Project2: Microwave device testing and communication system construction

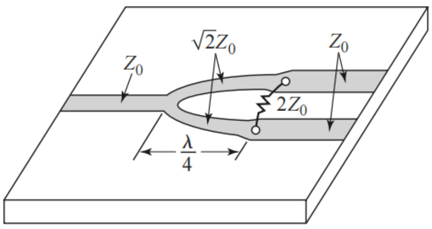
Author	Name	Student ID
	Name: 赵宣懿	Student ID: 12111101
	Name: 张皓东	Student ID: 12113010
	Name: 宋宜航	Student ID: 12112725
	Name: 王振宇	Student ID: 12112719

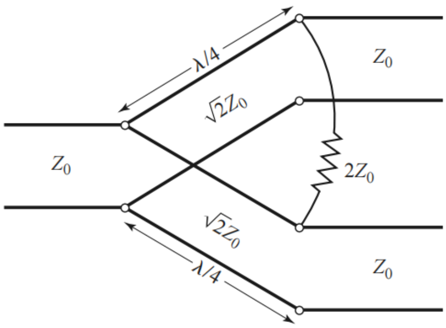
Theorical Analysis

1. Wilkinson Power Divider

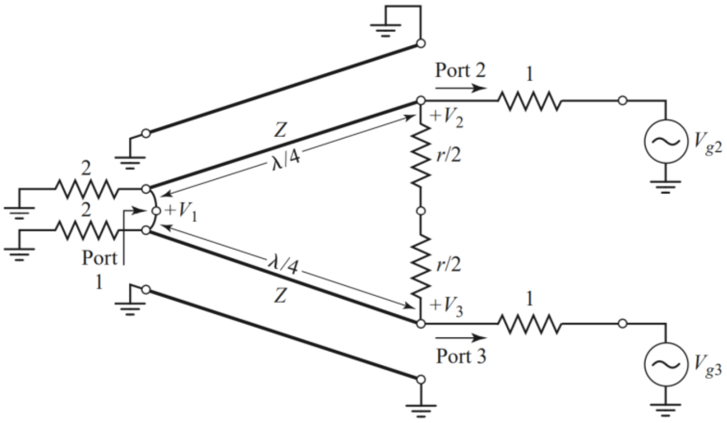
Wilkinson power divider is a lossy three-port network that can be made having all ports matched with isolation between output ports. The lossless T-junction divider suffers from the disadvantage of not being matched at all ports, and it does not have isolation between output ports. The resistive divider can be matched at all ports, but even though it is not lossless, isolation is still not achieved. By contrast, the Wilkinson power divider use a resistor achieve that all ports matched and isolation between outputs ports, which can be made with arbitrary power division.

In project 2, we mainly talk about the 3dB Wilkinson power divider, whose models are shown below both in microstrip line form and corresponding transmission line circuit, respectively.





To analyze this, we apply even-odd mode, in which we reduce the circuit to two simpler ones driven by symmetric and antisymmetric sources at output ports. For simplicity, we can normalize all impedances to the characteristic impedance Z_0, and the transmission line circuit with voltage generators at the output ports is shown below:

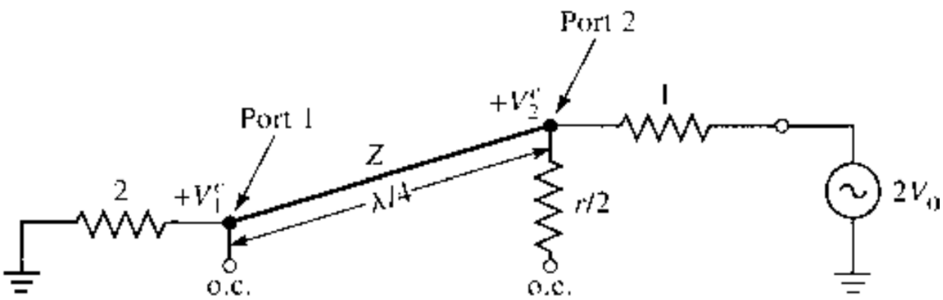


Now define two separate modes of excitation for the circuit the even mode, where $V_{g2} = V_{g3} = 2V_0$, and the odd mode, where $V_{g2} = -V_{g3} = 2V_0$. Superposition of these two modes effectively produces an excitation of $V_{g2} = 4V_0$ and $V_{g3} = 0$, from which we find the scattering parameters of the network. We now treat these two modes separately.

- **Even mode**

  For even-mode excitation, $V_{g2} = V_{g3} = 2V_0$, so $V_2^e = V_3^e$, and therefore no current flows through the $r/2$ resistors or the short circuit between the inputs of the two transmission lines at port 1.

  We can then bisect the network above with open circuits at these points to obtain the network below:



  Looking into port 2, we can see an impedance:

$$Z_{in}^e = \frac{Z^2}{2}$$

  Thus, if $Z = \sqrt{2}$, port 2 will be matched for even-mode excitation, then $V_2^e = V_0$ since $Z_{in}^e = 1$. The $r/2$ resistor is superfluous in this case since one end is open-circuited. Then, we find V_1^e from the transmission line equations. If we let $x = 0$ at port 1 and $x = -\lambda/4$ at port 2, we can write the voltage on the transmission line section as:

$$V(x) = V^+(e^{-j\beta x} + \Gamma e^{j\beta x})$$

Then,

$$V_2^e = V\left(-\frac{\lambda}{4}\right) = jV^+(1 - \Gamma) = V_0$$

$$V_1^e = V(0) = V^+(1 + \Gamma) = jV_0\frac{\Gamma + 1}{\Gamma - 1}$$

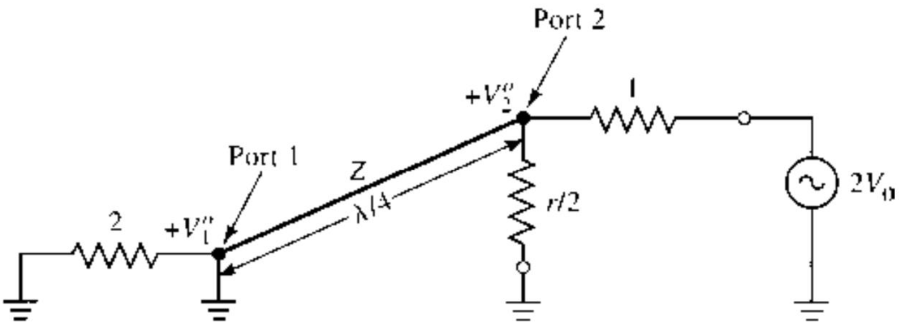
The reflection coefficient Γ is that seen at port 1 looking toward the resistor of normalized value 2, so

$$\Gamma = \frac{2 - \sqrt{2}}{2 + \sqrt{2}}$$

$$V_1^e = -jV_0\sqrt{2}$$

- **Odd mode**

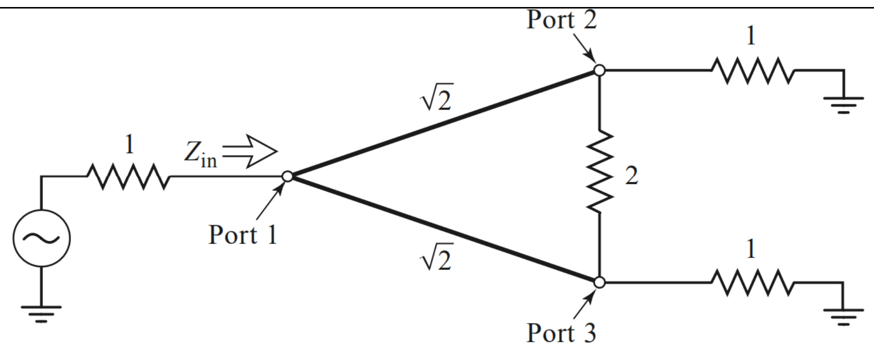
For odd-mode excitation, $V_{g2} = -V_{g3} = 2V_0$, so $V_2^o = -V_3^o$, and there is a voltage null along the middle of the whole transmission line circuit. We can then bisect this circuit by grounding it at two points on its midplane to give the network below:



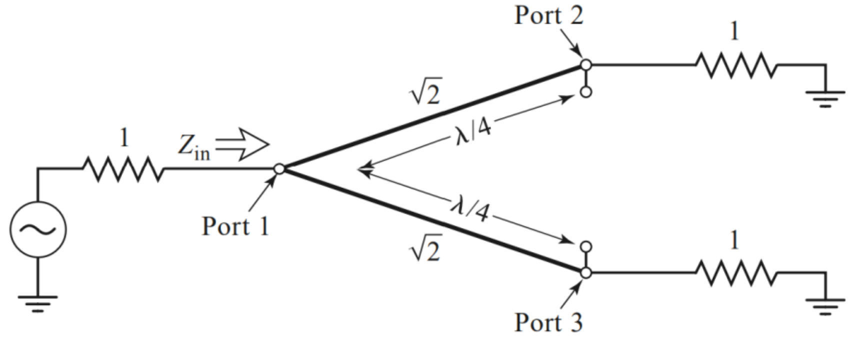
Looking into port 2, we see an impedance of $r/2$ since the parallel-connected transmission line is $\lambda/4$ long and shorted at port 1, and so looks like an open circuit at port 2. Thus, port 2 will be matched for odd-mode excitation if $r = 2$. Then $V_2^o = V_0$ and $V_1^o = 0$, for this mode of excitation all power is delivered to the r/2 resistors, with none going to port 1.

Finally, we must find the input impedance at port 1 of the Wilkinson divider when ports 2 and 3 are terminated in matched loads.

The resulting circuit is shown below,



where it is seen that this is like an even mode of excitation since $V_2 = V_3$. No current flows through the resistor of normalized value 2, so it can be removed, leaving the circuit:



We then have the parallel connection of two quarter-wave transformers terminated in loads of unity (normalized). The input impedance is:

$$Z_{in} = \frac{1}{2}(\sqrt{2})^2 = 1$$

In summary, we can establish scattering parameters for the 3dB Wilkinson power divider,

- Z_{in} = 1 at port 1, so

$$S_{11} = 0$$

- ports 2 and 3 matched for even and odd modes, so

$$S_{22} = S_{33} = 0$$

- symmetry due to reciprocity, so

$$S_{12} = S_{21} = \frac{V_1^e + V_1^o}{V_2^e + V_2^o} = \frac{-j}{\sqrt{2}}$$

- symmetry of ports 2 and 3, so

$$S_{13} = S_{31} = \frac{-j}{\sqrt{2}}$$

- due to short or open at bisection, so

$$S_{23} = S_{32} = 0$$

Then, we can construct the scattering matrix for the Wilkinson power,

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

From the scattering matrix, we find that when the divider is driven at port 1 and the outputs are matched, no power is dissipated in the resistor. Therefore, the divider is lossless when the outputs are matched, only reflected power from ports 2 or 3 is dissipated in the resistor. Also, $S_{23} = S_{32} = 0$, so ports 2 and 3 are isolated. And as $S_{11} = S_{22} = S_{33} = 0$, the Wilkinson power divider is perfectly matched.

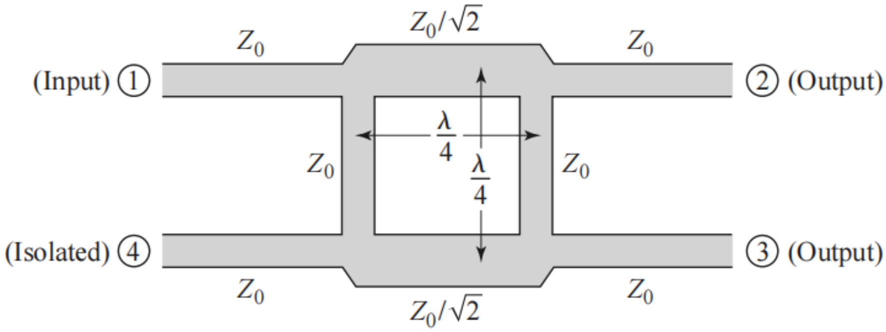
2. Branch Line Coupler

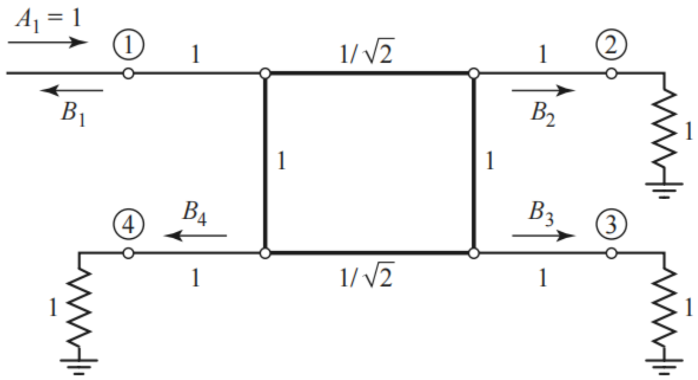
The branch line coupler is 3 dB directional couplers with a 90° phase difference in the outputs of the through and coupled arms. The basic operation of the branch-line coupler is as follows. With all ports matched, power entering port 1 is evenly divided between ports 2 and 3, with a 90° phase shift between these outputs. No power is coupled to port 4 (the isolated port). The scattering matrix has the following form:

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

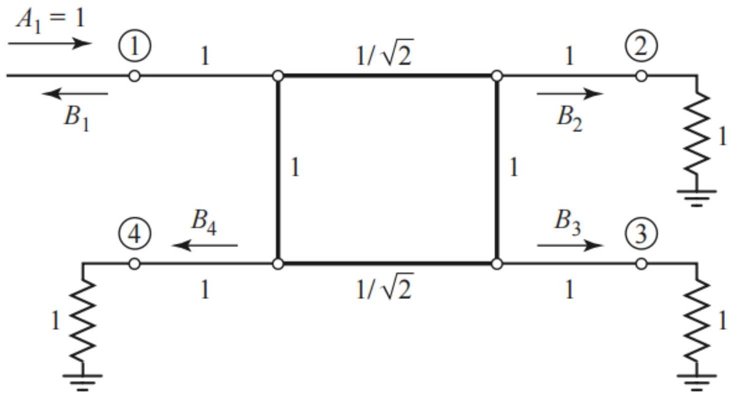
As the branch line hybrid has a high degree of symmetry, any port can be used as input port. The output ports will always be on the opposite side of the junction from the input port, and the isolated port will be the remaining port on the same side as the input port. This symmetry is reflected in the scattering matrix, as each row can be obtained as a transposition of the first row.

The geometry of a branch line coupler and Circuit of the branch-line hybrid coupler in normalized form are shown below, separately.

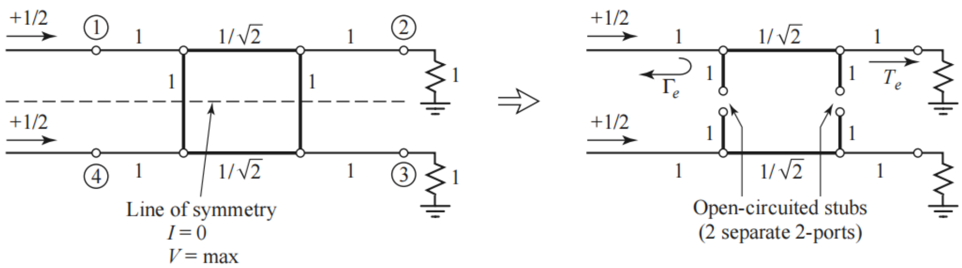


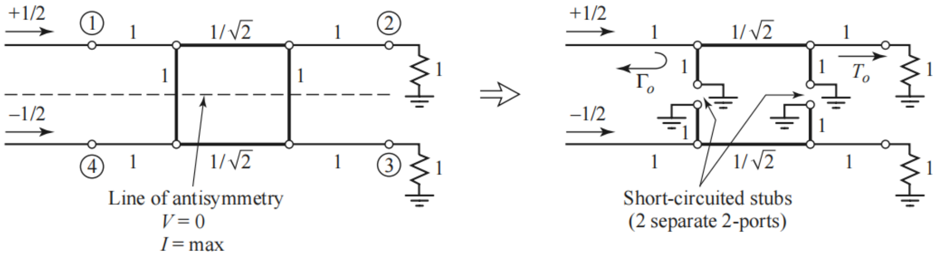


In the schematic circuit of the branch line coupler in normalized form, the common ground return for each transmission line is not shown. We assume that a wave of unit amplitude $A_1 = 1$ is incident at port 1.



The circuit can be decomposed into the superposition of an even-mode excitation and an odd-mode excitation as shown below, separately. And since the circuit is linear, the actual response (the scattered waves) can be obtained from the sum of the responses to the even and odd excitations.





Because of the symmetry or anti-symmetry of excitation, the four-port network can be decomposed into a set of two decoupled two-port networks. Because the amplitudes of incident waves for these two-ports are $\pm {}^{1}/_{2}$, the amplitudes of the emerging wave at each port of the branch-line hybrid can be expressed as:

$$B_1 = \frac{1}{2}\Gamma_e + \frac{1}{2}\Gamma_o,$$

$$B_2 = \frac{1}{2}T_e + \frac{1}{2}T_o,$$

$$B_3 = \frac{1}{2}T_e - \frac{1}{2}T_o,$$

$$B_4 = \frac{1}{2}\Gamma_e - \frac{1}{2}\Gamma_o,$$

where $\Gamma_{e,o}$ and $T_{e,o}$ are the even- and odd-mode reflection and transmission coefficients for the two-port networks. First, consider the calculation of $\Gamma_{e,o}$ and $T_{e,o}$ for the even-mode two-port circuit, which can best be done by multiplying the ABCD matrix of each cascade component in that circuit.

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_e = \underbrace{\begin{bmatrix} 1 & 0 \\ j & 1 \end{bmatrix}}_{\substack{\text{Shunt} \\ Y = j}} \underbrace{\begin{bmatrix} 0 & j/\sqrt{2} \\ j\sqrt{2} & 0 \end{bmatrix}}_{\substack{\lambda/4 \\ \text{Transmission} \\ \text{line}}} \underbrace{\begin{bmatrix} 1 & 0 \\ j & 1 \end{bmatrix}}_{\substack{\text{Shunt} \\ Y = J}} = \frac{1}{\sqrt{2}} \begin{bmatrix} -1 & j \\ j & -1 \end{bmatrix}$$

Then convert the ABCD matrix into S-parameter matrix, which are equivalent to the reflection and transmission coefficients:

$$\Gamma_e = \frac{A + B - C - D}{A + B + C + D} = \frac{(-1 + j - j + 1)/\sqrt{2}}{(-1 + j + j - 1)/\sqrt{2}} = 0,$$

$$T_e = \frac{2}{A + B + C + D} = \frac{2}{(-1 + j + j - 1)/\sqrt{2}} = \frac{-1}{\sqrt{2}}(1 + j).$$

Similarly, for the odd mode we obtain:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_o = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & j \\ j & 1 \end{bmatrix}$$

which gives the reflection and transmission coefficients as:

$$\Gamma_o = 0,$$
$$T_o = \frac{1}{\sqrt{2}}(1 - j)$$

Using the equation above we can get that:

$B_1 = 0$ (port 1 is matched)

$B_2 = -\frac{j}{\sqrt{2}}$ (port 1 is matched)

$B_3 = -\frac{1}{\sqrt{2}}$ (port 1 is matched)

$B_4 = 0$ (no power to port 4)

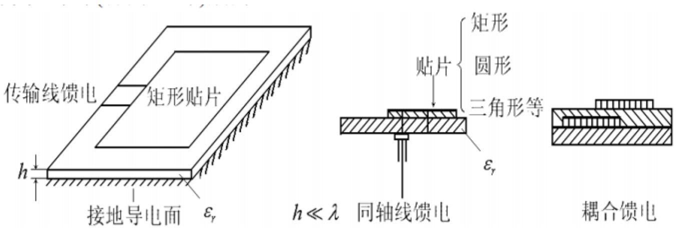
These results agree with the first row and column of the scattering matrix. the remaining elements can easily be found by transposition.

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

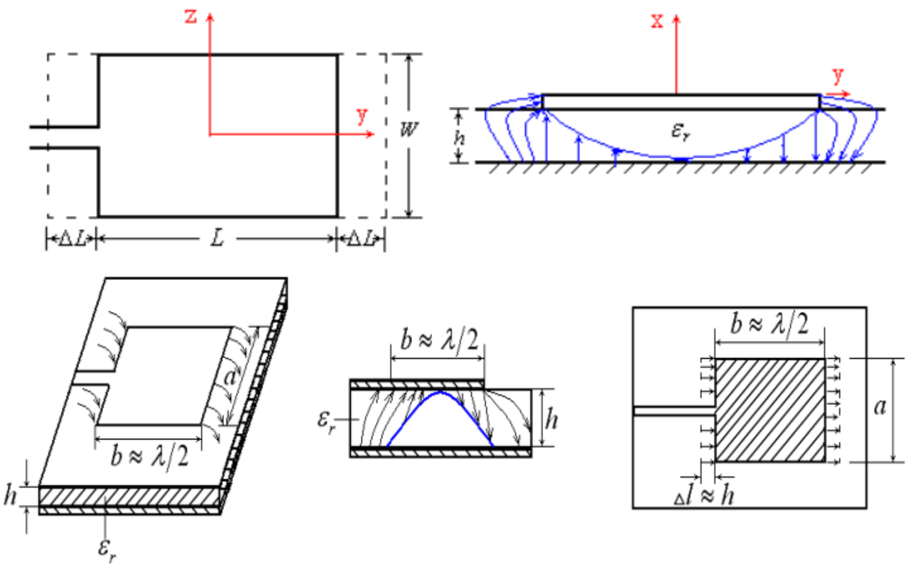
In practice, due to the quarter-wavelength requirement, the bandwidth of a branch-line hybrid is limited to $10\%–20\%$. However, as with multi sections matching transformers and multi-hole directional couplers, the bandwidth of a branch-line hybrid can be increased to a decade or more by using multiple sections in cascade. In addition, the basic design can be modified for unequal power division and/or different characteristic impedances at the output ports. Another practical point to be aware of is the fact that discontinuity effects at the junctions of the branch-line coupler may require that the shunt arms be lengthened by $10^{\circ} - 20^{\circ}$.

3. Patch Antenna

The microstrip patch antenna contains a metal patch, which is far thinner than the wavelength ($h \ll \lambda_0$, λ_0 is the wavelength in free space), as shown below. Microstrip patches are designed that there is maximum radiation in the side emission direction of the patch, which is acceptable through the selection of different patch shape excitation way to achieve. Selecting different shape of patch group can also achieve end shooting radiation. For rectangular patch, patch length L generally satisfies $\lambda_0/3 < L < \lambda_0/2$. Microstrip patch and ground are separated by a thin sheet of medium called substrate.



The radiation mechanism of microstrip antenna is actually high-frequency electromagnetic leakage. If a microwave circuit is not completely closed by the conductor, the discontinuities in circuit will cause electromagnetic radiation. For example, at the open end of microstrip circuit, electromagnetic radiation (leakage) will also be generated at discontinuities such as abrupt change in structure size and bending. When frequency is low, the electrical size of these parts is small, so the electromagnetic leakage is small. But as frequency increases, the electrical size increases, and the leakage also increases. After a special design, that is, the size of enlarged sheet, and make it work in a resonant state. Radiation is obviously enhanced, radiation efficiency is greatly improved, and become an effective antenna.



(a) (b) (c)

The electric field in the medium substrate between the radiating patch and the conducting ground plate does not change along the direction a of the patch width and the direction h of the thickness. It only changes along the direction of length b, whose structure is shown in Figure (a) above. The radiation field can be considered to be generated by the edge field on the two open ends of the patch along the length direction, as shown in Figure (b)(c) above. The edge field is decomposed into horizontal and vertical components. Because the patch length $b \approx \lambda/2$, the vertical electric field components at the two open ends are reversed, and the fields generated by these components in space cancel each other out (or are very weak). And the horizontal component of the electric field is in phase. The far-region radiation field is mainly generated by the component field. The maximum radiation direction is perpendicular to the patch, which can be seen from the analysis that the rectangular microstrip antenna can be equivalent to two slot antennas which are $\lambda/2$ apart and in the same phase excitation. This is the explanation of the transmission line model analysis method of microstrip antenna. And in project 2, we used the way of microstrip line feed to stimulate. (Coaxial probe feeding is adopted in class).

The theoretical calculations of patch antenna parameters are shown below.

- **Radiating Patch**

The material of dielectric substrate we use is FR4 epoxy with a thickness of $1.6\,mm$ and relative dielectric constant $\varepsilon_r = 4.4$, while the center frequency f_c equals $2.4GHz$. According to the equation below:

$$W_p = \frac{c}{2f_c}\sqrt{\frac{2}{\varepsilon_r + 1}} = \frac{3\times 10^8\, m/s}{2\times 2.4GHz}\sqrt{\frac{2}{4.4+1}} = 38.04mm$$

we can get the width of radiating patch $W_p = 38.04\text{mm}$

And due to

$$\frac{W_p}{h} = \frac{38.04mm}{1.6mm} = 23.775 > 1$$

we can calculate effective dielectric constant ε_e by the following equation:

$$\varepsilon_e = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[1 + 12\frac{h}{W_p}\right]^{-\frac{1}{2}}$$

$$= \frac{4.4+1}{2} + \frac{4.4-1}{2}\left[1 + 12\frac{1.6mm}{38.04mm}\right]^{-\frac{1}{2}} = 4.086$$

The length of microstrip antenna is generally chosen as ${\lambda_e}/{2}$, where λ_e is the wavelength of guided wave within the substrate, which is:

$$\lambda_e = \frac{c}{f\sqrt{\varepsilon_e}}$$

However, take the edge shortening sound deadening in count, we need to reconsider the length of patch L_p:

$$\Delta L_p = 0.412\text{h}\frac{(\varepsilon_e + 0.3)\left(\frac{W}{h} + 0.264\right)}{(\varepsilon_e - 0.258)\left(\frac{W}{h} + 0.8\right)}$$

$$L_p = \frac{c}{2f\sqrt{\varepsilon_e}} - 2\Delta L_p$$

$$= \frac{3 \times 10^8\, m/s}{2 \times 2.4GHz \times \sqrt{4.086}}$$

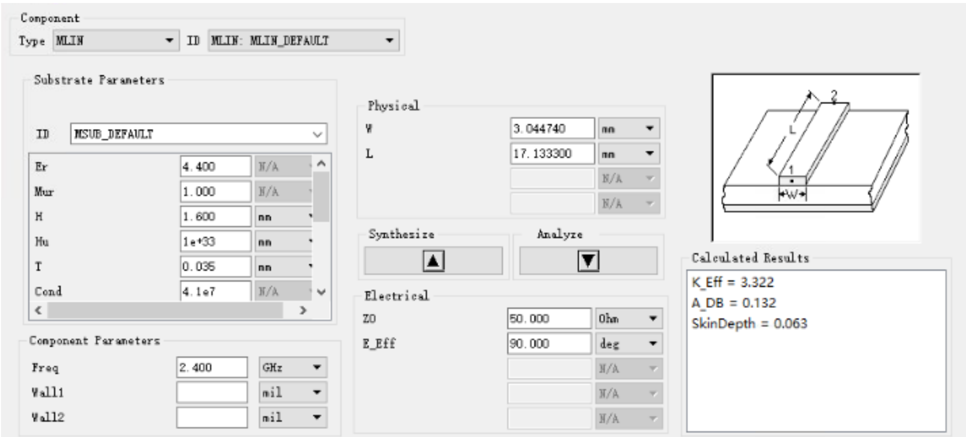
$$-2 \times 0.412 \times 1.6mm \frac{(4.086 + 0.3)\left(\frac{38.04mm}{1.6mm} + 0.264\right)}{(4.086 - 0.258)\left(\frac{38.04mm}{1.6mm} + 0.8\right)}$$

$$= 30.18\text{mm}$$

We get the length of patch $L_p = 30.18mm$

- **Feed**

We use a 50Ω microstrip transmission line and we calculate the length and width of the feed line through *LineCalc* in ADS:

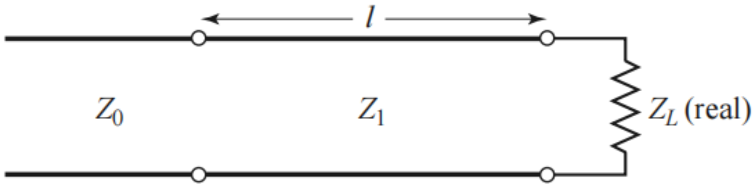


We obtain the width of the feed line $W_f = 3.045\ mm$ and the length $L_f = 17.133mm$. The length of transmission line depends on its electrical length, and there is no mandatory limit since it is matched.

- **Impedance Transformer**

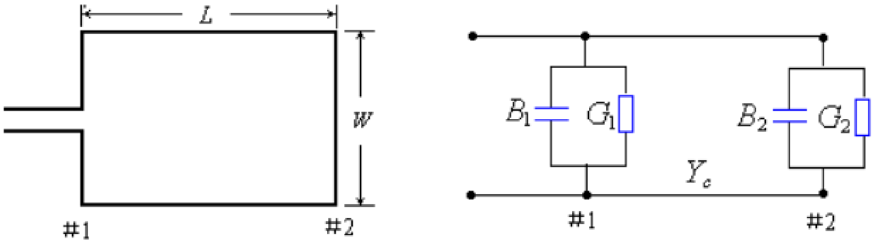
The single-section quarter-wave matching transformer circuit is shown below, with the characteristic impedance of the matching section given as:

$$Z_1 = \sqrt{Z_0 Z_L}$$



So first we calculate the input impedance magnitude of the microstrip patch.

The microstrip antenna can be converted to the following equivalent circuit.



The equivalent conductance on one side of the patch can be written as Y_1:

$$Y_1 = G_1 + jB_1$$

$$G_1 = \frac{W}{120\lambda_0}\left[1 - \frac{1}{24}(\beta h)^2\right], \qquad \frac{h}{\lambda_0} < \frac{1}{10}$$

$$B_1 = \frac{W}{120\lambda_0}[1 - 0.636\ln\beta h], \qquad \frac{h}{\lambda_0} < \frac{1}{10}$$

Due to symmetry, so the $B_1 = B_2$

So, the conductance of the microstrip patch is $Y = 2B_1$

Simplifying the above equation to:

$$G_1 = \begin{cases} \frac{1}{90}\left(\frac{W}{\lambda_0}\right)^2, & W \ll \lambda_0 \\ \frac{1}{120}\left(\frac{W}{\lambda_0}\right), & W \gg \lambda_0 \end{cases}$$

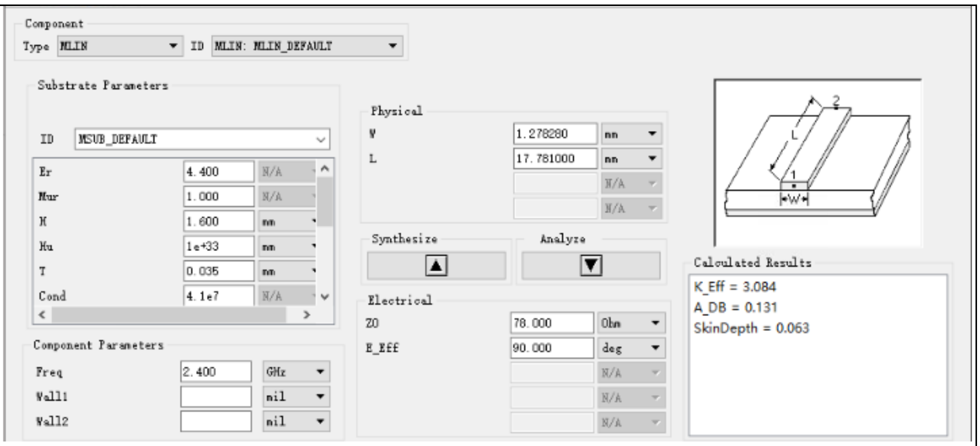
Here we have $W < \lambda$, and we can calculate:

$$R_{in} = \frac{1}{Y_{in}} = 120.6\ ohm$$

So, the characteristic impedance Z_t of quadrature wavelength converter

$$Z_t = \sqrt{Z_{t0}R_{in}} = \sqrt{50\ ohm \times 120.6\ ohm} = 78\ ohm$$

Then the length and width of the feed line can be calculated in ADS:



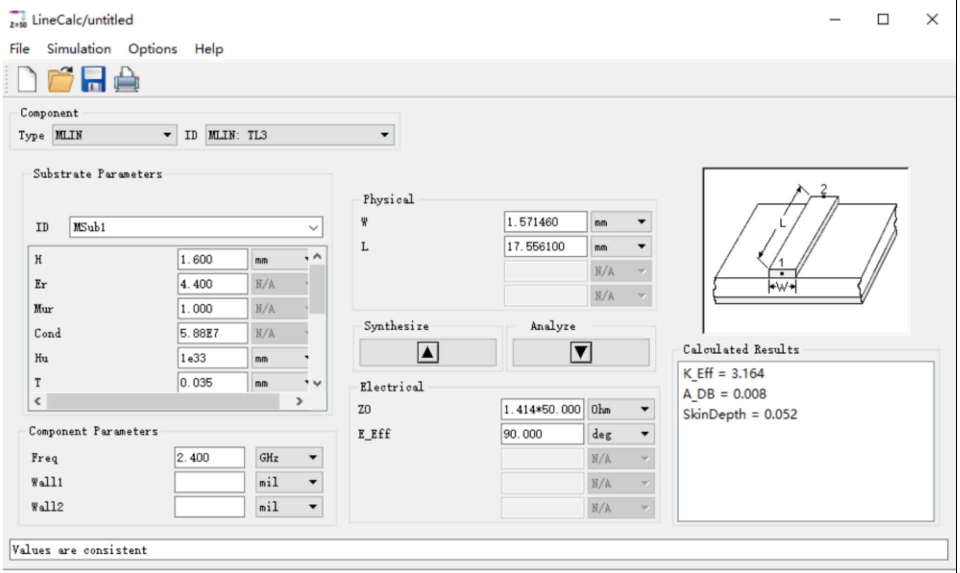
After *LineCalc*, we obtain the length and width of transformer $L_t = 17.78\,mm$, $W_t = 1.28\,mm$.

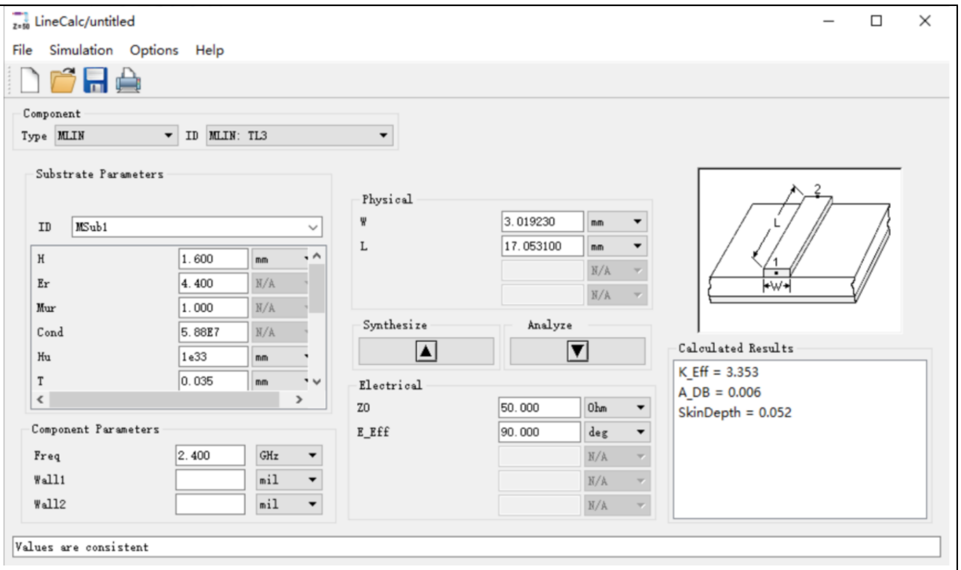
HFSS & ADS Simulation

1. 2.4GHz Wilkinson Power Divider

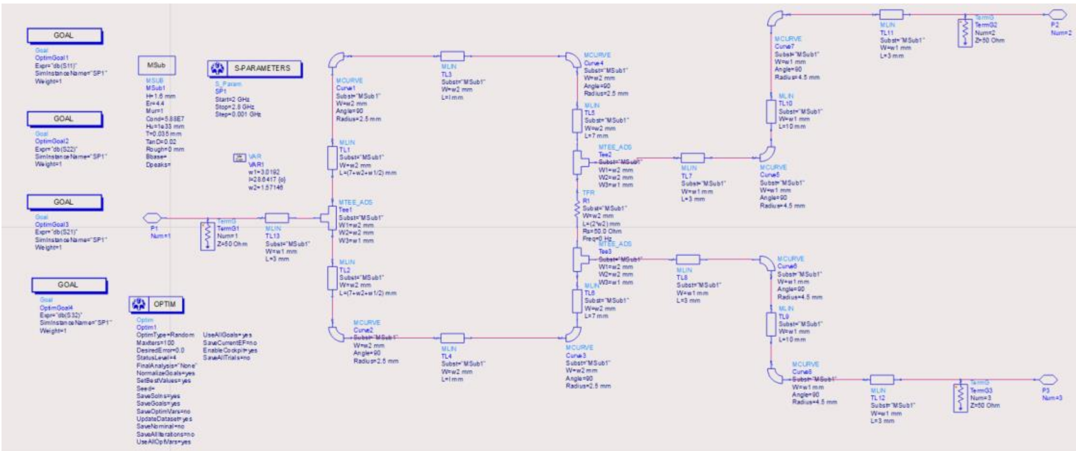
(1) ADS Modeling

Apply LineCalc tools to calculate the width of transmission line again. Below results are widths of $Z0 = 50\ \Omega$ and $Z0 = 50\sqrt{2}\ \Omega$ separately. So, we change the $w1 = 1.57146mm$, and $w2 = 3.01923$mm.

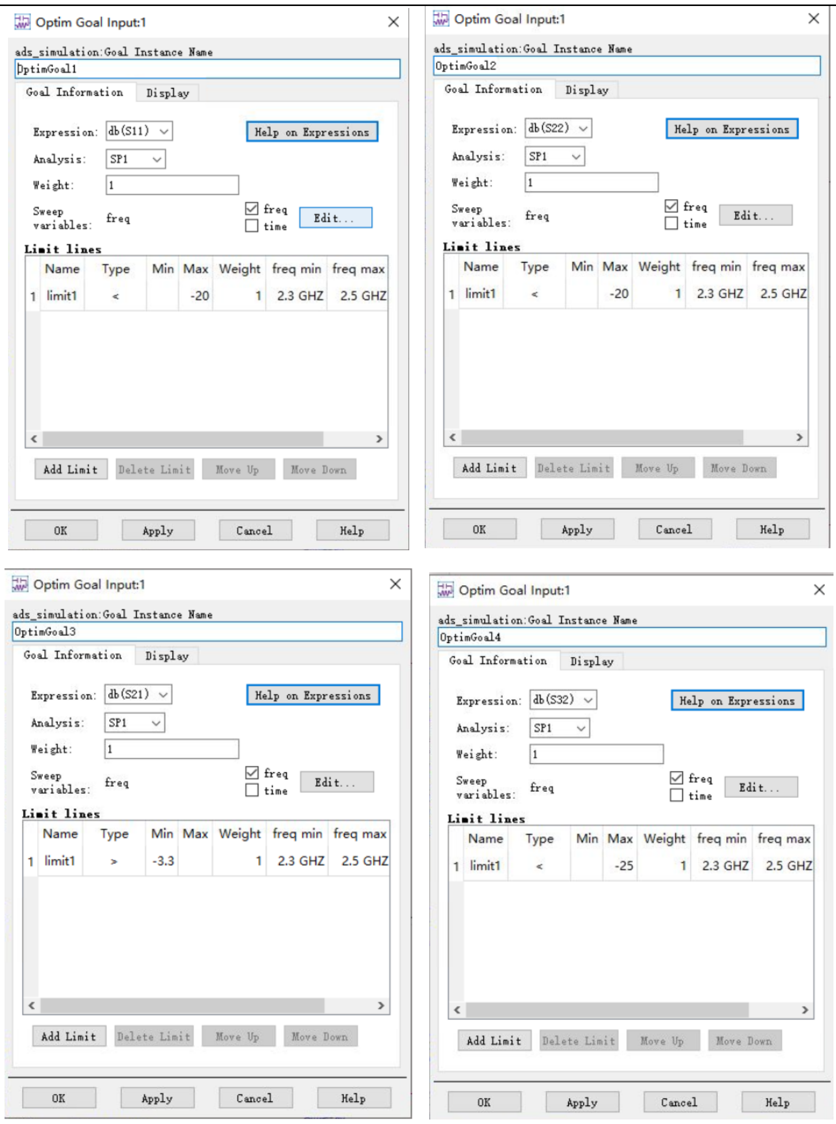




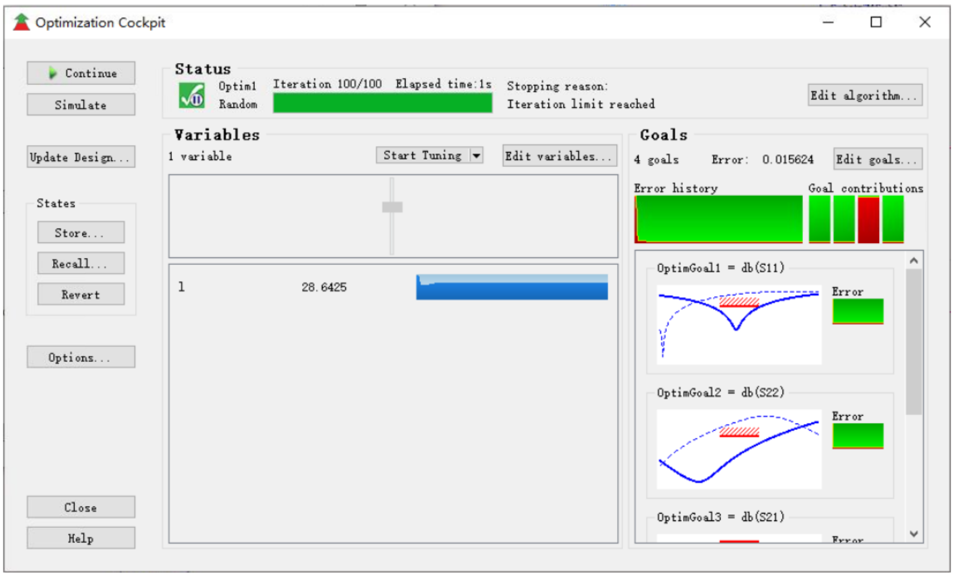
According to the principle of Wilkinson power divider, the length of $Z0 = 50\sqrt{2}\ \Omega$ should be a quarter wavelength. Adjust the microstrip line length to operate at $2.4GHz$. The ADS circuit diagram of 3dB Wilkinson power divider are shown below.



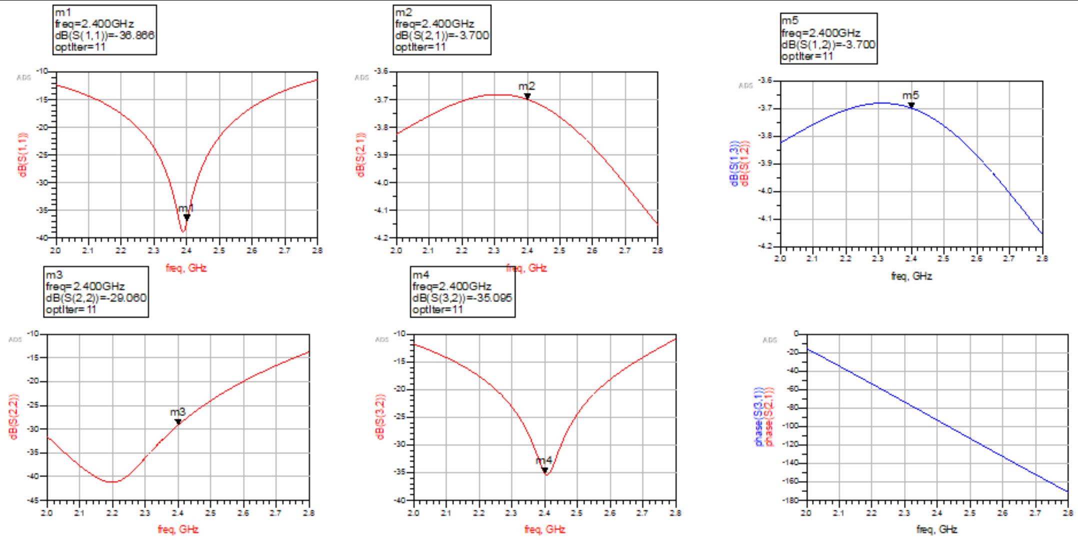
To meet the S-parameter requirements of the device, we need to optimize the parameters Ls using the optimization settings of the ADS, whose optimization range of Ls is $20 - 30mm$. Here are the optimized settings:



The result of the optimized parameter $Ls = 28.6425mm$ is shown below:

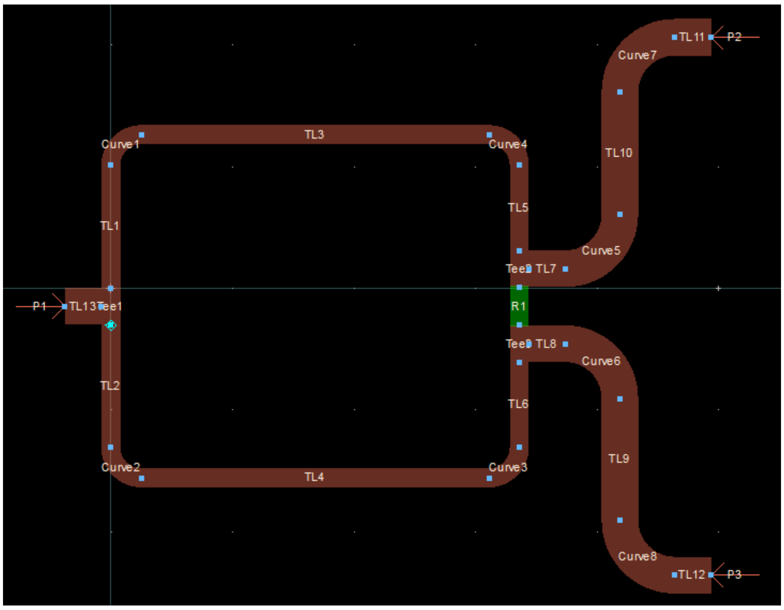


The result of S-parameter after optimization in ADS is shown below,

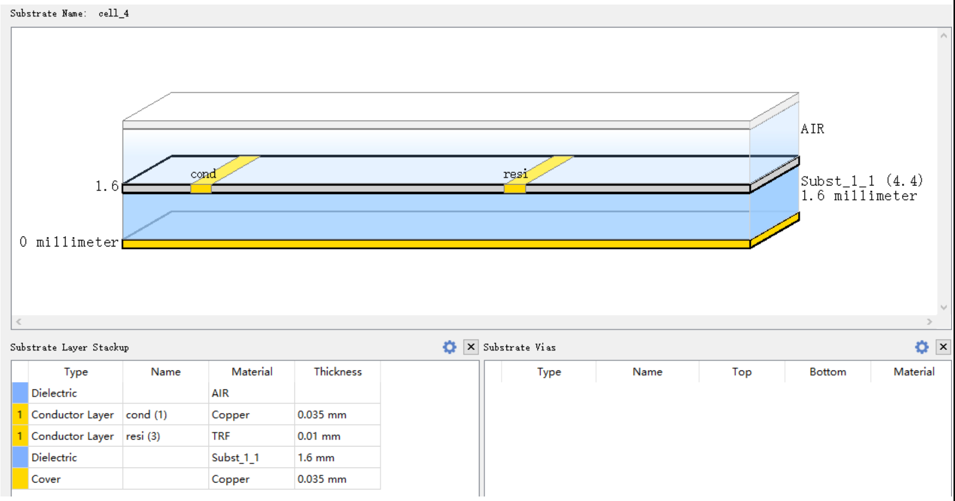


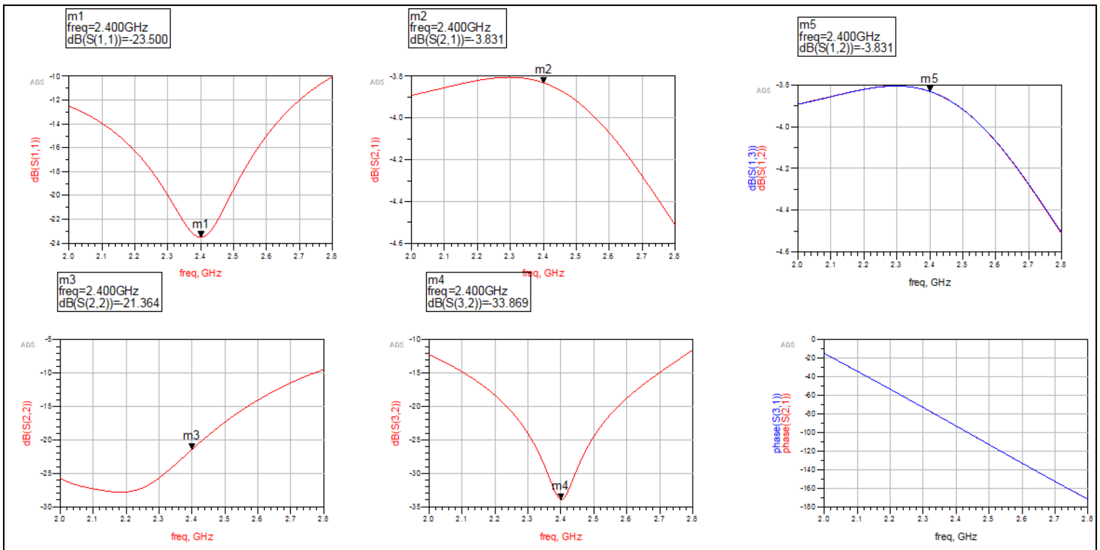
which meets our requirements.

Then, we can generate the layout as follows:



And the substrate is:





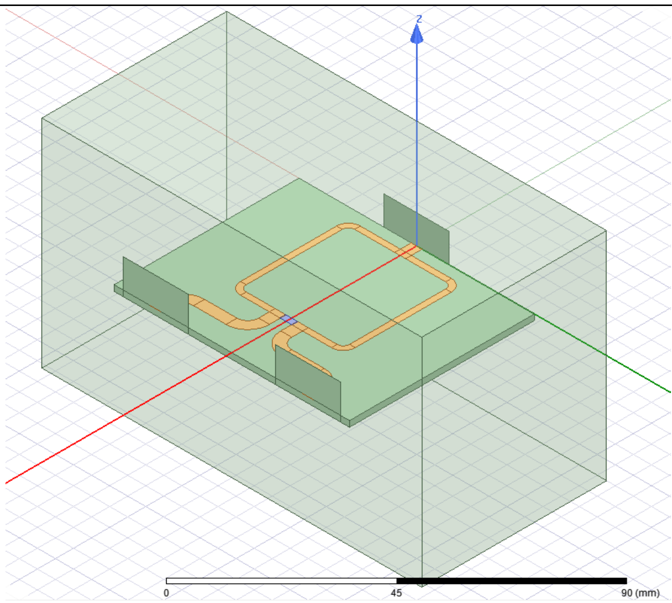
The S-parameters through the EM simulation is as above, which differ from the simulated ideal S-parameters mentioned earlier, showing some degradation, which is because the EM simulation in the layout considers spatial effects and the influence of parasitic parameters, which is normal.

(2) HFSS Modeling

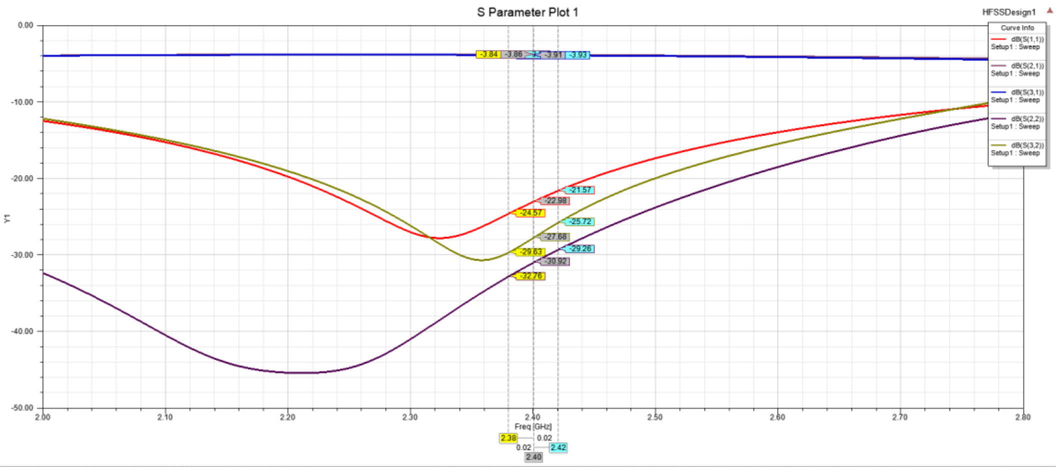
Then we can export this to HFSS to do EM simulation. Add substrate with FR4 and the radiation box. The whole model is shown as below. And we should assign excitation by adding the wave port for the three ports.

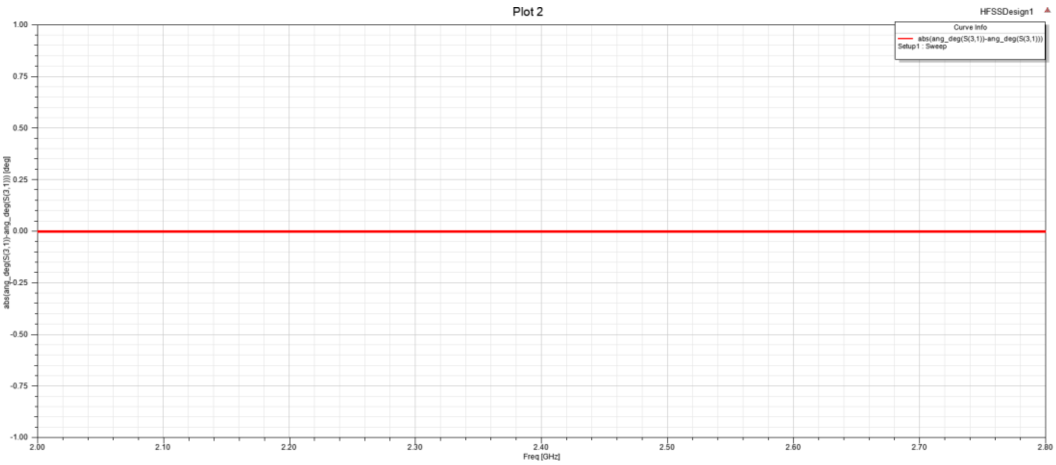
The corresponding variables are shown below.

Name	Value	Unit	Evaluated Va...	Type	Description	Read-only	Hidden	Sweep
extent_x_pos	-3.7857	mm	-3.7857mm	Design		☐	☐	☑
extent_y_pos	-18.1002	mm	-18.1002mm	Design		☐	☐	☑
extent_x_size	28.7647	mm	28.7647mm	Design		☐	☐	☑
extent_y_size	33.1813	mm	33.1813mm	Design		☐	☐	☑
_0_lower_ele...	0	mm	0mm	Design		☐	☐	☑
cond_lower_...	0	mm	0mm	Design		☐	☐	☑
resi_lower_e...	0	mm	0mm	Design		☐	☐	☑
l	51.1717	mm	51.1717mm	Design		☐	☐	☑
l_y	22.5906	mm	22.5906mm	Design		☐	☐	☑
h	1.6	mm	1.6mm	Design		☐	☐	☑



The S-parameters after simulation is as follows:





From the graph, it can be observed that although there is some attenuation in the S-parameters due to the more realistic nature of the HFSS simulation, it still meets our parameter requirements.

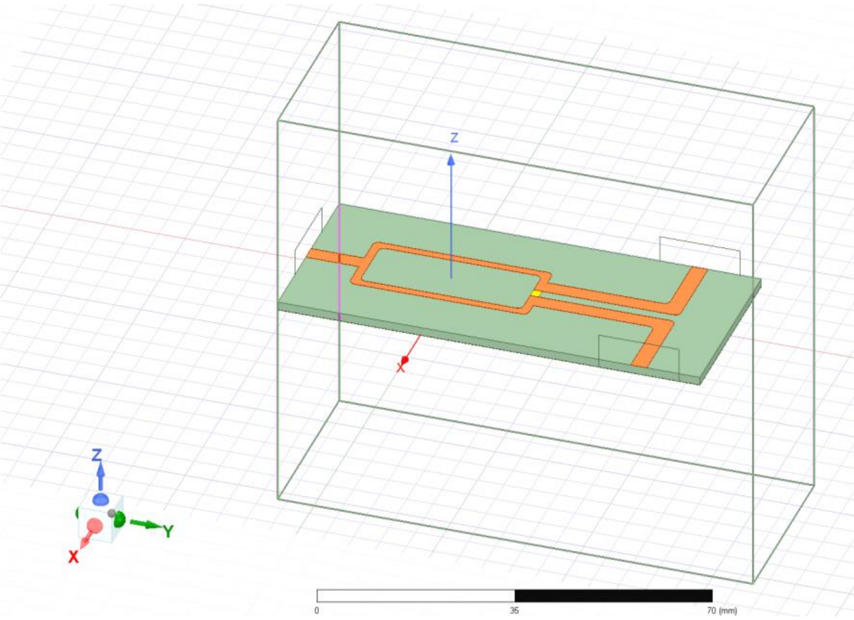
2. 950MHz Wilkinson Power Divider

(1) Modeling

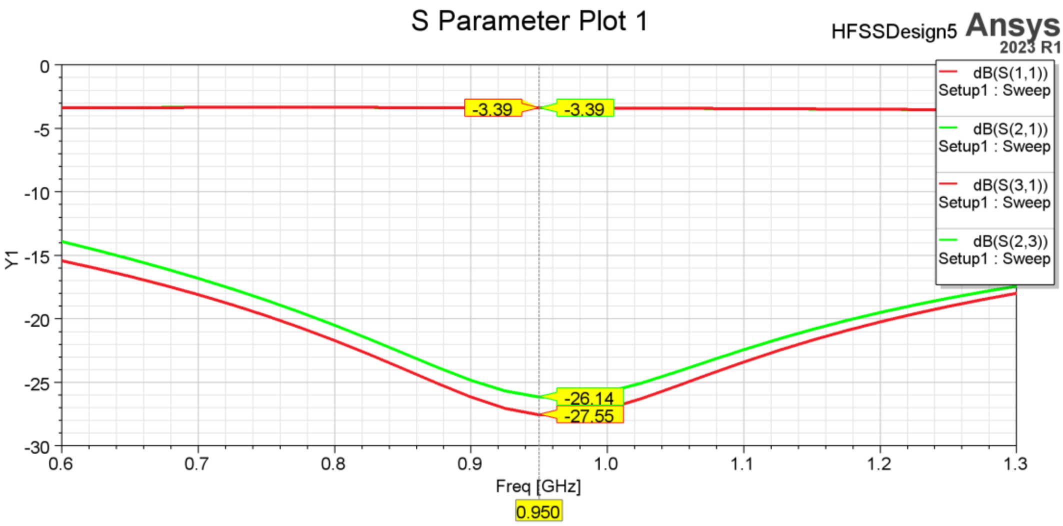
Change the local variables in HFSS as shown below.

Name	Value	Unit	Evaluated Val...	Type	Descripti...	Rea...	Hid...	Swe...
hs	1.6	mm	1.6mm	Design		☐	☐	☑
hc	0.035	mm	0.035mm	Design		☐	☐	☑
Lf	10	mm	10mm	Design		☐	☐	☑
Wf	3.01331	mm	3.01331mm	Design		☐	☐	☑
L1	30.495	mm	30.495mm	Design		☐	☐	☑
W1	1.5677	mm	1.5677mm	Design		☐	☐	☑
L2	7.3125	mm	7.3125mm	Design		☐	☐	☑
L3	22.7975	mm	22.7975mm	Design		☐	☐	☑
Rs	2	mm	2mm	Design		☐	☐	☑

And we get the WPD as shown below.



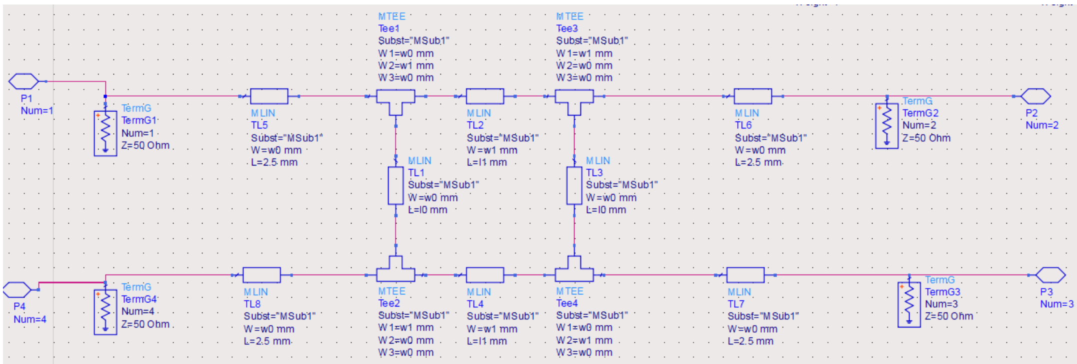
(2) Simulation and Optimetric Result



3. Branch Line Coupler

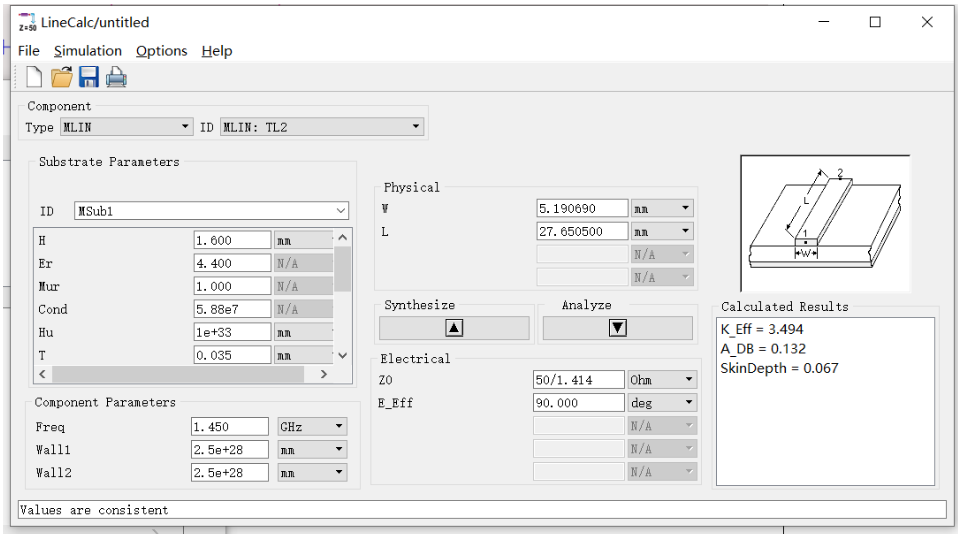
(1) ADS

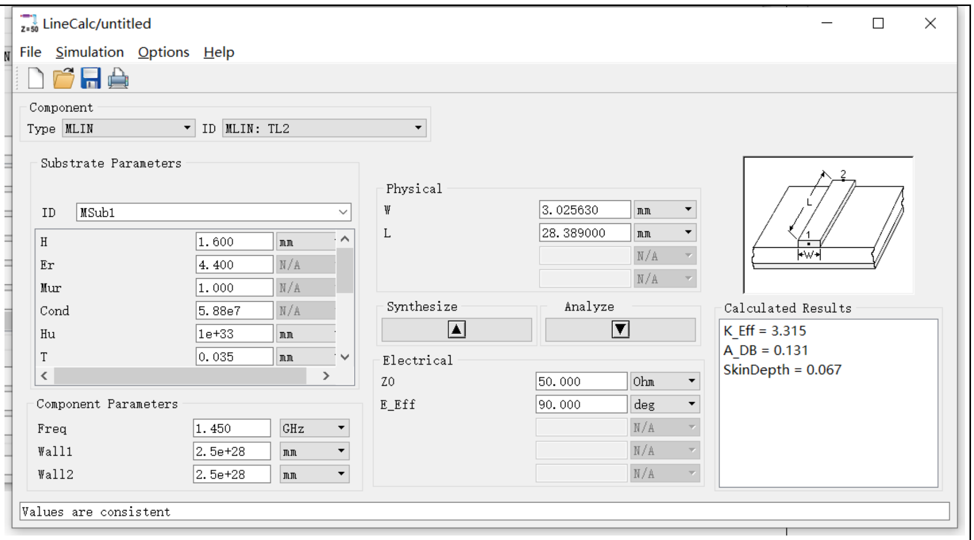
Schematic



Parameters setting

TanD=0.02 when FR4 be our substrate.

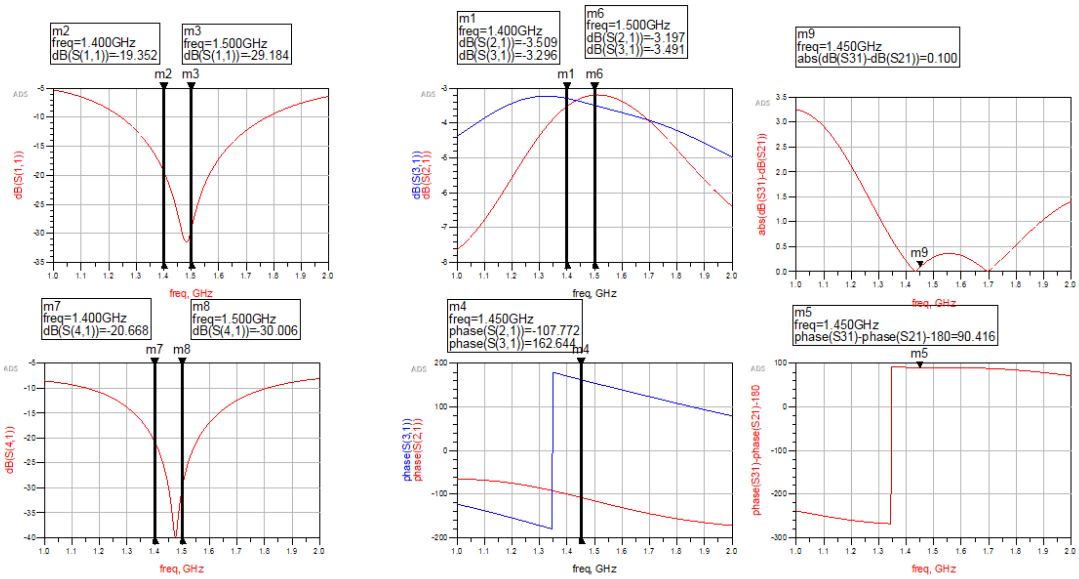




Then calculate the width of transmission line of both.

$$W1 = 5.19mm, W2 = 3.025mm$$

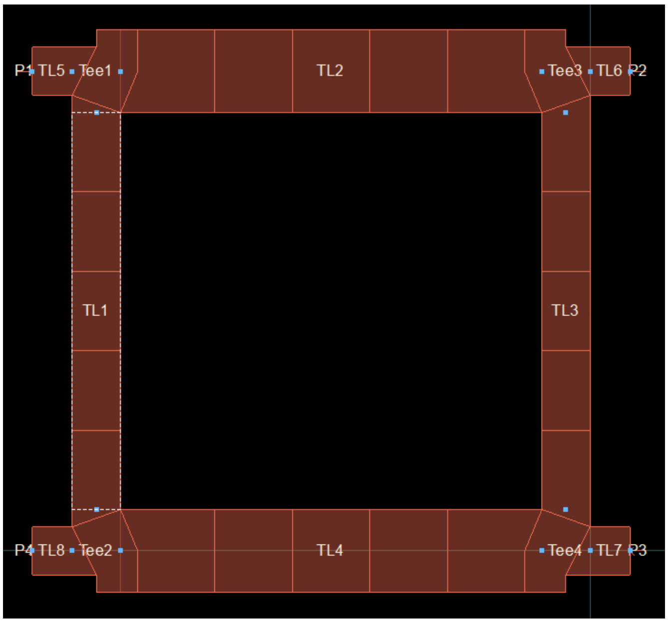
Next, we set goals to simulate to reach the requirements.



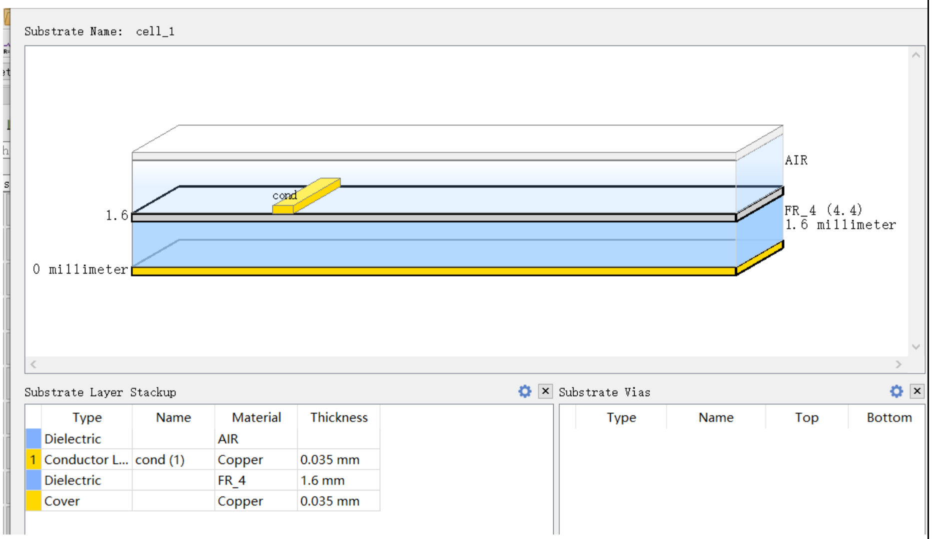
When TanD=0.02 of FR4, the tangent loss will cause S21 and S31 being closed to -3.3dB. But they will not be larger than -3.3dB meanwhile in the case of meeting other requirements.

(2) EM simulation

Layout

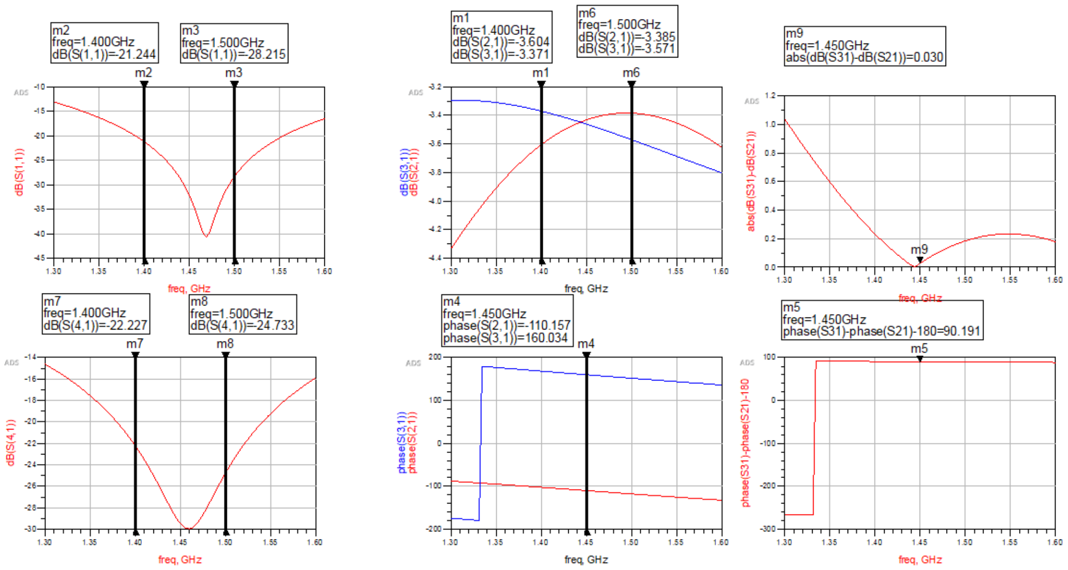


Substrate



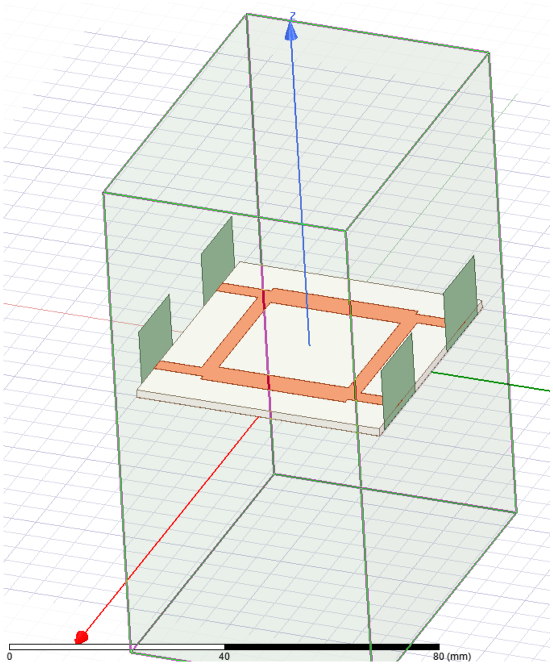
We adjust the thickness of copper to 0.035mm.

Simulation result



(3) HFSS

Model



FR4 as the substrate material.

HFSS parameters settings

Name	Value	Unit	Evaluated...	Type	Description	Read-o...	Hidden	Sweep
Lf	10	mm	10mm	Design		☐	☐	☑
L1	30.807235882233	mm	30.80723...	Design		☐	☐	☑
L2	29.525978432617	mm	29.52597...	Design		☐	☐	☑
hs	1.6	mm	1.6mm	Design		☐	☐	☑
hc	0.035	mm	0.035mm	Design		☐	☐	☑
Wf	3.02563	mm	3.02563m...	Design		☐	☐	☑
W1	3.02563	mm	3.02563m...	Design		☐	☐	☑
W2	5.19069	mm	5.19069m...	Design		☐	☐	☑

Optimization result

OptimizationSetup1

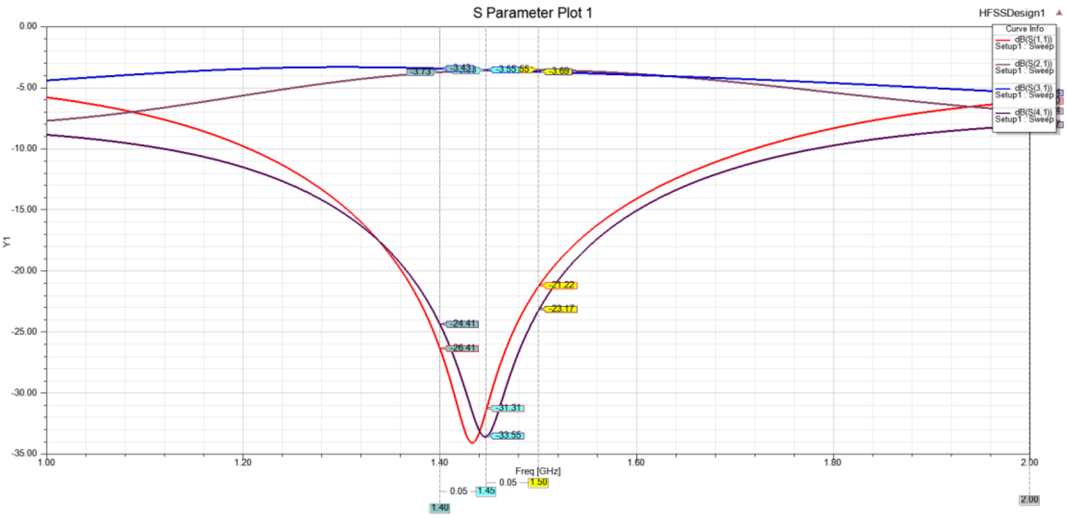
Result | Profile

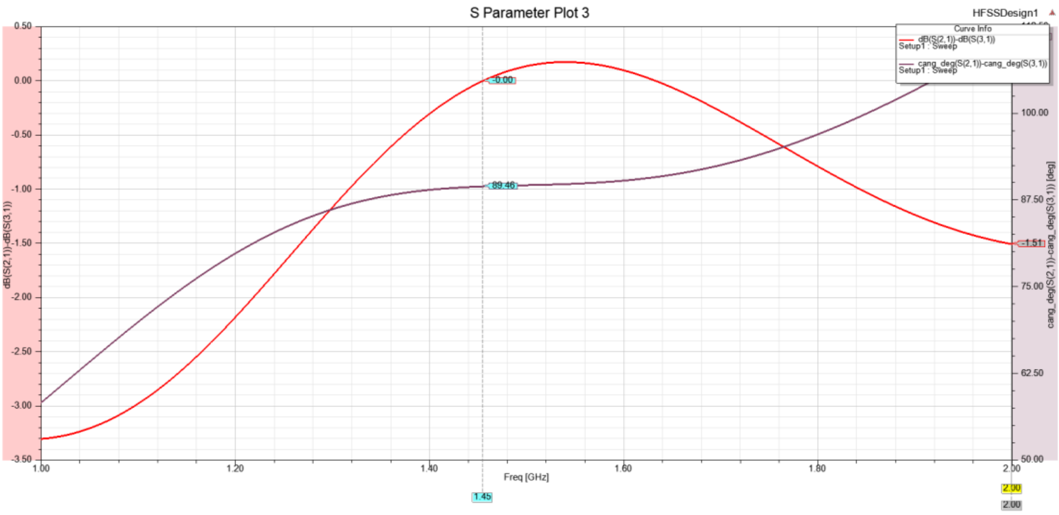
View: (•) Table ( ) Plot

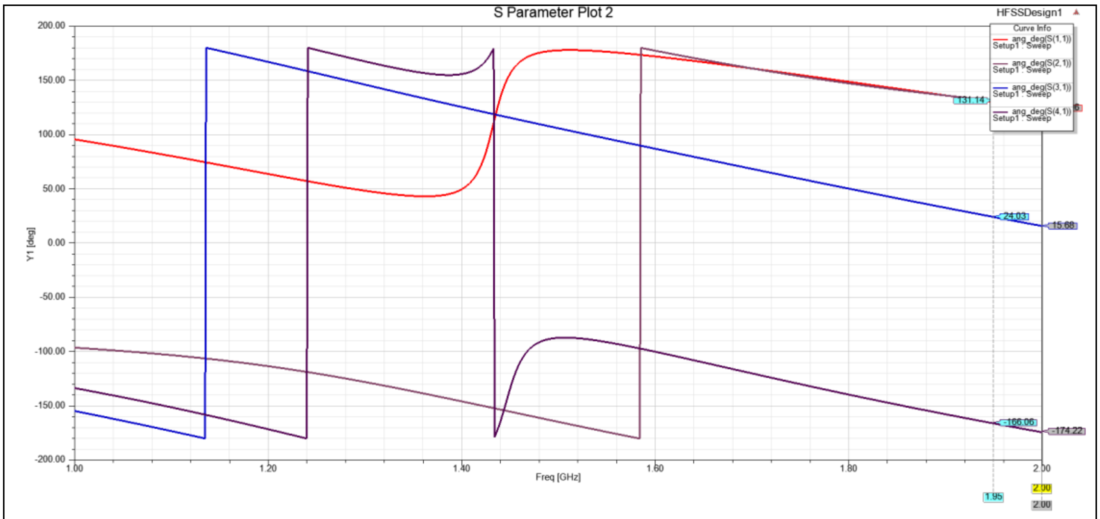
Evaluation	L1	L2	Cost
9	30.81mm	29.53mm	0.16332
17	30.82mm	29.52mm	0.16419
1	30.81mm	29.71mm	0.16453
8	30.61mm	29.71mm	0.16589
13	30.81mm	29.69mm	0.16796
12	31.04mm	29.53mm	0.17797
15	30.81mm	29.36mm	0.18794
16	30.93mm	29.45mm	0.2155
14	30.58mm	29.53mm	0.22215
10	30.75mm	29.6mm	0.24602
7	30.81mm	28.09mm	0.33631
11	30.8mm	29.7mm	0.34955
5	30.54mm	29.47mm	0.38575
2	32.45mm	29.71mm	0.39003
6	29.16mm	29.71mm	0.4931
3	30.81mm	31.33mm	0.60224
4	28.09mm	27.27mm	18.493

We choose L1=30.81mm, L2=29.53mm to make it out.

S-parameters results







Except S21 and S31 out of range which due to the TanD(0.02), others all satisfy the requirements.

4. Patch Antenna

(1) Modeling

Change the local variables in HFSS as shown below.

Name	Value	Unit	Evaluated Va...	Type	Description	Read-only	Hidden	Sweep
H	1.6	mm	1.6mm	Design		☐	☐	☑
W0	45	mm	45mm	Design		☐	☐	☑
L0	30.05	mm	30.05mm	Design		☐	☐	☑
W1	1.2	mm	1.2mm	Design		☐	☐	☑
L1	17.71	mm	17.71mm	Design		☐	☐	☑
W2	3.1	mm	3.1mm	Design		☐	☐	☑
L2	15	mm	15mm	Design		☐	☐	☑

$$W_p = 45\text{mm}$$

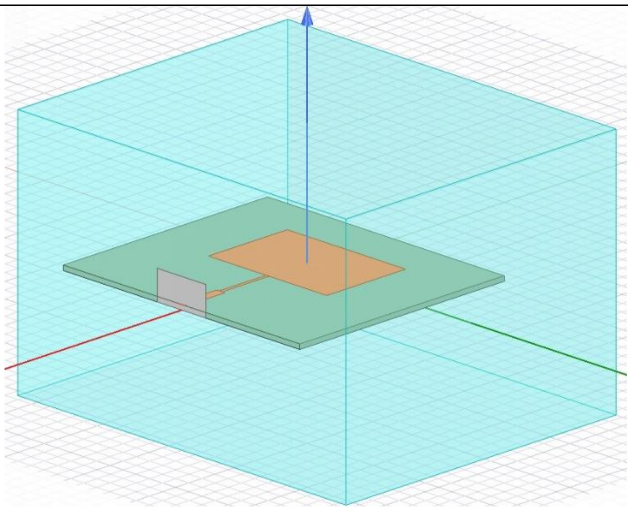
$$\frac{W_p}{h} = \frac{45mm}{1.6mm} = 28.15 > 1$$

$$\varepsilon_e = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[1 + 12\frac{h}{W_p}\right]^{-\frac{1}{2}} = 4.12$$

$$\Delta L_p = 0.412\text{h}\frac{(\varepsilon_e + 0.3)\left(\frac{W}{h} + 0.264\right)}{(\varepsilon_e - 0.258)\left(\frac{W}{h} + 0.8\right)} = 0.74$$

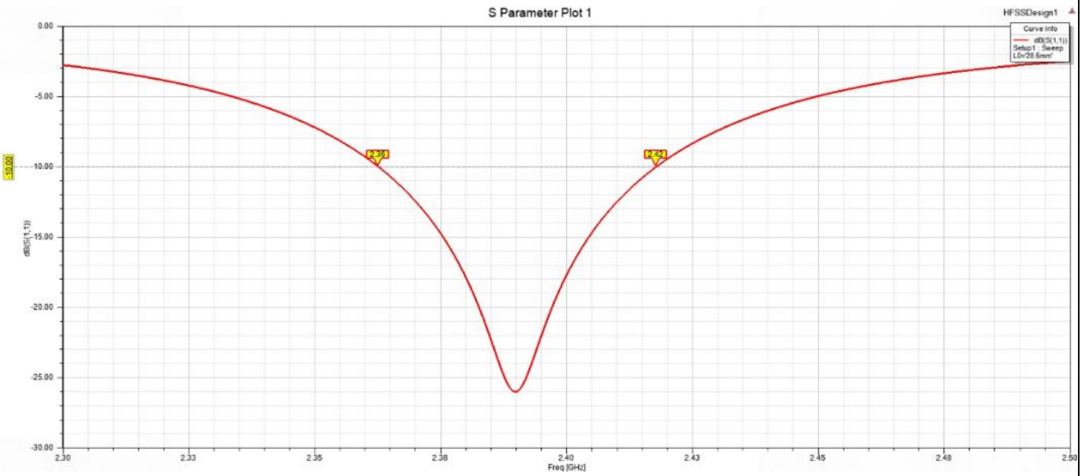
$$L_p = \frac{c}{2f\sqrt{\varepsilon_e}} - 2\Delta L_p = 30.05$$

And we get the patch antenna model as shown below.



(2) First Simulation Result

Reset the frequency sweep region to (2.3GHz, 2.5GHz), and we get the $dB(S_{11})$ as shown below.

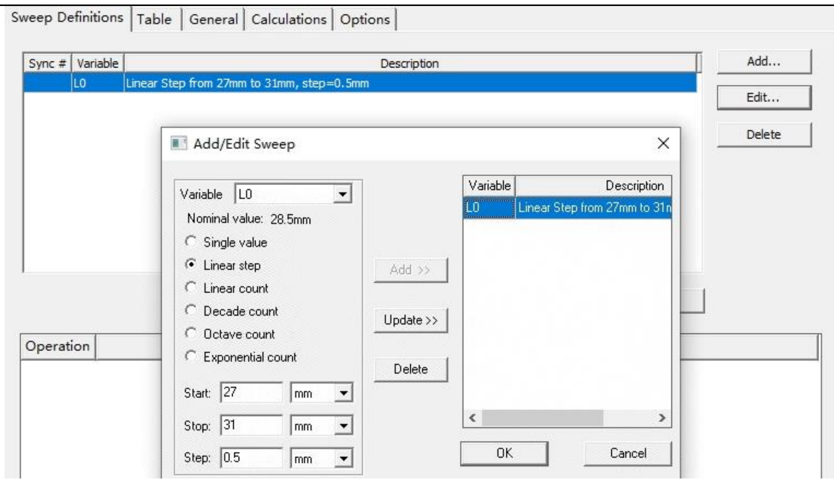


However, although the results have satisfied the experimental requirements, they are not perfect because the lowest point of $dB(S_{11})$ is not taken at the set center frequency $f_c = 2.4GHz$.

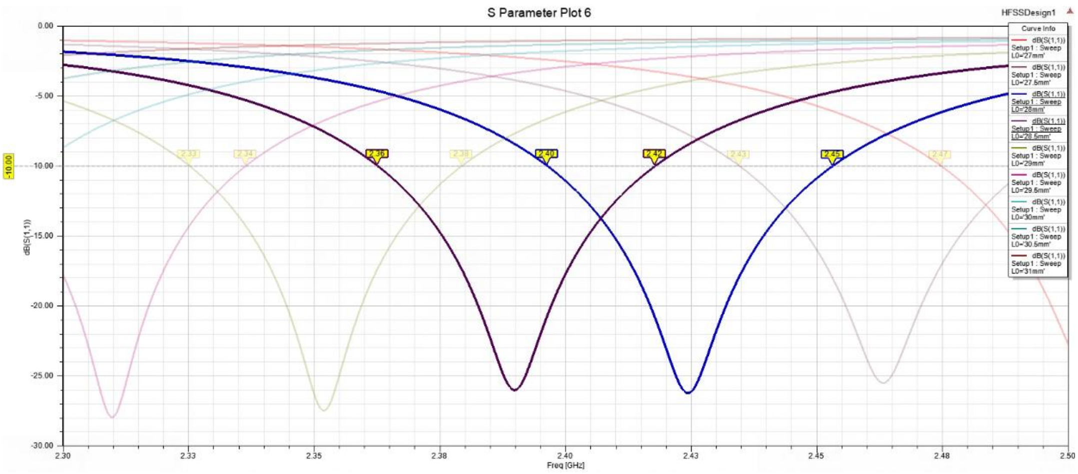
So, we do need *Optometrics*.

(3) Optimetrics

First, we estimate a general range of L_0in the *Sweep Analysis* settings, say $27 - 31mm$ with step of $0.5mm$.

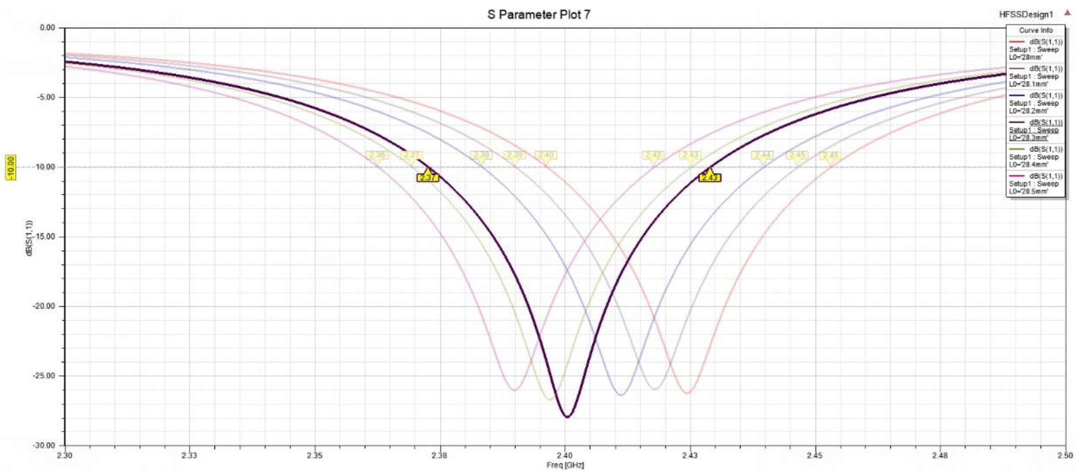


And we get the corresponding $dB(S_{11})$ result of each L_0.

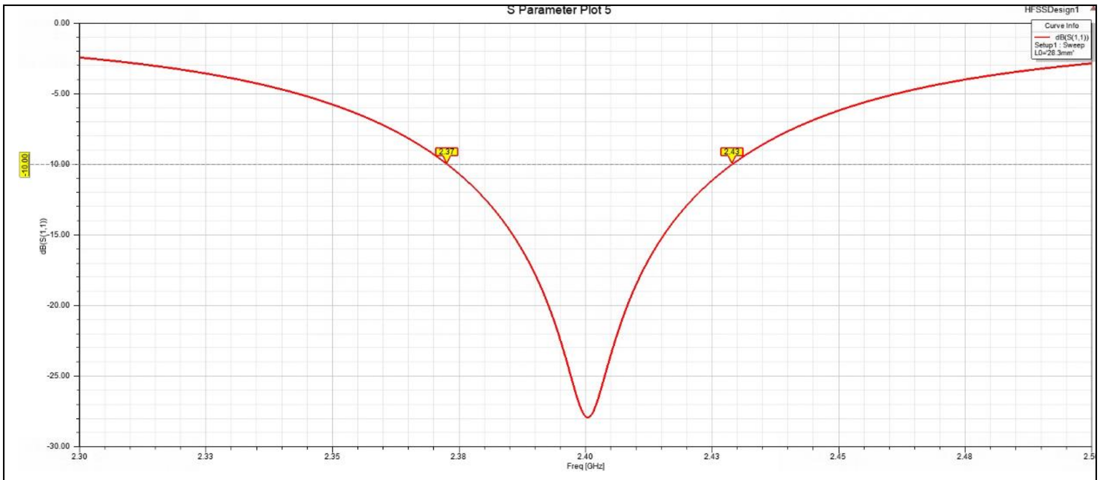


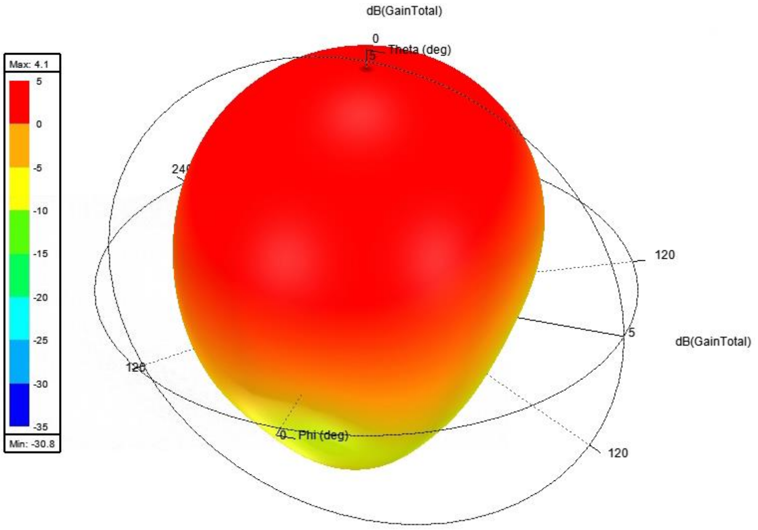
From the figure above, we can see when $L_0 = 28mm$ or $L_0 = 28.5mm$, the $dB(S_{11})$ results are much better than others.

So, we set the sweep range to $28 - 28.5mm$ with step of $0.1mm$.



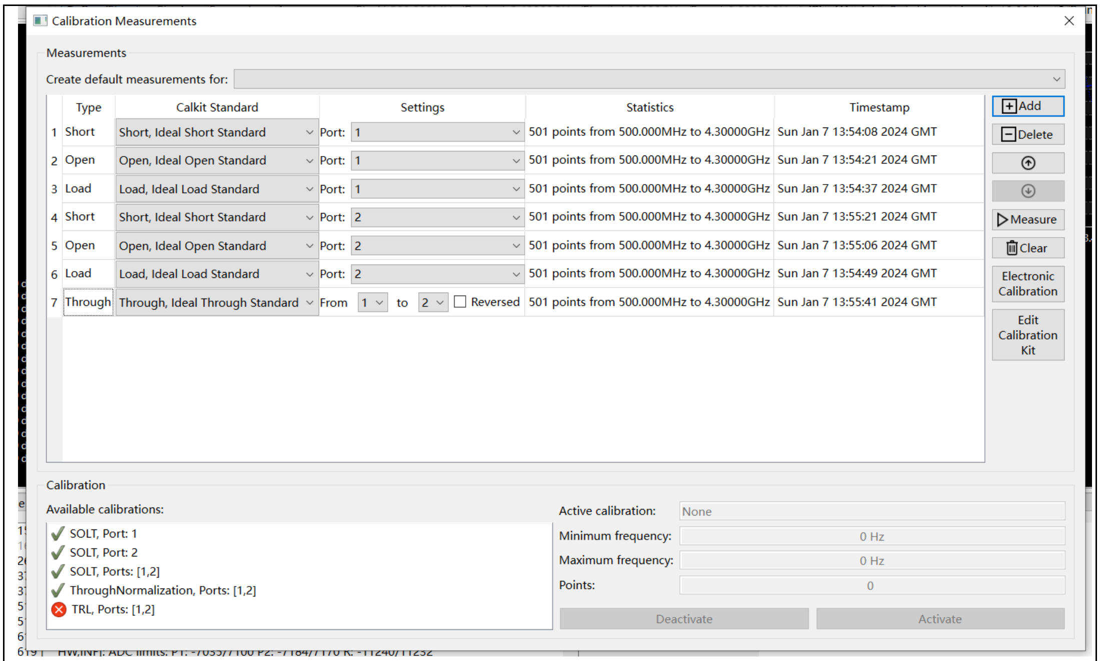
As we can see, when $L_0 = 28.3mm,$ the corresponding $dB(S_{11})$ result is the best. So, the final L_0 we need is $28.3mm$. The final $dB(S_{11})$ and 3D polar of far field are shown below, respectively.

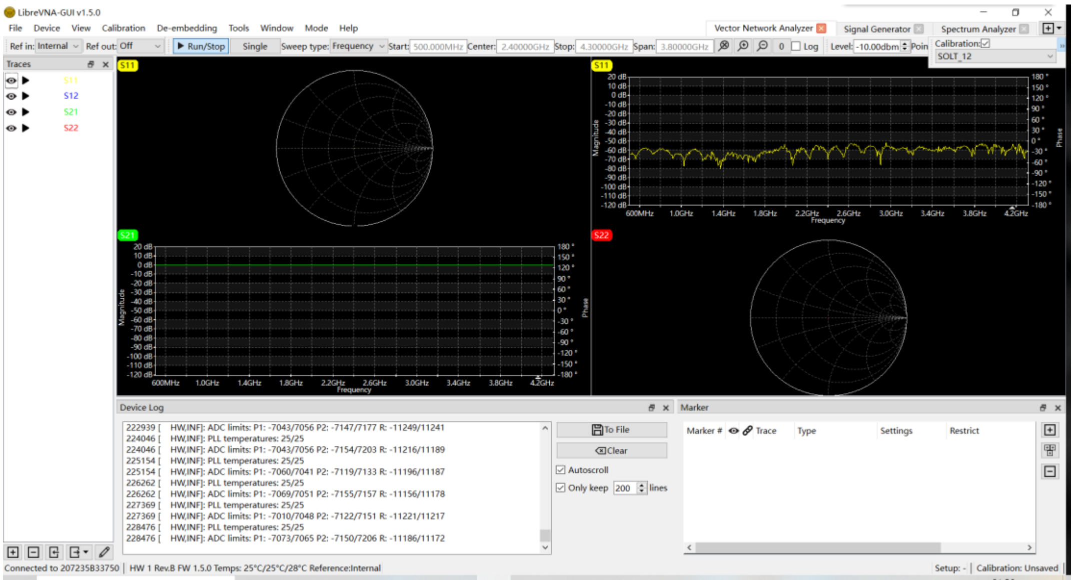




LibreVNA Test Result

1. Calibration Result

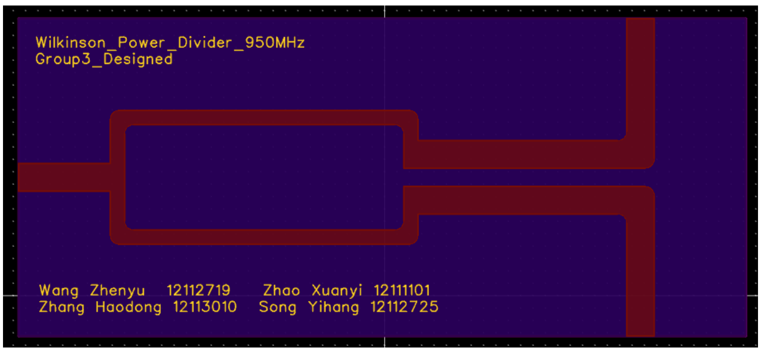


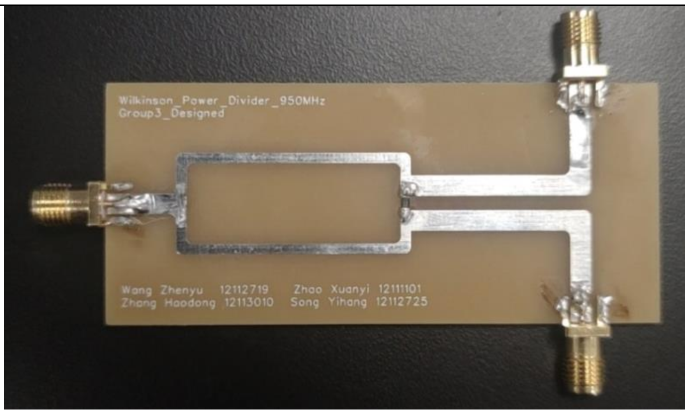


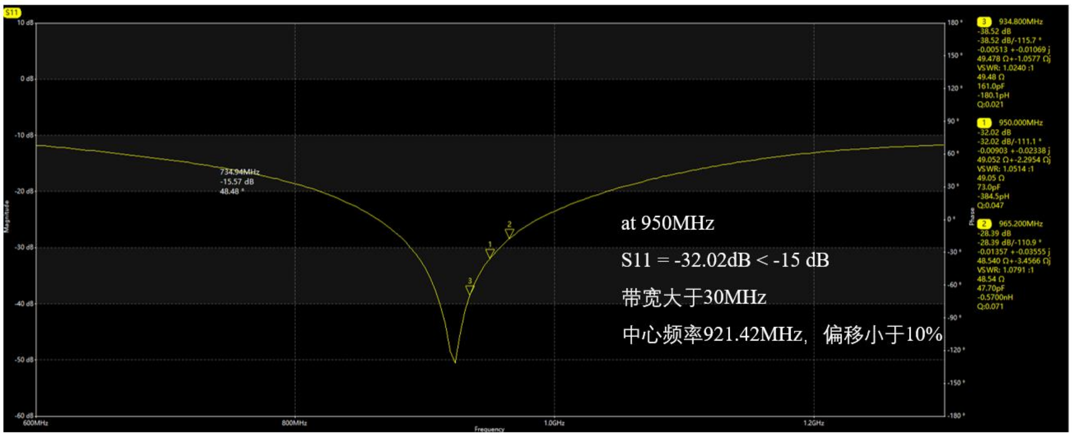
2. Device Test Results

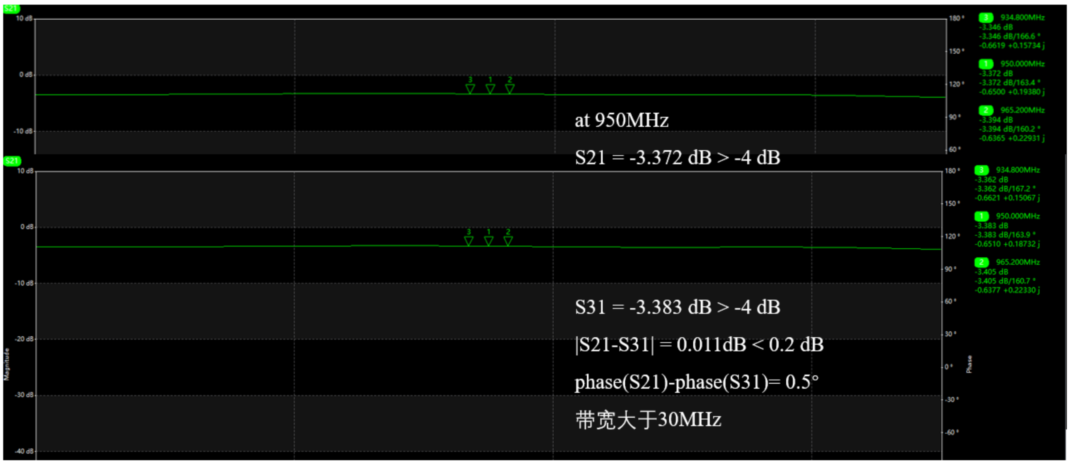
(1) 950MHz Wilkinson Power Divider

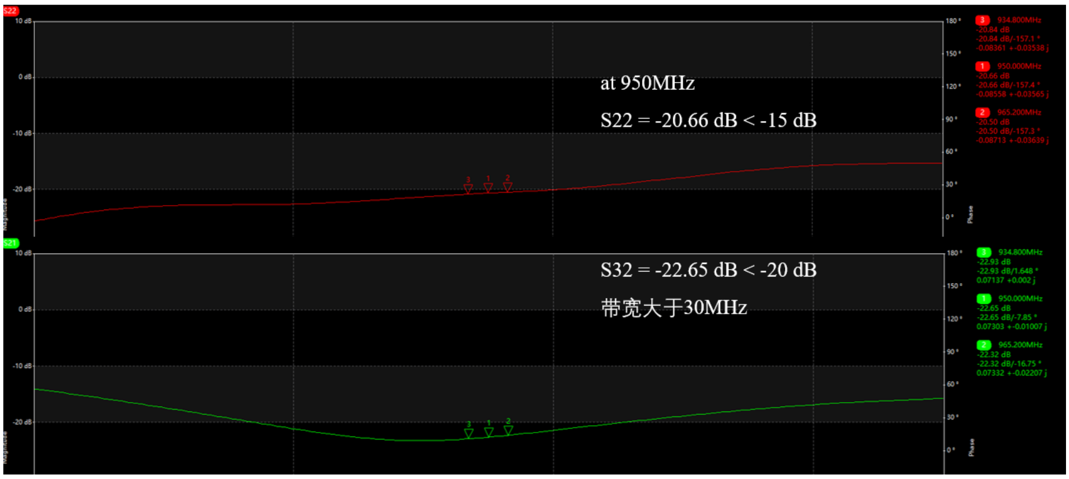
威尔金森功分器 950M： S11<-15dB；S22<-15dB；S21>-4dB；|S21-S31|<0.2dB；S21、S31 相位差<1°；S32<-20dB；满足以上条件的带宽大于 30MHz；中心频率偏移<10%（950MHz 的 10%范围是 0.885GHz-1.045GHz）




Wilkinson_Power_Divider_950MHz
Group3_Designed
Wang Zhenyu 12112719
Zhang Haodong 12113010
Zhao Xuanyi 12111101
Song Yihang 12112725

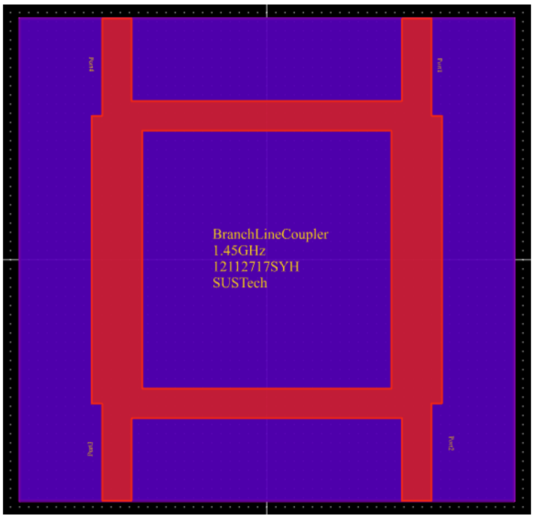

at 950MHz
S11 = -32.02dB < -15 dB
带宽大于30MHz
中心频率921.42MHz，偏移小于10%

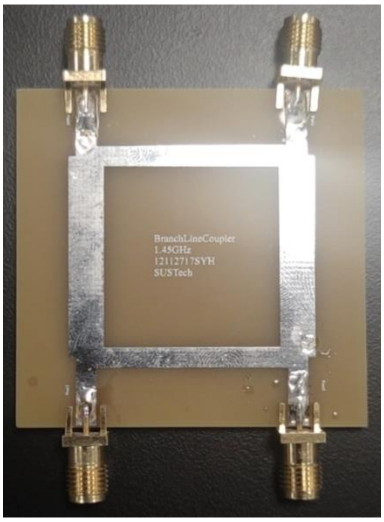

at 950MHz
S21 = -3.372 dB > -4 dB
S31 = -3.383 dB > -4 dB
|S21-S31| = 0.011dB < 0.2 dB
phase(S21)-phase(S31)= 0.5°
带宽大于30MHz

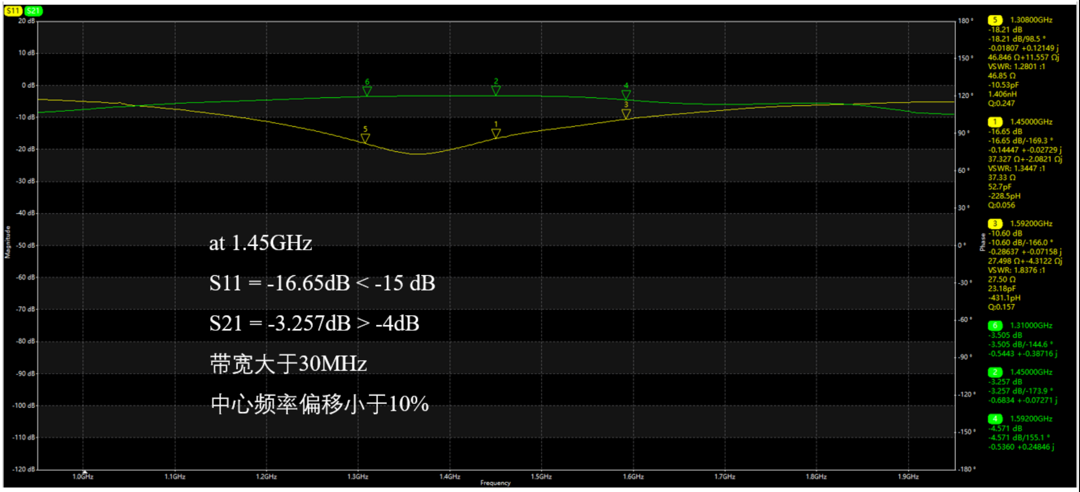

at 950MHz
S22 = -20.66 dB < -15 dB
S32 = -22.65 dB < -20 dB
带宽大于30MHz

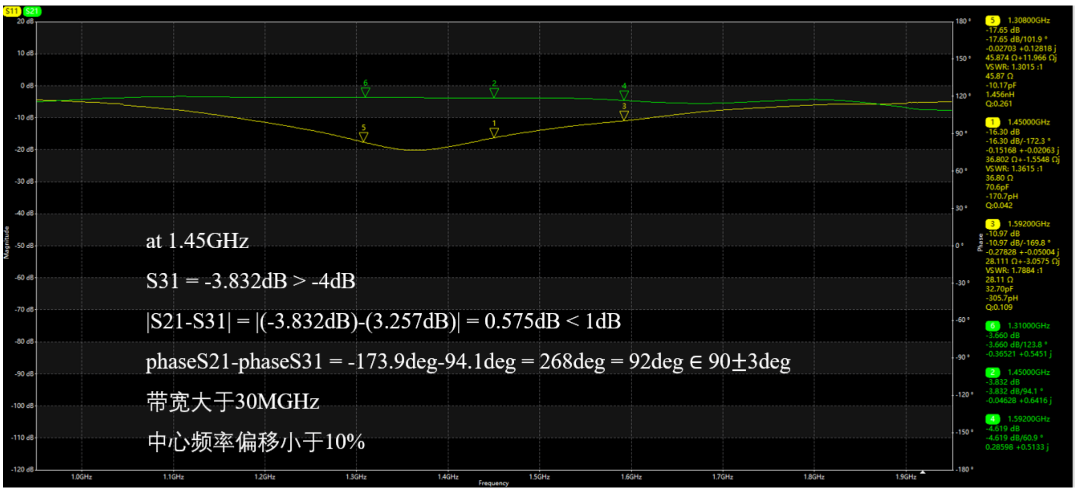
(2) 1.45GHz Branch Line Coupler

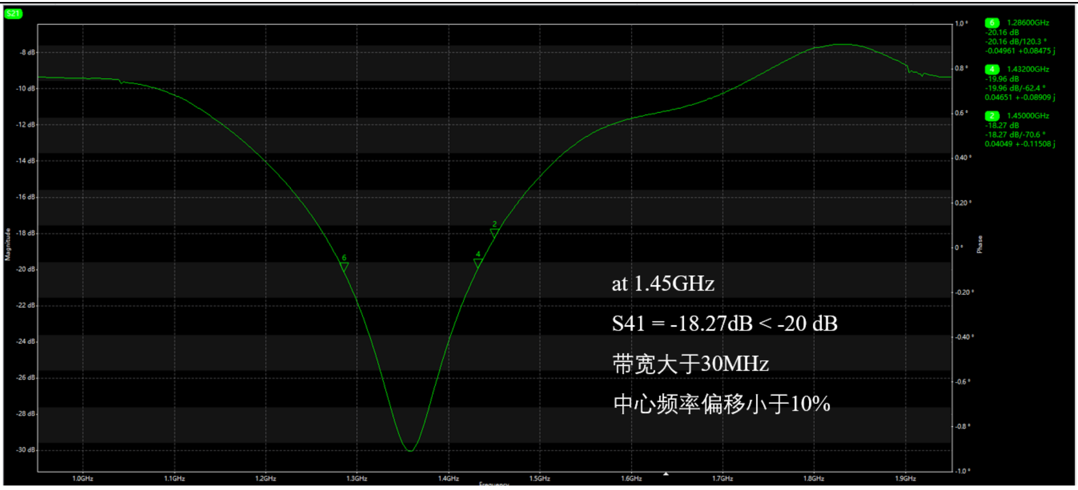
分支线耦合器 1.45G：S11<-15dB；S21>-4dB；S31>-4dB；|S21-S31|<1dB；S21、S31 相位差 90±3°；S41<-20dB；满足以上条件的带宽大于 30MHz；中心频率偏移<10%（1.45GHz 的 10%范围是 1.305GHz-1.595GHz）





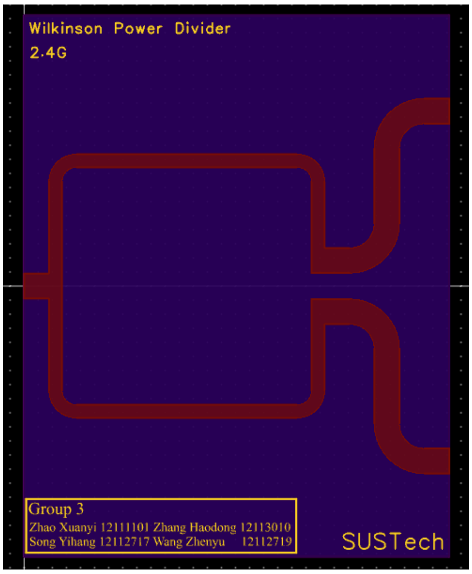


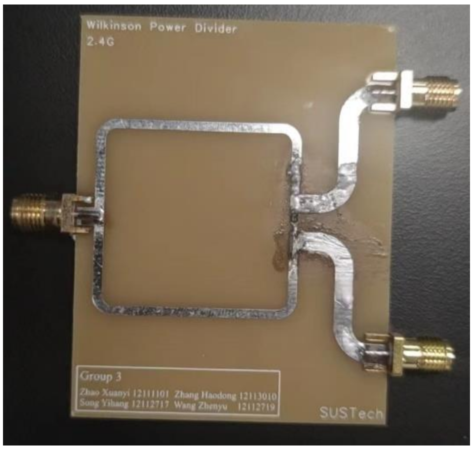


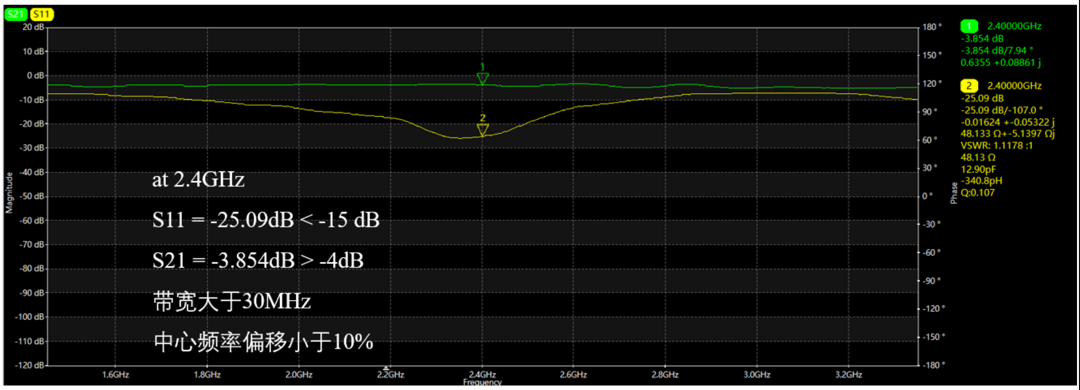


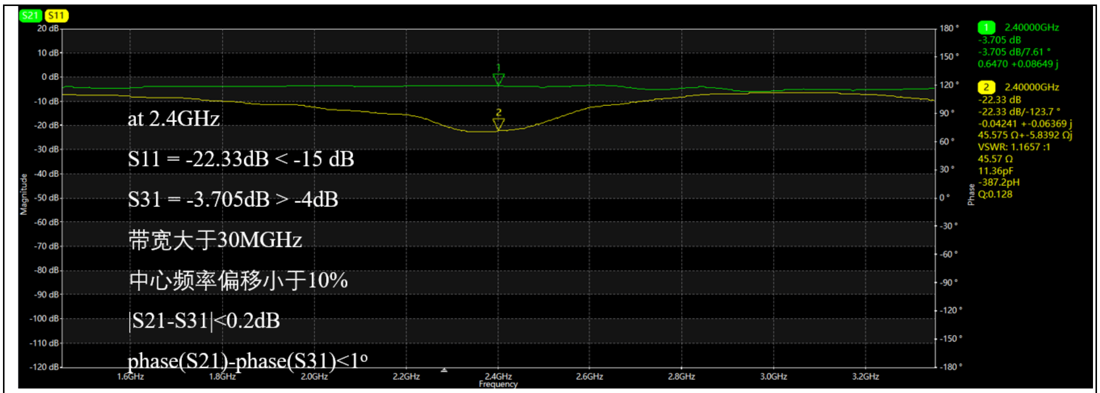
(3) 2.4GHz Wilkinson Power Divider

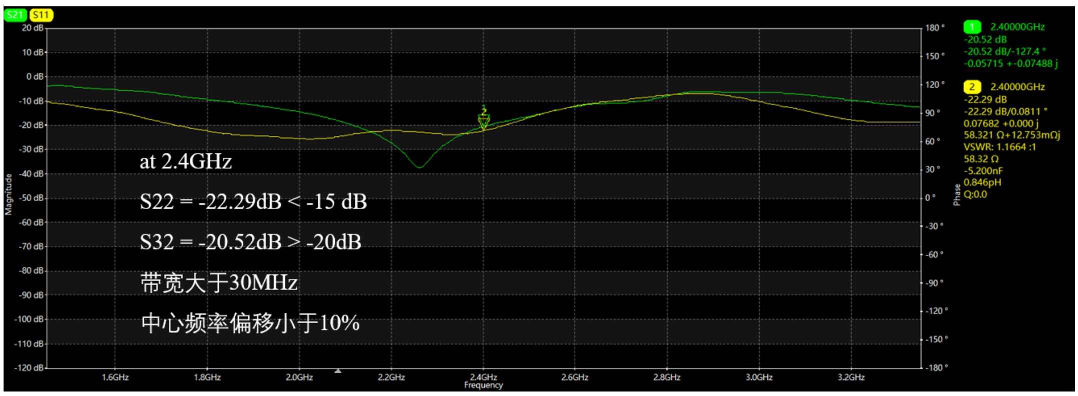
威尔金森功分器 2.4G: S11<-15dB; S22<-15dB; S21>-4dB; |S21-S31|<0.2dB; S21、S31 相位差<1° ; S32<-20dB; 满足以上条件的带宽大于 30MHz; 中心频率偏移<10% (2.4GHz 的 10%范围是 2.16GHz-2.64GHz)





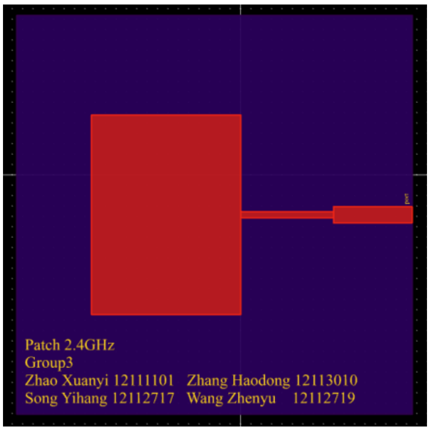


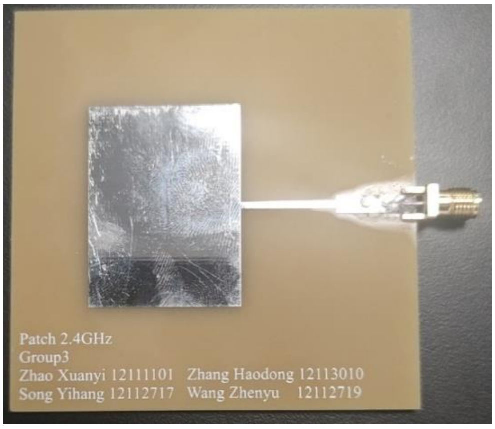


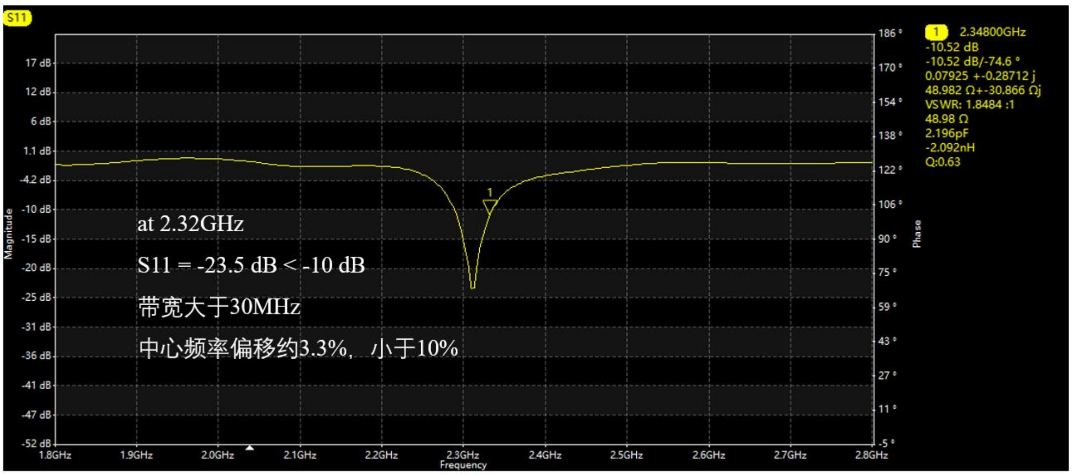


(4) 2.4GHz Patch Antenna

微带贴片天线 2.4G：S11<-10dB；满足以上条件的带宽大于 30MHz；中心频率偏移<10%（2.4GHz 的 10%范围是 2.16GHz-2.64GHz）





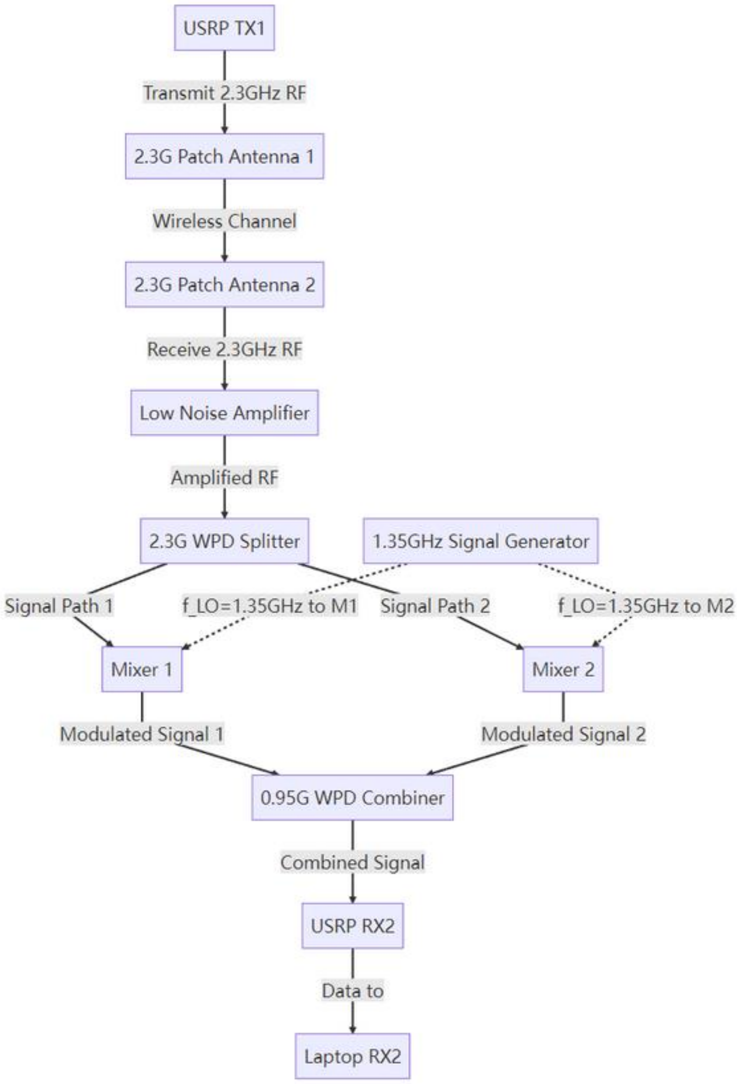


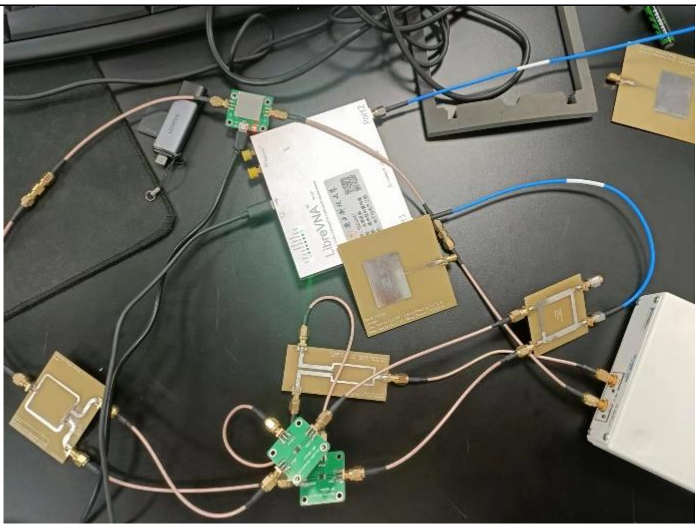
3. Analysis of test Results & Directions for Improvement

As can be seen from the demonstration diagrams above, the S-parameters, bandwidths, and the degree of center frequency shift of these four devices are all fully satisfied. Furthermore, the wireless communication system built from these devices works well, we don't have a clear idea of how to improve the devices for the time being.

Wireless Communication System Construction

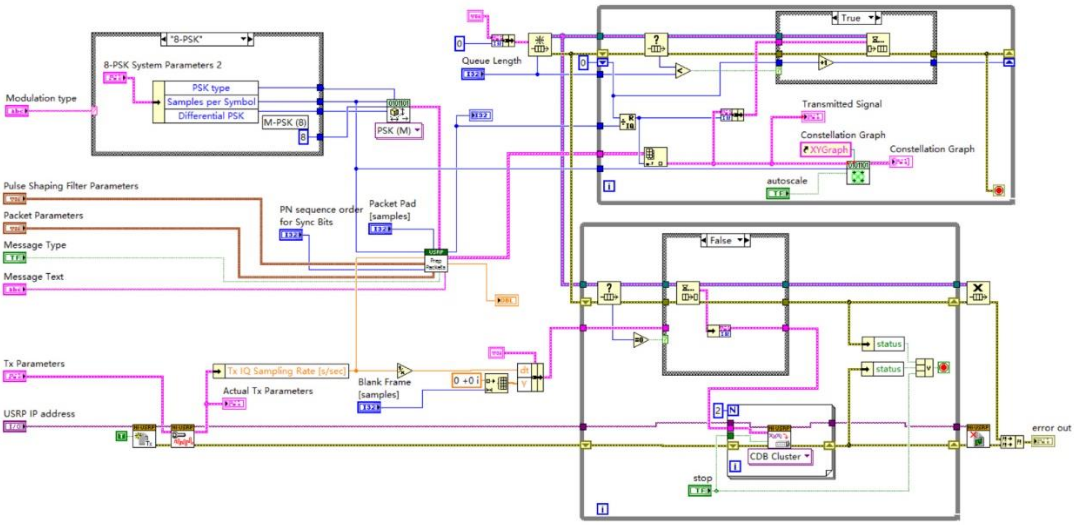
1. System Block Diagram and Physical Diagram



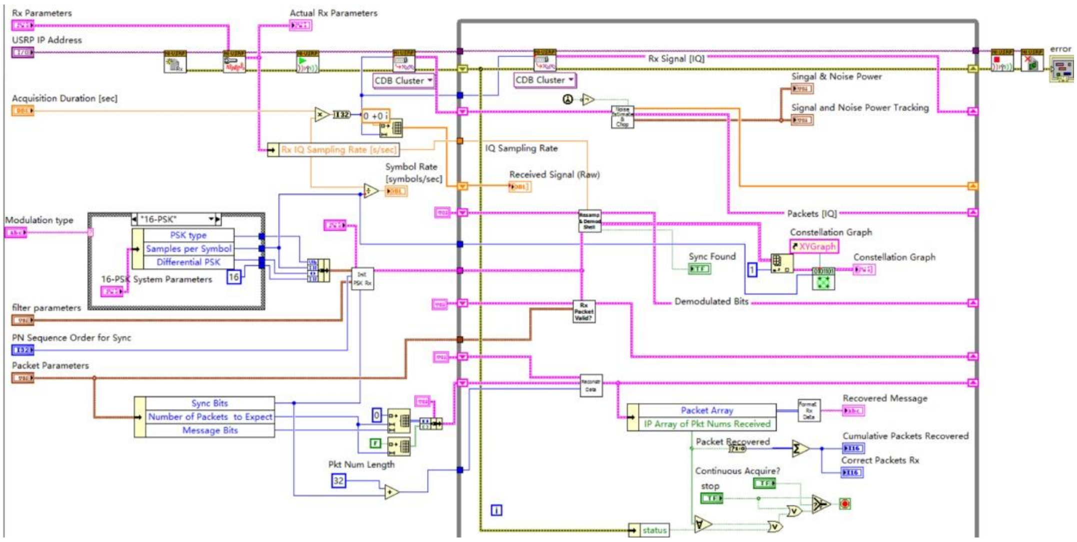


2. Text Transmission

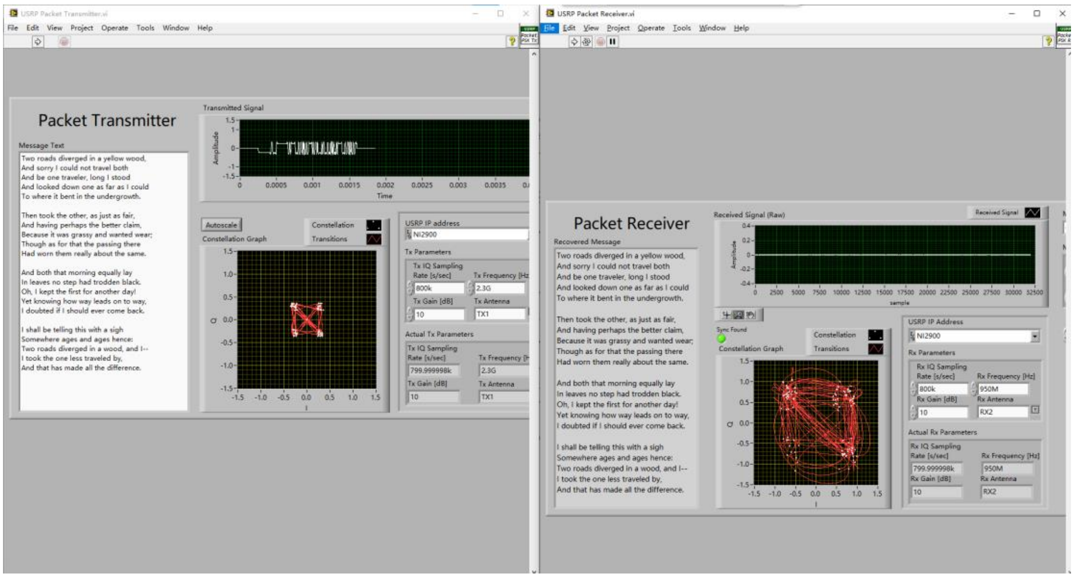
➢ Transmitter LabView Program



➢ Receiver LabView Program

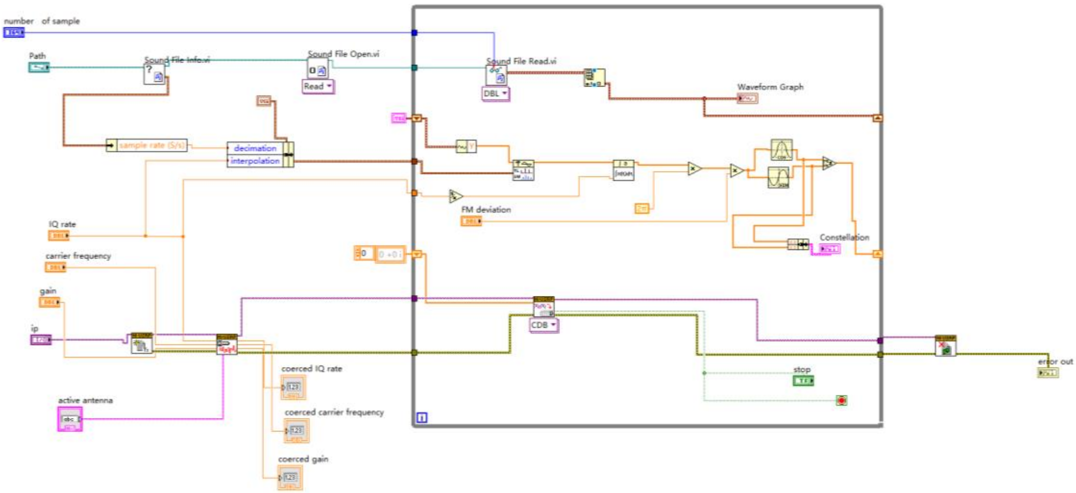


➢ **Result** (See PPT for video presentation of results)

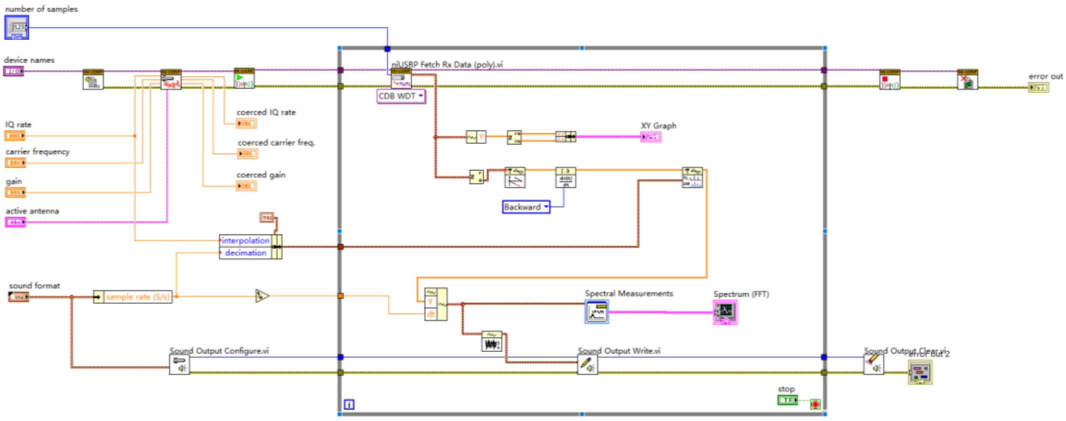


3. Voice Transmission

➢ **Transmitter LabView Program**



➢ **Receiver LabView Program**



- **Result**

  See PPT for video presentation of results.

Experience

1. During the simulation process, we overlooked the dimensions of the resistor in the Wilkinson power divider, leading to difficulties during the subsequent physical soldering. We successfully addressed this issue by bridging the pins with shortened jumper wires. This experience emphasizes the importance of considering real-world dimensions and constraints when simulating and designing devices.
2. Additionally, when using the mixer, special attention must be paid to its input ports. The local oscillator signal enters through the LO port, the radio frequency signal enters through the RF port, and care should be taken not to confuse them. Connecting to the wrong port may result in unsuccessful mixing.
3. Even after down-conversion, signals may still exist at the original frequency, indicating signal leakage. Therefore, careful consideration of frequency selection is crucial.
4. Issue with pre-mixer frequency offset: Is there a frequency offset? Answer: Yes, our system exhibits a certain degree of frequency offset. The initially designed 2.4 GHz system performs optimally at 2.3 GHz. Consequently, during actual testing, we selected 2.3 GHz as the input signal frequency and 950 MHz as the output signal frequency to achieve the best system performance.

Team Division of Labor (25%, 25%, 25%, 25%)

- **赵宣懿：**Introduction to the principle of patch antenna; Theoretical basis for patch antenna design dimensioning; HFSS and ADS simulation of patch antenna; LibreVNA testing of patch antenna; Wireless communication system construction (voice transmission); PPT production.
- **张皓东：** Introduction to the principle of 2.4GHz Wilkinson Power Divider; Theoretical basis for 2.4GHz Wilkinson Power Divider design dimensioning; HFSS and ADS simulation of 2.4GHz Wilkinson Power Divider; LibreVNA testing of 2.4GHz Wilkinson Power Divider; Wireless communication system construction (text transmission); PPT production.

➢ **王振宇：** Introduction to the principle of 950MHz Wilkinson Power Divider; Theoretical basis for 950MHz Wilkinson Power Divider design dimensioning; HFSS and ADS simulation of 950MHz Wilkinson Power Divider; LibreVNA testing of 950MHz Wilkinson Power Divider; Wireless communication system construction (voice transmission); PPT production. ➢ **宋宜航：** Introduction to the principle of Branch Line Coupler; Theoretical basis for Branch Line Coupler design dimensioning; HFSS and ADS simulation of Branch Line Coupler; LibreVNA testing of Branch Line Coupler; Wireless communication system construction (text transmission); PPT production.	
Score	100