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A BROADBAND MICROWAVE GAIN EQUALIZER

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Abstract—In this paper, we propose a new kind of broadband microwave gain equalizer in microstrip circuit. The equalizer uses open stepped impedance resonators (SIRs) to increase the adjust parameter so that the equation curve can be more flexible. Simplified topology of the gain equalizer is used to make the match net easier. The power distribute on each resistance is analyzed and the error analysis of the resistance values is done. Finally we design and manufacture a gain equalizer, and the measured results show that the equalization curve meets requirements well and prove that this structure is practical and effective.

1. INTRODUCTION

Gain equalizer is one of the important parts in Microwave Power Module (MPM), which is used to flatten the output gain of traveling wave tube (TWT). Nowadays, MPM is widely used in many potential system applications, such as radar, communication, aircraft and electronic warfare [1–4], which make a higher demand on the volume and weight of the equalizer. The miniaturization of it becomes one of the focus studies.

Gain equalizer used to be made of waveguide or coaxial with electromagnetic wave absorber in its cavity resonators [5–8]. However, this structure is large and not convenient to integrate. Because microstrip circuit is easy to process with a small size, it is very useful in making gain equalizer. Traditional microstrip gain equalizer contains several resonant branches and resistances connecting the branches to the main transmission line. However, these resonant branches have only two parameters to adjust, the width and length. This limits its ability to control the equalization curve [9–12]. To

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solve this problem, the gain equalizer proposed here is fulfilled by open stepped impedance resonators (SIRs) with thin film resistance integrated. This structure provides a new adjustable parameter, the ratio of SIR impendence, which is useful in controlling the equalization curve. Another important parameter of gain equalizer is the return loss at I/O port. Luckily there are many researches about how to get a good match in wide band, which may give us a lot of references [13–18]. In this paper, we propose a simple topology to get the I/O port matched. In this way, the design parameters are reduced a lot, and the match process becomes more convenient.

2. THE DESIGN METHODOLOGY OF GAIN EQUALIZER

2.1. The Topology of Gain Equalizer

We begin the analyses of the gain equalizer by studying a single resonator. There is a simple resonator composed of resistance, inductance and capacitance (as shown in Fig. 1).

The input impedance and admittance are defined as Z, which can be expressed as follows:

$$Z = R + \frac{1-\omega^2 CL}{j\omega C}; \quad |Z| = \sqrt{R^2 + \left|\frac{1-\omega^2 CL}{\omega C}\right|^2} \tag{1}$$

The scattering matrix can be obtained as follows:

$$\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \frac{1}{2+Z_0/Z}\begin{bmatrix} -Z_0/Z & 2 \\ 2 & -Z_0/Z \end{bmatrix} \tag{2}$$

So its transfer function is:

$$H(\omega) = \frac{2\times(j\omega CR + 1 - \omega^2 CL)}{2\times(j\omega CR + 1 - \omega^2 CL) + j\omega C Z_0} \tag{3}$$

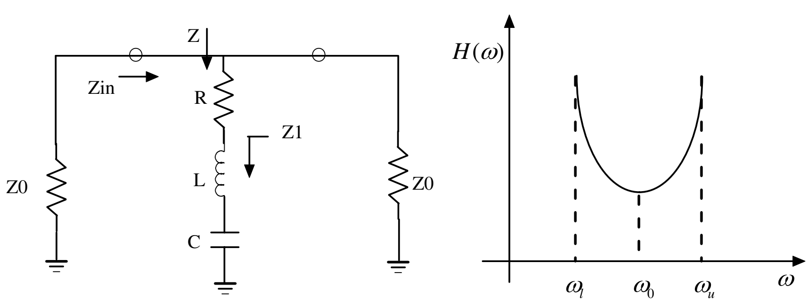

Figure 1. Circuit model of resonator and its frequency response.

When resonance structures are united with a good match, their total transfer function is:

$$H_{total}(\omega) = \sum_{i=1}^{n} H_i(\omega) \tag{4}$$

Putting a series of resonators together, carefully choosing resonance frequency and resonance Q value, and designing the match net well, an appropriate equalized response can be obtained. This is the basic principle of the gain equalizer.

For simplifying the match process, a simpler topology of the gain equalizer is used here (Fig. 2). Compared to the gain equalizer reported in [9, 10], this match net removes the match sections at the I/O port and only uses the quarter wave transmission line between two adjacent resonators. To avoid the couplings of the two adjacent resonators, the quarter wave transmission line can be added by integer multiple of the wavelength.

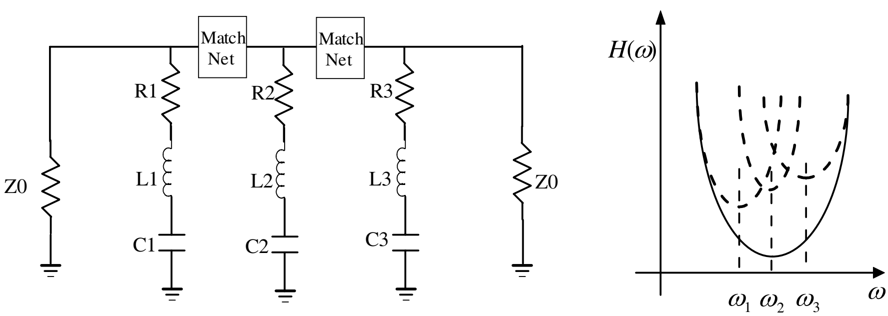

Figure 2. Circuit model and frequency response of gain equalizer.

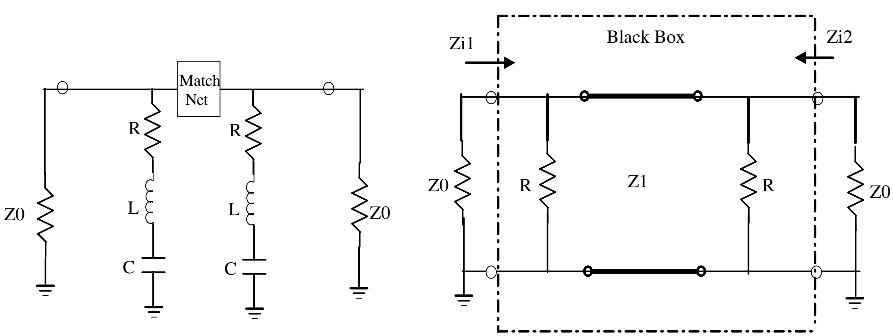

Figure 3. The simplified topology and its equivalent circuit model when resonate.

Fig. 3 shows a description of the simplified topology and its equivalent circuit model as an example. Regarding the equalizer circuit model as a black box, the transmission matrix can be obtained as follows by using microwave network theory [19].

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \dfrac{jZ_1}{R} & jZ_1 \\ j\dfrac{1+(Z_1/R)^2}{Z_1} & \dfrac{jZ_1}{R} \end{bmatrix} \tag{5}$$

The image impedance of the black box is:

$$\begin{aligned} Z_{i1} &= \sqrt{\frac{AB}{CD}} = \frac{Z_1}{\sqrt{1+(Z_1/R)^2}} \\ Z_{i2} &= \sqrt{\frac{DB}{CA}} = \frac{Z_1}{\sqrt{1+(Z_1/R)^2}} \end{aligned} \tag{6}$$

So the match condition can be calculated as follows:

$$\frac{Z_1}{\sqrt{1+(Z_1/R)^2}} = Z_0 \Rightarrow Z_1 = \frac{Z_0}{\sqrt{1-(Z_0/R)^2}}, \quad R > Z_0 \tag{7}$$

This calculated value is derived with the ideal model, ignoring the influences of discontinuity and edge capacitance at the open end. The accurate value can be optimized by using simulation software.

2.2. Microstrip Open SIR in Gain Equalizer

The basic structure of an open SIR is shown in Fig. 4. A distinct feature of the SIR is that the bandwidth of the resonator can be adjusted by changing the impedance ratio R_z ($R_z = Z_2/Z_1$), which is a useful parameter and can help in making the attenuation curve more flexible. Assume that the electrical lengths of the two transmission lines are the same. We give two important expressions as follows [20].

The resonance condition:

$$\theta_1 = \theta_2 = \mathrm{ctg}^{-1}\sqrt{R_z} \tag{8}$$

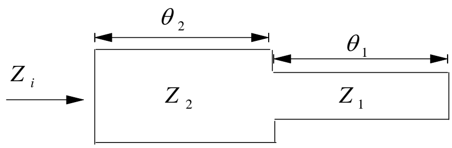

Figure 4. The structure of open SIR.

And the spurious frequency f_{SA}:

$$\frac{f_{SA}}{f_0} = \frac{\theta_{SA}}{\theta_0} = \frac{\pi - \theta_0}{\theta_0} = \frac{\pi}{-\mathrm{ctg}^{-1}\sqrt{R_z}} - 1 \tag{9}$$

The fundamental resonance frequency is represented as f_0, while the lowest spurious frequency is f_{SA}. Consider $\theta_1 = \theta_2 = \theta_0$.

It is shown that if we change the value of R_z, not only the resonance condition but also the distance between fundamental and spurious frequency will be changed. The uniform impedance resonator (UIR) used in [9–11] is one of the conditions in SIR ($R_z = 1$). So the analysis here is still useful in UIR.

To explain it more vividly, simulation is done by HFSS. By changing the ratio of impedance in SIR, while using the Equation (8) to get the electrical length of the SIR, the bandwidth of the notch can be easily controlled (Fig. 5). From Fig. 5, we see that the bandwidth

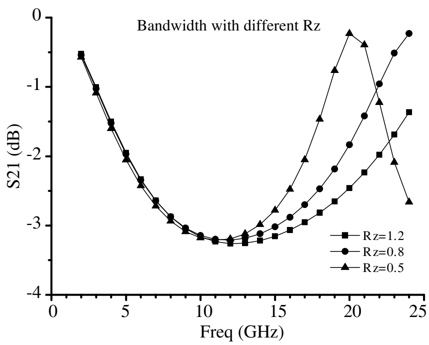

Figure 5. Bandwidth control of the microwave gain equalizer.

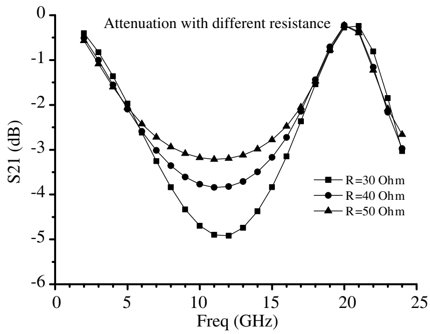

Figure 6. Attenuation control of the microwave gain equalizer.

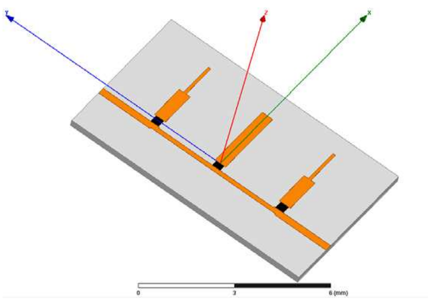

Figure 7. Simulation model of equalizer in HFSS.

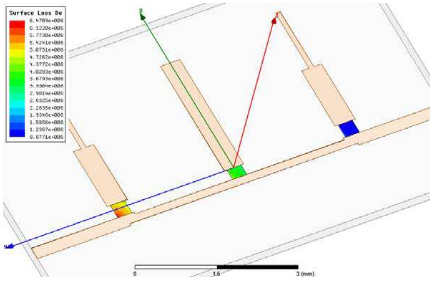

Figure 8. The surface power loss density distribute on resistances.

grows with the increase of ratio of impedance in SIR.

Keep the ratio of SIR unchanged so that the physical length of the SIR will be the same when the SIR resonates. From Equation (2), we know that if we adjust the value of the resistance which absorbs the power, the transmission coefficient can be controlled. The results are shown in Fig. 6. It can be seen that the attenuation grows when the value of absorb resistance decreases. Using the two methods introduced above we can control the equalization curve easily.

3. SIMULATION AND MEASUREMENT

3.1. Simulation and Manufactured

To prove the usefulness of these strategies introduced above, a gain equalizer is designed and manufactured. The simulation is done in a 3D-electro-magnetic simulation software HFSS. Fig. 7 shows the simulation model of the gain equalizer. The substrate is Al_2O_3 ($\varepsilon_r = 9.8$, thickness $h = 0.254\,\mathrm{mm}$). The microstrip is gilt, and the square part is TaN thin film resistance. The thin film resistances are integrated on a substrate directly, which can reduce the influence of parasitic parameters.

The software HFSS can calculate the electromagnetic field distribution on the equalizer. From the electromagnetic field we can get the Poynting vector in the area where the resistance is. Then the surface loss density can be obtained from the Poynting vector. At last we integrate the loss density function in the area of each resistance to calculate the power absorbed by each resistance. This process can be expressed as:

$$P_{loss} = \iint\limits_{x,y} \mathrm{Re}((\bar{E} \times \bar{H}^*) \cdot \bar{n}) dxdy \tag{10}$$

The surface power loss density and the power absorbed by each resistance are respectively shown in Fig. 8 and Fig. 9. The input power here is 1 W. Fig. 8 shows that the first resistance near the input port has the highest surface power loss density. To improve the power capacity of the gain equalizer, the resistance near input port should have a higher power capacity. From Fig. 9, we know the exact power absorbed by the resistance, which can be useful in choosing a suitable power capacity of the resistance.

Also we have done the error analysis of the resistance values. We do $+/-5\%$ parameter scanning of the resistance values, while keeping the other parameters unchanged. The results are shown in Fig. 10. When the resistance value changes, the S_{21} of the network changes

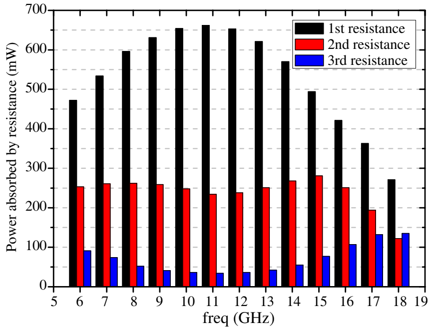

Figure 9. The power absorbed by each resistance.

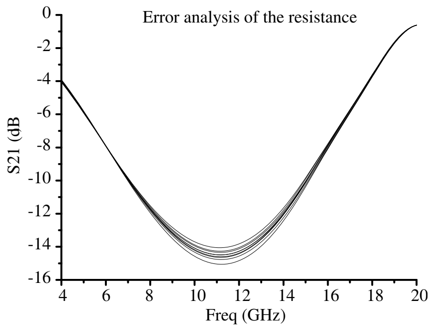

Figure 10. Error analysis of the resistance values.

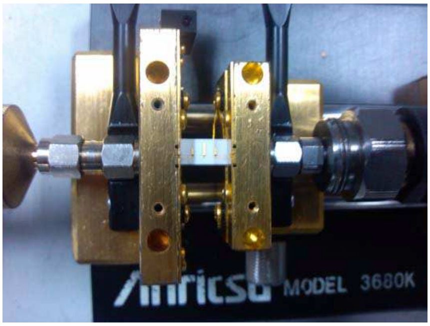

Figure 11. Test fixture Anritsu3680K.

Figure 12. Photo of the microwave gain equalizer.

little, and the maximum attenuation of S_{21} changes within $+/-0.5$ dB. It Shows that the design has a good stability.

3.2. Measurement and Analysis

The measurement is made with the help of vector network analyzer of Agilent and test fixture Anritsu3680 K (Fig. 11). A photograph of the equalizer is shown in Fig. 12.

The comparison of simulated and measured results is shown in Fig. 13, which shows that the measurement result of the equalization curve meets the requirements well with only a few differences about 0.5 dB near the area where the maximum attenuation is. From the analysis in Fig. 10, we know that the difference is caused by the processing errors of the resistance.

Figure 14 shows the whole measurement results of gain equalizer.

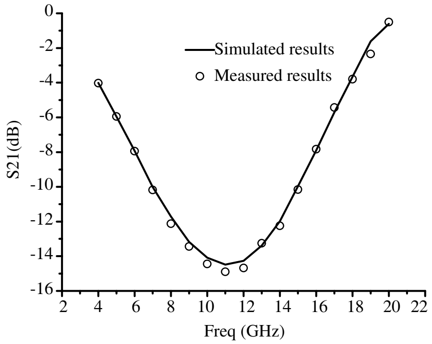

Figure 13. Compare of simulation result and measured result.

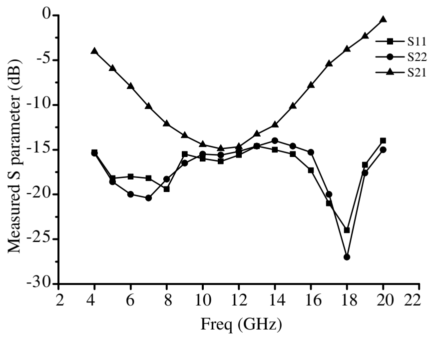

Figure 14. The measured result of gain equalizer.

The insertion loss is about 0.5 dB at 20 GHz and the max attenuation about 15 dB at 11 GHz. The value of equalization is about 14.5 dB. We get a good match both at the input and output ports, and the return loss at the two ports are both better than 15 dB, from 4 GHz to 20 GHz.

4. CONCLUSION

In this paper, a new kind of broad band microwave gain equalizer using open stepped impedance resonators is proposed. It has more advantages of adjustable frequency and quality factor compared to the microstrip branch resonator used before. Thin film resistance is made directly on the substrate which is more accurate than the paster resistance. This structure is small and convenient to be integrated. Good matches are got at the input and output ports with a new topology. This method is simpler and has fewer design parameters. The return losses at I/O port are both better than 15 dB from 4 GHz to 20 GHz. Also the power absorbed by each resistance and the error analysis of the resistance values are analyzed. These results are useful in the design process to estimate the power capacity of the gain equalizer. The measurement results meet requirements well, proving that this structure is practical and effective and will propel the researches in gain equalizer in MPM.

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