolation properties

of the designed and fabricated backshort. The short is logged in the section of WR340 waveguide sandwiched between the transitions

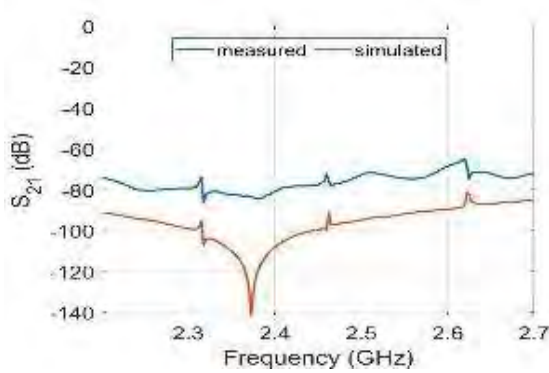


Figure 21. Fabricated waveguide non-contacting backshort with PTFE spacers.

A comparative assessment of this work with other existing tuners was performed and presented in Table 2. This highlights the merits of this work in terms of design and overall performance.

Table 2. Comparative assessment of this work with similar tuners.

Ref.	Tuner type	State	Freq.	Result
[26]	single-stub	n/a	V-band	≤ 1.0 dB with the insertion loss
[27]	triple-stub	n/a	X-band	~ 29 ETR
[28]	single-stub	automatic	W-band	≤ 1.0 dB with the insertion losses
[29]	triple-stub	n/a	S-band	≤ 1.0 dB
[30]	triple-stub	n/a	S-band	≤ -10 dB (Γ_L)
This work	double-stub	automatic	S-band	-78 dB transmission

Limitation and Future Direction

While the proposed tuner demonstrated stable and temperature-independent performance at 2.45 GHz, it was however limited to this specific tuning range (S-band), which restricts its versatility in applications requiring broader frequency adaptability. Furthermore, improvements in the precision of the tuning mechanism could enhance performance in environments with rapid or extreme impedance variations. Future research works could focus on optimizing the tuner's design to support multi-band frequency ranges, which would extend its usability across various industrial microwave applications. Exploring materials and components that further improve thermal resilience could also enhance stability under more demanding conditions. Implementing more advanced feedback algorithms other than the closed-feedback system may improve real-time adaptability,

providing faster response times and finer control in dynamic operational settings.

CONCLUSIONS

In this work, the design of an impedance tuner developed to be used in an automatic waveguide network tuning system for high power microwave networks has been presented. Simulations show that the tuner can match a wide range of network impedances. A slotted backshort for the tuner has been designed, fabricated, and tested. The plunger filled waveguide structure was analyzed, and the properties of the various propagation modes were also examined. The measured results show that the backshort is capable of reflecting incident power strongly and recording an impressive -78 dB transmission at 2.45 GHz. The Magic-T hybrid, at the time of writing, is under construction. The complete automatic tuning system shall be presented in future work. The system shall be suited for on-line monitoring and control of network parameters in industrial waveguide applications.

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