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G-Shaped Defected Microstrip Structure Based Method of Reducing Crosstalk of Coupled Microstrip Lines

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Abstract—The suppression of crosstalk by combining the defected microstrip structure (DMS) with step-shaped transmission lines is proposed to address the problem of crosstalk between microstrip lines of the printed circuit board. This method suppresses the crosstalk between the microstrip lines by constructing two step-shaped coupled microstrip lines and etching the designed G-shaped DMS on one of the microstrip lines. Simulation and actual measurement results show that the combination of G-shaped DMS and step-shaped transmission line can effectively suppress crosstalk and reduce the far-end crosstalk by approximately 20 dB in the frequency range of 4–5 GHz. The actual measurement results in the vector network analyzer coincide with the high-frequency structure simulator simulation results.

1. INTRODUCTION

The rapid development of electronic products has continuously increased the wiring density and operating frequency of integrated circuits and printed circuit board (PCB) [1]. The crosstalk between microstrip lines is increasingly becoming serious. Crosstalk, which is one of the key affecting factors of the quality of signal transmission, will seriously endanger the normal operation of the system. Therefore, suppressing crosstalk has become a major issue in recent years.

At present, the proposed crosstalk suppression methods mainly include increasing the coupling distance of the microstrip lines [2, 3], adding a serpentine wiring structure between the microstrip lines [4], and grounding through grounding holes at both ends of the protective wiring [5–7]. However, these methods consume too much limited area resources on the PCB board. As the PCB is developed toward smaller and more refined space, the application range of such methods is gradually narrowing. Moreover, protective wiring with grounding holes at both ends will cause resonance at some specific frequency points. In [8], the crosstalk can be suppressed by reducing the coupling length of microstrip lines. However, further reducing the coupling length between devices with a limited layout is difficult and limits the wiring efficiency. In [9], a step-shaped transmission line is created to reduce crosstalk. However, the effect of crosstalk suppression is limited, and the far-end crosstalk is only improved by 4 dB. In [10], the crosstalk between microstrip lines is reduced by etching an S-shaped defected microstrip structure (DMS) on the microstrip lines. Simulation and analysis show that the structure can suppress crosstalk in some frequency bands. However, this structure cannot achieve the suppression of crosstalk in the full frequency band. In addition, the S-shaped DMS is more complicated, and 11 lines need to be etched between the microstrip lines, which is technically difficult.

In this paper, a novel G-shaped DMS is designed based on the structure of References [9] and [10]. A method is proposed to reduce the crosstalk between coupled microstrip lines by combining G-shaped DMS with the step-shaped transmission line. The method can suppress crosstalk to a greater extent,

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and the structure is simpler and easier to implement. Modeling simulation and actual measurement are conducted through high-frequency structure simulator (HFSS) software and vector network analyzer (VNA) to verify the effectiveness of the proposed method. The simulation and actual measurement results show that the proposed method can effectively suppress the crosstalk between coupled microstrip lines. The method can also reduce the crosstalk by nearly 20 dB in the frequency range of 4–5 GHz.

2. THEORY ANALYSIS

If the distance between two microstrip lines on the PCB board is small, then noise will be generated on the adjacent microstrip lines when one of them transmits signals. This phenomenon is called crosstalk. Crosstalk is a phenomenon of energy transferring to adjacent microstrip lines, and the essential reason for formation is electromagnetic coupling [11].

2.1. Theoretical Analysis of Step-Shaped Transmission Line

Figure 1 shows the coupled microstrip lines signal transmission model. An excitation signal is added to port 1, and no excitation signal is added at the other ends. At this time, transmission line A is the aggressor line, and transmission line B is the victim line.

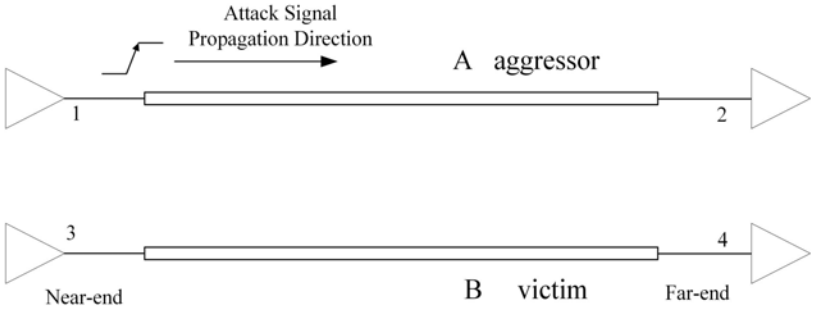

Figure 1. Coupling microstrip lines signal transmission model.

A far-end crosstalk signal will be obtained at port 4. The far-end crosstalk can be modeled as [12]

$$V_{\mathrm{FEXT}}(t) = \frac{1}{2}\left(\frac{C_m}{C_T} - \frac{L_m}{L_S}\right) \cdot TD \cdot \frac{\mathrm{d}V_a(t - TD)}{\mathrm{d}t} \tag{1}$$

where TD is the propagation time through the transmission line, $V_a(t)$ the applied voltage at the aggressor line, C_m the per unit length mutual capacitance, C_T the sum of self and mutual capacitance, L_m the per unit length mutual inductance, L_S the self inductance, and t the time.

Equation (1) shows that the far-end crosstalk is mainly determined by the difference in the ratio of capacitive coupling C_m/C_T and inductive coupling L_m/L_S. Therefore, step-shaped transmission lines can be used to reduce crosstalk [9]. As shown in Figure 2, the step-shaped transmission line divides two coupled microstrip lines into multiple sections, and a stepped structure with some width and length is created on each section. The change in the distance between the microstrip lines will change the electric field strength between the microstrip lines and then affect the size of the capacitive coupling. Therefore, the values of far-end crosstalk in different sections are also different. The closer part of the two microstrip lines will cause larger capacitive coupling, and the longer part will cause smaller capacitive coupling. By optimizing the length and width of the steps in each section, the capacitive coupling of the closer part of the microstrip lines is larger than the inductive coupling; the far-end crosstalk is a positive number; the longer part of the capacitive coupling is smaller than the inductive coupling; and the far-end crosstalk is a negative number. The whole microstrip lines can be compensated completely or partially by using the superposition of crosstalk and adding the far-end crosstalk of each section.

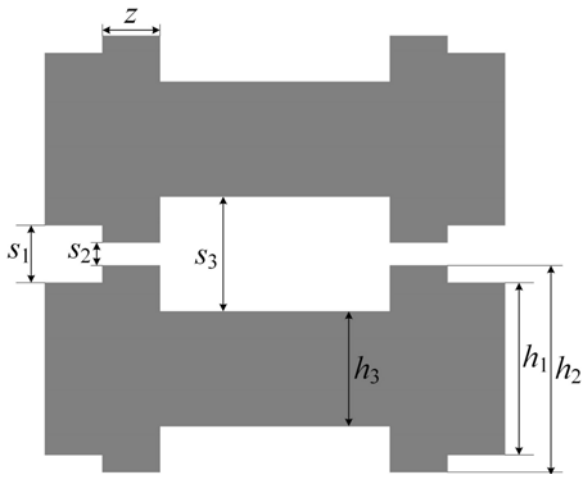

Figure 2. Geometric structure of step-shaped transmission line.

2.2. Theoretical Analysis of DMS

DMS is proposed based on a defective ground structure (DGS). The DMS can reduce the electromagnetic interference and the crosstalk noise between circuit and ground because it has a complete metal ground plate, so it is more suitable for microwave integrated circuits [13]. According to Figure 1, the reason for crosstalk generation is that electromagnetic coupling causes the aggressor line to generate interference currents on the victim line when current flows through the aggressor line, and the signal passing through the victim line changes. DMS produces defects in the microstrip lines by etching periodic or non-periodic structures of various shapes on the victim line. This kind of defect can show stopband characteristics when a current flows through the microstrip lines. Then, this condition prevents the energy radiated by the microstrip lines at a close distance, reduces the total capacitive coupling in the circuit, and suppresses crosstalk.

In summary, the DMS and step-shaped transmission line are combined. The step structure is used to change the ratio difference between the capacitive coupling C_m/C_T and inductive coupling L_m/L_S in different sections of the microstrip lines. The DMS is also used to reduce the total capacitive coupling in the circuit. The crosstalk suppression effect can be better achieved.

3. SUPPRESSION OF CROSSTALK BY COMBINED G-SHAPED DMS AND STEP-SHAPED TRANSMISSION LINE

The G-shaped DMS structure is combined with a step-shaped transmission line to achieve better suppression of crosstalk. The designed G-shaped DMS structure is shown in Figure 3. DGS can be equivalent to a simple parallel LC resonant circuit in the microwave system [14, 15], and DMS and DGS have the same stopband characteristics. Thus, the DMS can also be equivalent to a simple parallel LC resonant circuit.

Figure 4 shows the equivalent circuit of the coupled microstrip lines with etched G-shaped DMS. Each planar microstrip line in the plane microstrip lines structure can construct a π-type unit circuit with its inductance and capacitance. When the two microstrip lines are on the same plane, mutual inductance will occur between the microstrip lines. The two microstrip lines are equivalent to two substrates, which can be equivalent to the mutual capacitance between the microstrip lines. At the same time, the notch on the microstrip lines affects the current distribution of the microstrip lines, and the current change will affect the reference ground. Therefore, the effect of the DMS on the ground must be considered in the equivalent circuit. In Figure 4, L_1–L_4 represent the self-inductance of the microstrip lines; L_{m1} and L_{m2} denote the mutual inductance of the microstrip lines; C_1–C_6 represent the self-capacitance of the microstrip lines; C_7–C_9 denote the mutual capacitance of the microstrip lines; the LC parallel resonance circuit constructed by C_{11} and L_{11} is the equivalent circuit of the G-shaped

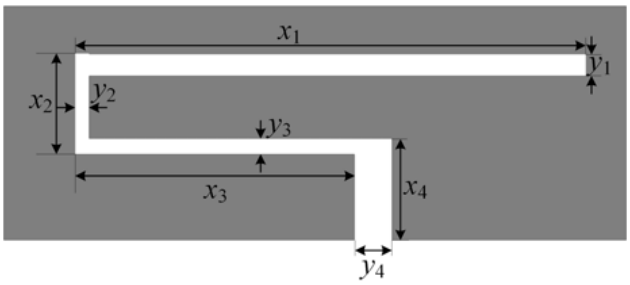

Figure 3. Geometric structure of G-shaped DMS.

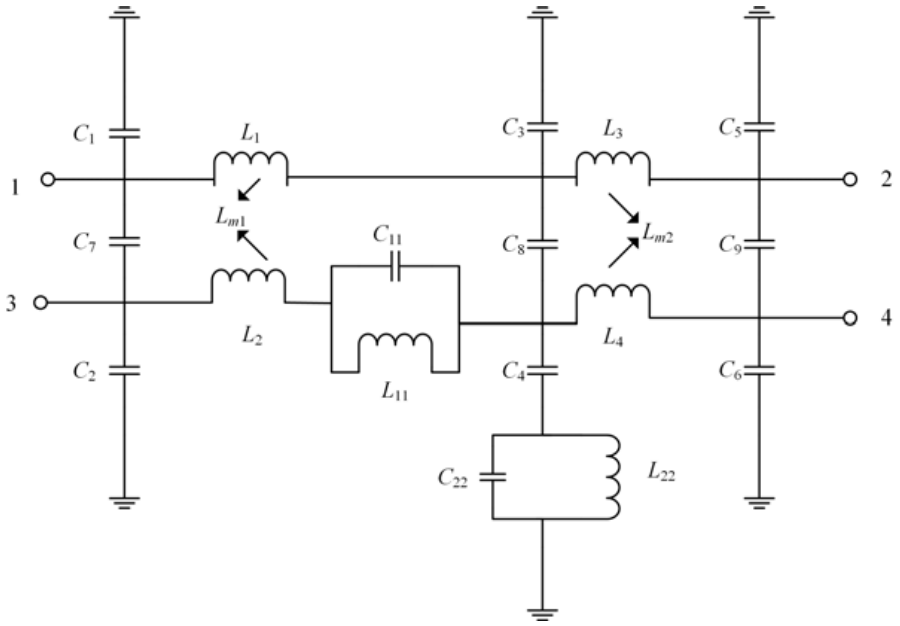

Figure 4. Equivalent circuit of the coupled microstrip lines with etched G-shaped DMS.

DMS, and the LC parallel resonance circuit constructed by C_{22} and L_{22} is the influence of the G-shaped DMS on the reference ground.

The expressions of equivalent circuit C_{11} and L_{11} of G-shaped DMS are [16]

$$C_{11} = \left(\frac{w_c}{Z_0 g_1}\right) \frac{1}{w_0^2 - w_c^2} \tag{2}$$

$$L_{11} = \frac{1}{4\Pi^2 f_0^2 C_{11}} \tag{3}$$

where w_0 is the resonant frequency of C_{11} and L_{11}; w_c is the cut-off frequency of C_{11} and L_{11}; Z_0 is the characteristic impedance of microstrip lines, which is $50\,\Omega$; g_1 is the normalized parameter of the 1st Butterworth low pass prototype, which is 2; f_0 is the frequency of stop band attenuation pole.

According to the impedance characteristics of the capacitor and the inductor, the higher-frequency signal easily reaches the output terminal through the capacitor, and the lower-frequency signal easily reaches the output terminal through the inductor. Therefore, when the input signal passes through the LC parallel resonant circuit constructed by C_{11} and L_{11}, the capacitor C_{11} plays a major role in suppressing the far-end crosstalk between ports 1 and 4. The capacitive coupling between ports 1 and 4 when the microstrip lines are not etched with the G-shaped DMS structure is shown in Figure 5(a). At this time, the total capacitive coupling C_{m1} is Equation (4), which is the sum of three parallel coupling capacitors. The capacitive coupling between ports 1 and 4 when the microstrip lines are etched with

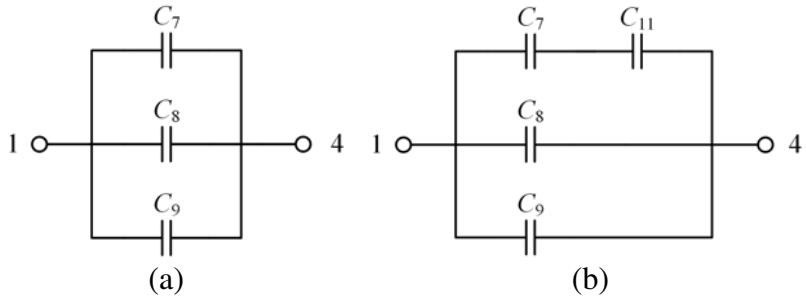

Figure 5. Total capacitive coupling from port 1 to port 4. (a) Without G-shaped DMS. (b) With G-shaped DMS.

the G-shaped DMS structure is shown in Figure 5(b). Coupling capacitors C_7 and C_{11} are connected in series, and the total capacitive coupling C_{m2} is Equation (5). Obviously, C_{m1} is smaller than C_{m2}. Therefore, the microstrip lines etching G-shaped DMS can reduce the far-end crosstalk between ports 1 and 4.

$$C_{m1} = C_7 + C_8 + C_9 \tag{4}$$

$$C_{m2} = \frac{C_7 \cdot C_{11}}{C_7 + C_{11}} + C_8 + C_9 \tag{5}$$

The designed DMS and step-shaped transmission line are combined to suppress crosstalk without changing the relevant parameters of the microstrip lines and meeting the impedance matching of the port. A step-shaped structure is realized by replacing each of the coupled transmission lines of multiple cascaded sections with some length on a pair of adjacent microstrip lines, and the G-shaped DMS is etched on the victim line. The impedance characteristics of the first and last sections are restricted to conform to the impedance matching, which is 50 ohm, while the impedance characteristic of the middle sections is optimized. The capacitive coupling after the microstrip lines is divided into different sections is changed. Accordingly, the ratio of the capacitive coupling C_m/C_T and the inductive coupling L_m/L_S of each section is changed. Moreover, the designed G-shaped DMS has stopband characteristics, which can be equivalent to a parallel LC circuit and can reduce the total capacitive coupling in the parallel microstrip lines.

According to the above analysis, the coupled microstrip lines structure combining the designed G-shaped DMS and the step-shaped transmission line is shown in Figure 6. Figure 6(a) is a top view of the coupled microstrip lines, and Figure 6(b) is a side view of the coupled microstrip lines. The structure divides the microstrip lines into four equal parts. The same step-shaped structure is created for each part of the microstrip lines, and the G-shaped DMS of same size is etched.

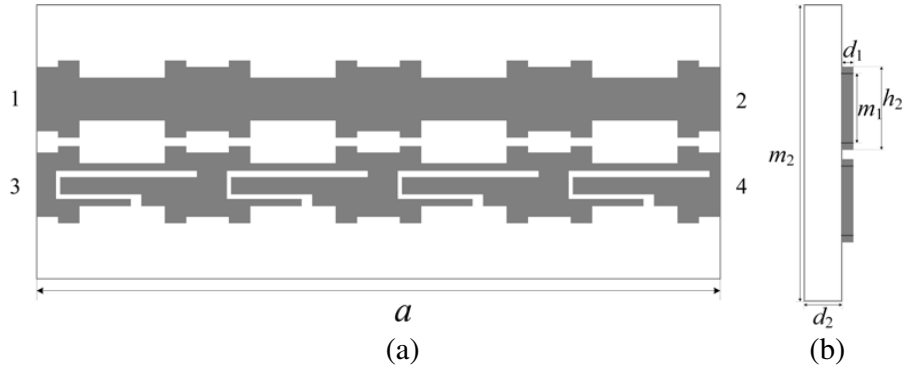

Figure 6. Coupled microstrip lines combining G-shaped DMS and step-shaped transmission line. (a) Top view. (b) Side view.

4. SIMULATION RESULTS AND ANALYSIS

The HFSS simulation software is used to model and simulate the structure shown in Figure 6 to verify the effect of combining the designed G-shaped DMS with the step-shaped transmission line to reduce the crosstalk between the coupled microstrip lines. The parameters of the G-shaped DMS and step-shaped structure are optimized to meet the requirement that the coupled microstrip lines combined with the G-shaped DMS and step-shaped transmission line have a reasonable insertion loss and generate the smallest far-end crosstalk. The specific dimensions of the final coupled microstrip lines after adjustment and optimization are given as follows: microstrip lines length is $a = 32\,\text{mm}$; microstrip lines width is $m_1 = 3\,\text{mm}$; microstrip lines thickness is $d_1 = 0.08\,\text{mm}$; dielectric substrate length is $a = 32\,\text{mm}$; dielectric substrate width is $m_2 = 12.8\,\text{mm}$; dielectric substrate thickness is $d_2 = 1.6\,\text{mm}$; dielectric substrate is FR4, dielectric constant $\varepsilon_r = 4.4$. The conductor is copper, and the characteristic impedance of microstrip lines is $50\,\Omega$. The dimensions of the step-shaped structure in each part are given as follows: $h_1 = 3\,\text{mm}$, $h_2 = 3.6\,\text{mm}$, $h_3 = 2\,\text{mm}$, $z = 1\,\text{mm}$, $s_1 = 1\,\text{mm}$, $s_2 = 0.4\,\text{mm}$, $s_3 = 2\,\text{mm}$. The dimensions of the G-shaped transmission line in each part are given as follows: $x_1 = 6.6\,\text{mm}$, $y_1 = 0.3\,\text{mm}$, $x_2 = 1.32\,\text{mm}$, $y_2 = 0.2\,\text{mm}$, $x_3 = 4\,\text{mm}$, $y_3 = 0.22\,\text{mm}$, $x_4 = 1.2\,\text{mm}$, $y_4 = 0.5\,\text{mm}$. The radiation conditions are set by HFSS, and the port excitation is set at the port for simulation.

The HFSS simulation results are shown in Figures 7 and 8. Figure 7 shows that the coupled microstrip lines combining the designed G-shaped DMS and the step-shaped transmission line can effectively reduce crosstalk in the range of 1–8 GHz, and the far-end crosstalk is reduced by around 20 dB within 4–5 GHz.

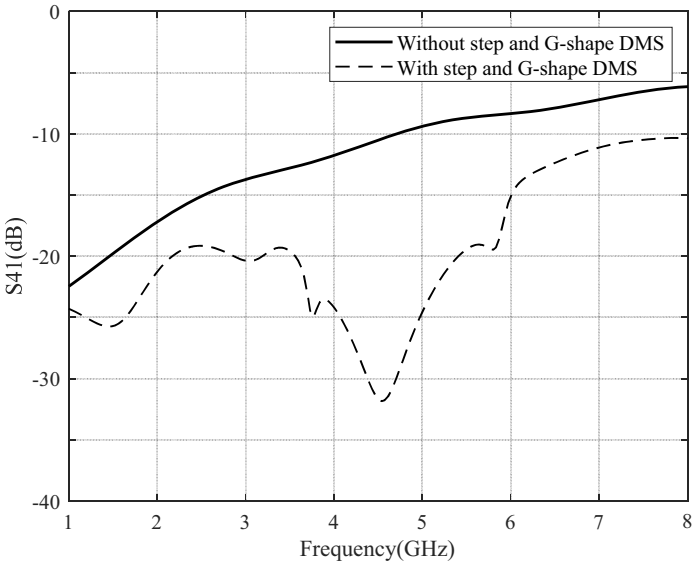

Figure 7. Comparison of the simulation results of microstrip lines S_{41} before and after the combination of G-shaped DMS and step-shaped transmission line.

Figure 8 shows the simulation results of microstrip lines insertion loss S_{21} before and after the combination of G-shaped DMS and step-shaped transmission line. The dotted line in Figure 8 is the simulation result of the microstrip lines S_{21} combined with the designed G-shaped DMS and step-shaped transmission line. As shown in Figure 8, the insertion loss S_{21} of the microstrip lines combined with G-shaped DMS and step-shaped structure has the same overall trend and smaller errors than the insertion loss S_{21} of the microstrip lines without DMS and the step structure.

A PCB is produced according to the HFSS simulation model and related parameters, as shown in Figures 9 and 10, to further verify the accuracy of the simulation results. Figure 9 shows the actual production diagram of measuring far-end crosstalk S_{41}. Figure 10 shows the actual production diagram of measuring insertion loss S_{21}. The VNA is used to test the S parameters of the PCB.

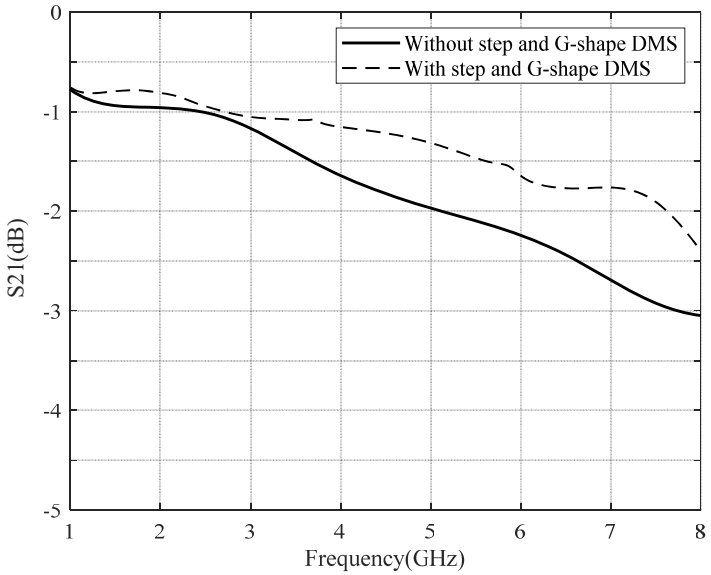

Figure 8. Comparison of the simulation results of microstrip lines S_{21} before and after the combination of G-shaped DMS and step-shaped transmission line.

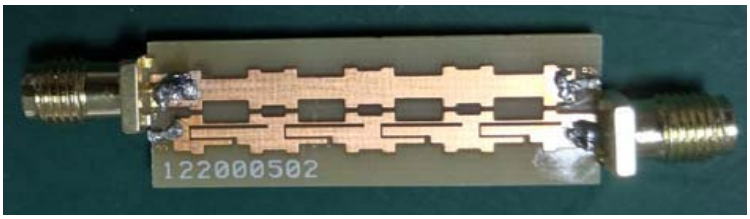

Figure 9. Actual production diagram of measuring far-end crosstalk S_{41}.

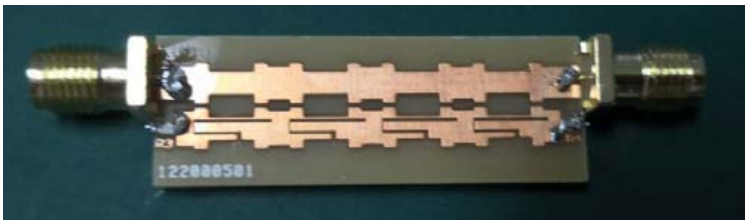

Figure 10. Actual production diagram of measuring insertion loss S_{21}.

The results of far-end crosstalk simulated by HFSS software are compared with the actual measurement results of a VNA, as shown in Figure 11. The pointed line in Figure 11 is the actual measurement result of the far-end crosstalk of the microstrip lines combined with the designed G-shaped DMS and step-shaped transmission line. The figure shows that the actual measured far-end crosstalk is consistent with the overall trend of the simulated far-end crosstalk results. The microstrip lines combined with the G-shaped DMS and the step-shaped transmission line can reduce the far-end crosstalk. This method can also reduce the far-end crosstalk by more than 20 dB in the range of 4–5 GHz.

At the same time, the results of insertion loss simulated by HFSS software are compared with the actual measurement results of a VNA, as shown in Figure 12. The pointed line in Figure 12 is the actual measurement result of the insertion loss of the microstrip lines combined with the designed G-shaped

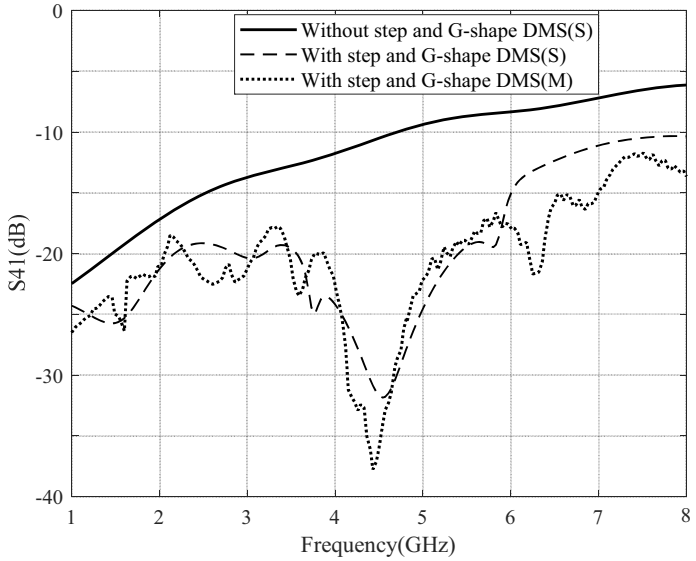

Figure 11. Comparison of the simulation and measured results of microstrip lines S_{41} before and after the combination of G-shaped DMS and step-shaped transmission line.

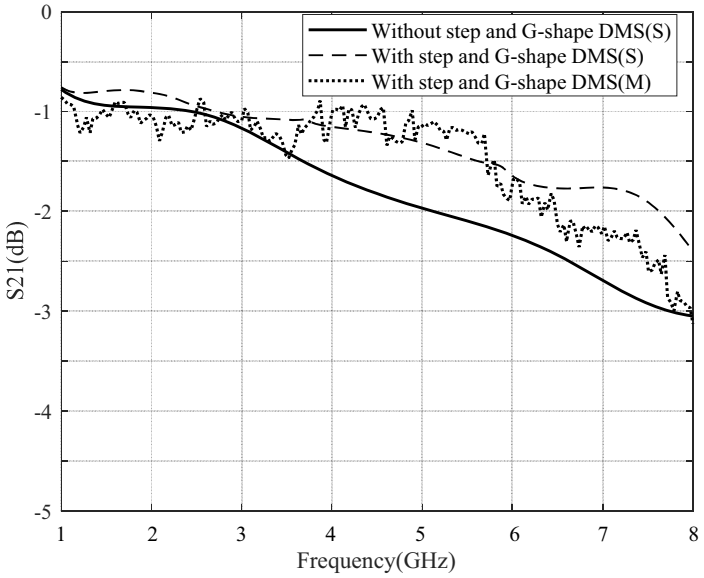

Figure 12. Comparison of the simulation and measured results of microstrip lines S_{21} before and after the combination of G-shaped DMS and step-shaped transmission line.

DMS and step-shaped transmission line. Figure 12 shows that the actual measured insertion loss is consistent with the numerical trend of the simulation results. However, a certain deviation exists in the amplitude in the entire frequency band. The deviation in measurement and simulation data may be caused by the error of PCB fabrication precision and the instability of the test environment.

In the next step, the crosstalk of microstrip lines can be suppressed in a wider frequency range by improving the G-shaped DMS and step-shaped structure.

5. CONCLUSION

In this paper, the crosstalk between the microstrip lines is reduced by creating a step-shaped transmission line and etching a designed G-shaped DMS on one of the microstrip lines. HFSS software simulation and VNA for actual measurement jointly verify that the proposed method can effectively suppress the far-end crosstalk between the coupled microstrip lines. The method can also reduce crosstalk by more than 20 dB in the frequency range of 4–5 GHz. In addition, the designed new structure does not further occupy PCBs space, and the structure is simple and easy to implement.

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