NTNU

TTT4205 Microwave Techniques

Waveguide assignment

Magne H. Å. Haneberg

November 11, 2015

1 Rectangular waveguide

1.1 Cut-off frequencies

The cut-off frequency of a hollow waveguide is given by

$$f_{\mathrm{c}}^{(\mathrm{mn})} = \frac{c}{2\pi\sqrt{\epsilon_{\mathrm{r}}\mu_{\mathrm{r}}}}\sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2}. \tag{1}$$

For a *WR75* waveguide, modes TE_{10} and TE_{20}, the cut-off frequencies are $f_{\mathrm{c}}^{(10)} = 7.8740\,\mathrm{GHz}$ and $f_{\mathrm{c}}^{(20)} = 15.748\,\mathrm{GHz}$.

The unimodal frequency band, ΔF, is given by

$$\Delta F = f_{\mathrm{c}}^{(20)} - f_{\mathrm{c}}^{(10)} = 7.8740\,\mathrm{GHz}. \tag{2}$$

1.2 Normalized propagation constants

The propagation constant of a hollow waveguide is given by

$$k_{\mathrm{z}}^{(\mathrm{mn})} = \sqrt{k_0^2\epsilon_{\mathrm{r}}\mu_{\mathrm{r}} - \left(\frac{m\pi}{a}\right)^2 - \left(\frac{n\pi}{b}\right)^2} \tag{3}$$

where

$$k_0 = \frac{2\pi f}{c} = \frac{2\pi f}{3\times 10^8\,\mathrm{m/s}}. \tag{4}$$

For a *WR75* waveguide, modes TE_{10} and TE_{20}, from 5 to 25 GHz, the propagation constants are shown in figure 1.

1.3 Phase and group velocities

The phase and group velocities for the TE_{10} mode in the unimodal frequency band is given by equations 5 and 6 and plotted in figure 2.

$$V_{\mathrm{ph}}^{(\mathrm{mn})} = \frac{\frac{c}{\sqrt{\epsilon_{\mathrm{r}}\mu_{\mathrm{r}}}}}{\sqrt{1 - \left(\frac{f_{\mathrm{c}}^{(\mathrm{mn})}}{f}\right)^2}} \tag{5}$$

$$V_{\mathrm{g}}^{(\mathrm{mn})} = \frac{c}{\sqrt{\epsilon_{\mathrm{r}}\mu_{\mathrm{r}}}}\sqrt{1 - \left(\frac{f_{\mathrm{c}}^{(\mathrm{mn})}}{f}\right)^2} \tag{6}$$

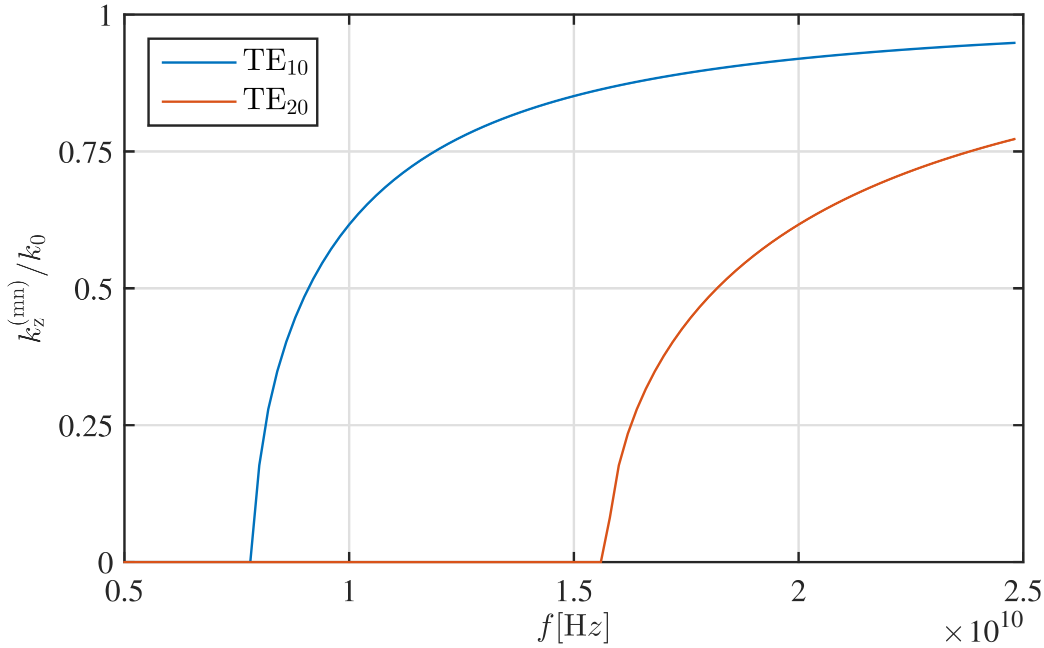

Figure 1: Normalized propagation constants in TE$_{10}$ and TE$_{20}$ modes.

1.4 Wave impedance

The wave impedance of a TE$_{\mathrm{mn}}$ mode is given by equation 7. This impedance is plotted for a TE$_{10}$ wave in figure 3 for the unimodal frequency band.

$$W_{\mathrm{mn}}^{\mathrm{TE}} = \frac{\dot{E}_{\mathrm{x}}^{(\mathrm{mn})}}{\dot{H}_{\mathrm{y}}^{(\mathrm{mn})}} = \frac{\omega\mu_{\mathrm{r}}\mu_0}{k_{\mathrm{z}}^{(\mathrm{mn})}} = \sqrt{\frac{\mu_{\mathrm{r}}\mu_0}{\epsilon_{\mathrm{r}}\epsilon_0}}\frac{1}{\sqrt{1-\left(\frac{f_{\mathrm{c}}^{(\mathrm{mn})}}{f}\right)^2}} \tag{7}$$

1.5 Cut-off frequencies in an ideal dielectric

The cut-off frequencies for a waveguide is given by equation 1. If the waveguide is filled with air, this corresponds to $\epsilon_{\mathrm{r}} = 1$. When filled with a perfect dielectric, ϵ_{r} changes value. However, it is still a real number.

The cut-off frequencies of a *WR75* waveguide filled with a perfect dielectric, $\epsilon_{\mathrm{r}} = 2.44$, are shown in table 1.

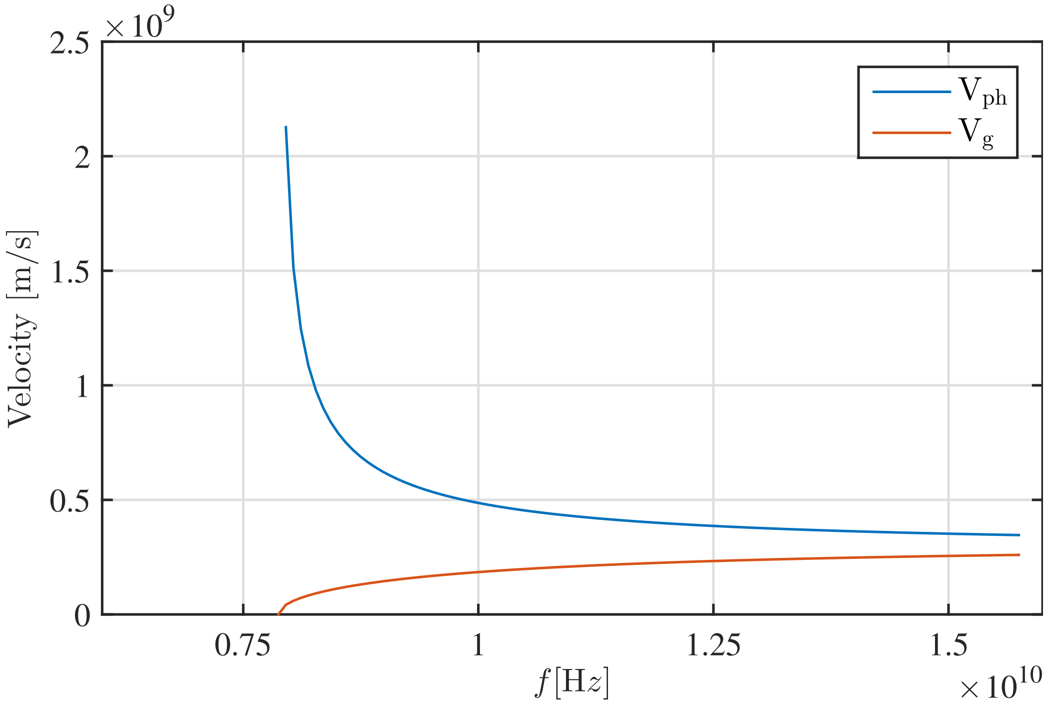

Figure 2: Phase and group velocities of TE_{10} wave in the unimodal frequency band

Table 1: Cut-off frequencies for waveguide filled with perfect dielectric

Mode	Cut-off frequency
TE_{10}	5.0408 GHz
TE_{20}	10.082 GHz

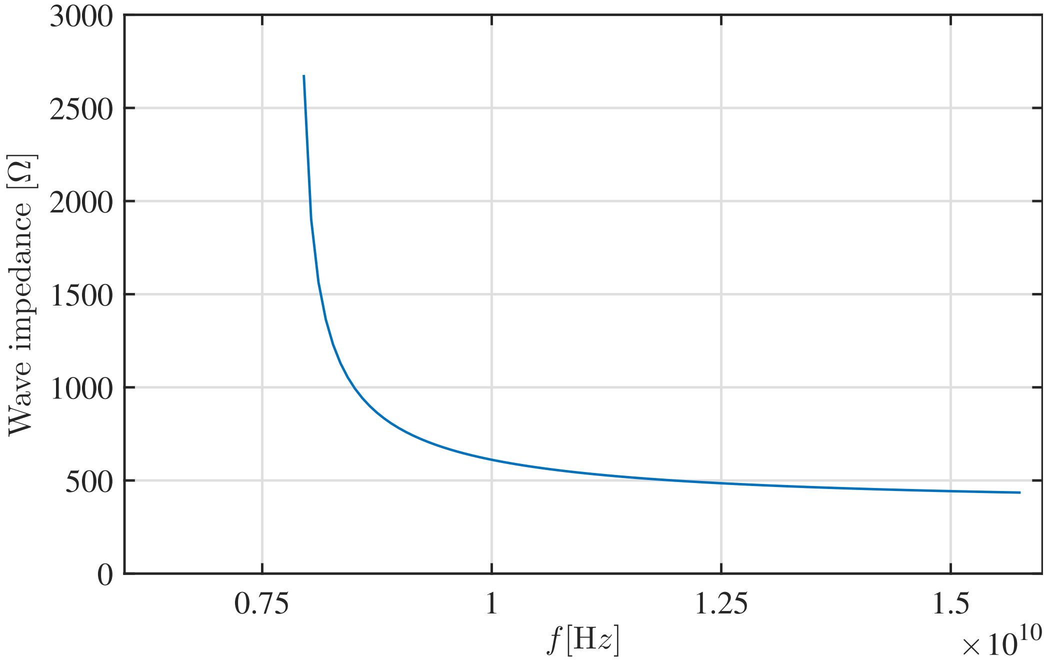

Figure 3: Wave impedance of TE_{10} wave in the unimodal frequency band

1.6 Dielectric loss constant

The dielectric loss constant for a mode TE_{mn} is given by equation 8.

$$\alpha_{\mathrm{d}}^{(\mathrm{mn})} = \frac{\Delta \bar{P}_{\mathrm{loss/unit}}}{2\bar{P}} \approx \frac{\epsilon_{\mathrm{r}}''}{\epsilon_{\mathrm{r}}'} \frac{\pi}{\lambda} \left(\frac{\Lambda_{\mathrm{mn}}}{\lambda} \right) \tag{8}$$

Λ_{mn} and λ are given by equations 9 and 10.

$$\Lambda_{\mathrm{mn}} = \frac{2\pi}{k_{\mathrm{z}}^{(\mathrm{mn})}} = \frac{\lambda}{\sqrt{1 - \left(\frac{f_{\mathrm{c}}^{(\mathrm{mn})}}{f} \right)^2}} \tag{9}$$

$$\lambda = \frac{2\pi}{k_0 \sqrt{\epsilon_{\mathrm{r}} \mu_{\mathrm{r}}}}. \tag{10}$$

According to [1],

$$\epsilon'' = \epsilon' \tan\delta = \epsilon_{\mathrm{r}} \epsilon_0 \tan\delta. \tag{11}$$

Given $\tan\delta = 0.002$, we have

$$\alpha_{\mathrm{d}}^{(\mathrm{mn})} \approx \frac{k_0 \sqrt{\epsilon_{\mathrm{r}} \mu_{\mathrm{r}}}}{1000 \sqrt{1 - \left(\frac{f_{\mathrm{c}}^{(\mathrm{mn})}}{f} \right)^2}}. \tag{12}$$

The dielectric loss constant is plotted for the unimodal frequency band in figure 4.

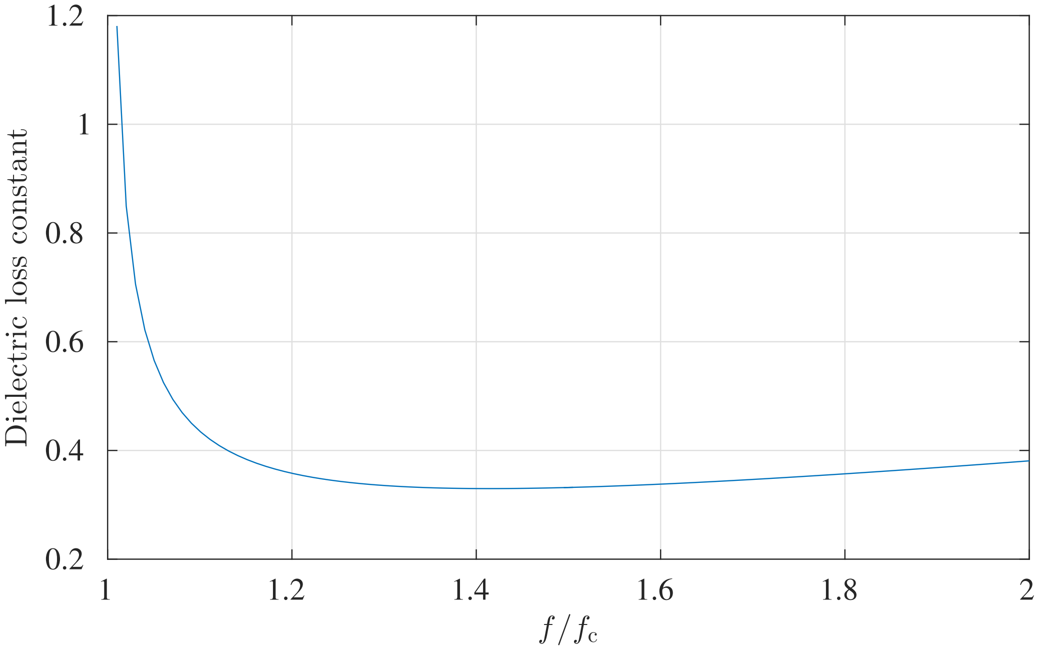

Figure 4: Dielectric loss constant, TE_{10} wave.

1.7 Equivalent capacitance and inductance

The equivalent capacitance and equivalent inductance are given by equations 13 and 14, respectively.

$$C_{\mathrm{equ}} \approx \frac{2k_{\mathrm{z}}^{(10)}b}{\pi^2 f W_{10}^{\mathrm{TE}}} \ln\left[\csc\left(\frac{\pi s}{2a}\right)\right] \tag{13}$$

$$L_{\mathrm{equ}} \approx \frac{a\mu_{\mathrm{r}}\mu_0}{2\pi\cot^2\left(\frac{\pi s}{2a}\left[1+\csc^2\left(\frac{\pi s}{2a}\right)\right]\right)} \tag{14}$$

The equivalent capacitance is plotted in figure 5.

The equivalent inductance does not vary with frequency. For the specified waveguide, $L_{\mathrm{equ}} \approx 2.1633 \times 10^{-8}$ H/m.

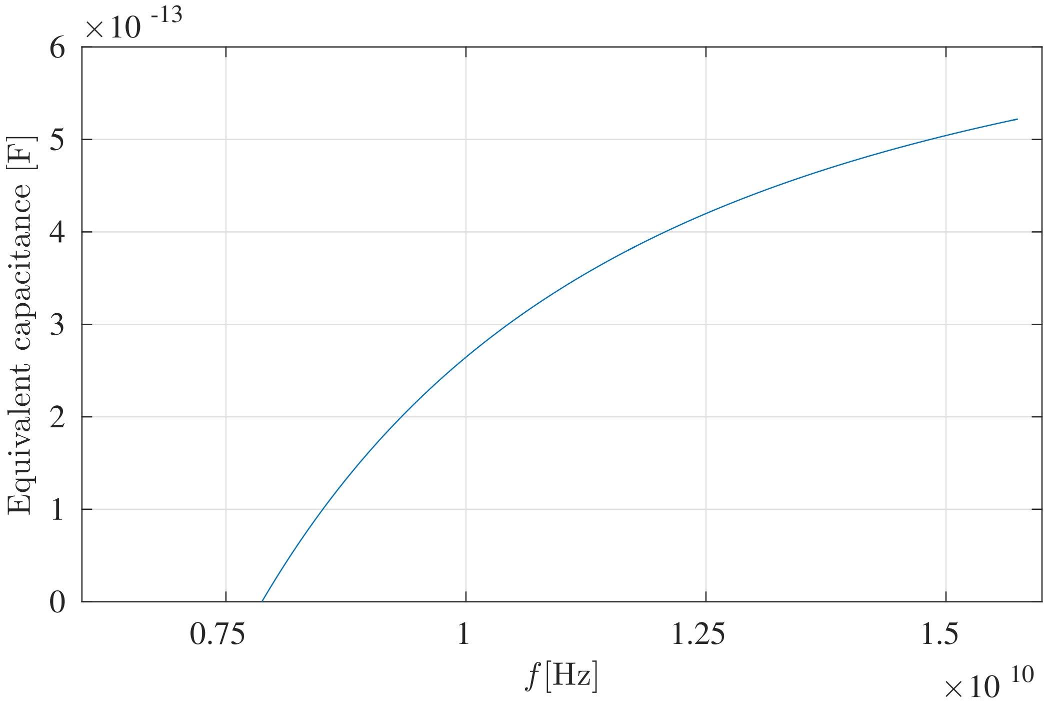

Figure 5: Equivalent capacitance over the unimodal frequency band.

2 Microstrip line

2.1 Effective permittivity

The effective permittivity relies on the geometry of the microstrip line. It is given by equation 15.

$$\epsilon_{\mathrm{eff}}(f=0) = \left(\frac{\beta}{k_0}\right)^2 = \frac{\epsilon_{\mathrm{r}}+1}{2} + \frac{\epsilon_{\mathrm{r}}-1}{2\sqrt{1+\frac{10h}{W}}} \tag{15}$$

Given the geometry $h = 1\,\mathrm{mm}$ and $W = (0.1\ldots4)\,\mathrm{mm}$, with $\epsilon_{\mathrm{r}} = 3.32$, the effective permittivity becomes as plotted in figure 6.

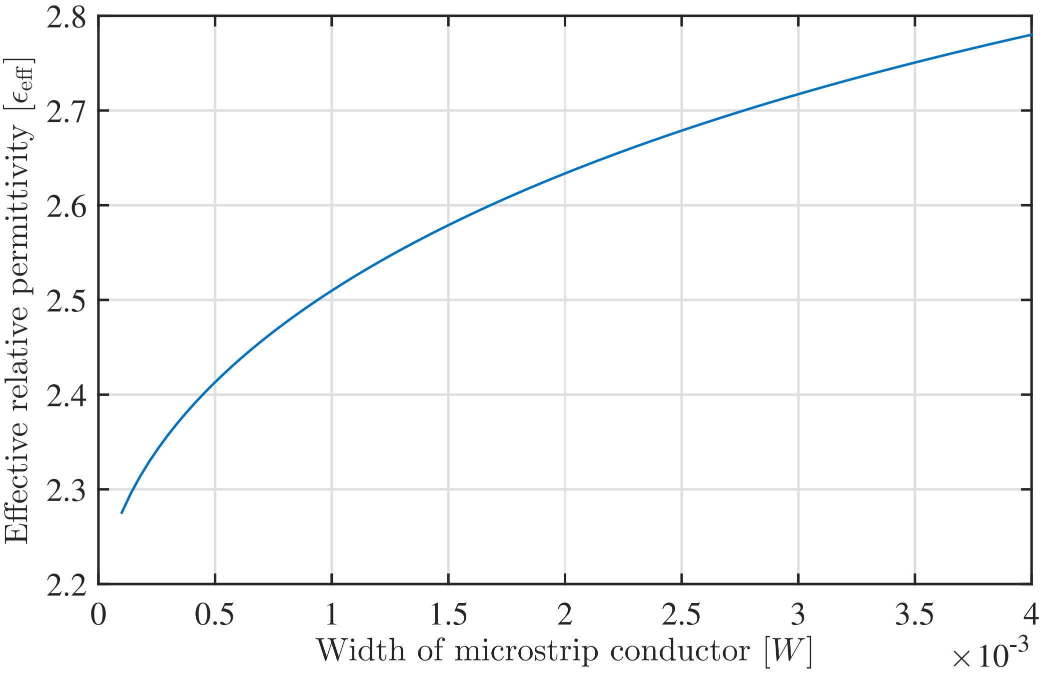

Figure 6: Effective relative permittivity of microstrip line.

2.2 Characteristic impedance

The characteristic impedance of a microstrip line is given by equation 16.

$$Z_{\mathrm{c}} = \begin{cases} \frac{60}{\epsilon_{\mathrm{eff}}(f=0)} \ln\left(\frac{8h}{W} + \frac{0.25W}{h}\right) & \frac{W}{h} < 1 \\ \frac{120\pi}{\sqrt{\epsilon_{\mathrm{eff}}(f=0)}\left[\frac{W}{h}+1.393+0.667\ln\left(\frac{W}{h}+1.444\right)\right]} & \frac{W}{h} \geq 1 \end{cases} \tag{16}$$

Given the geometry $h = 1\,\mathrm{mm}$ and $W = (0.1 \ldots 4)\,\mathrm{mm}$, with $\epsilon_{\mathrm{r}} = 3.32$, the characteristic impedance becomes as plotted in figure 7.

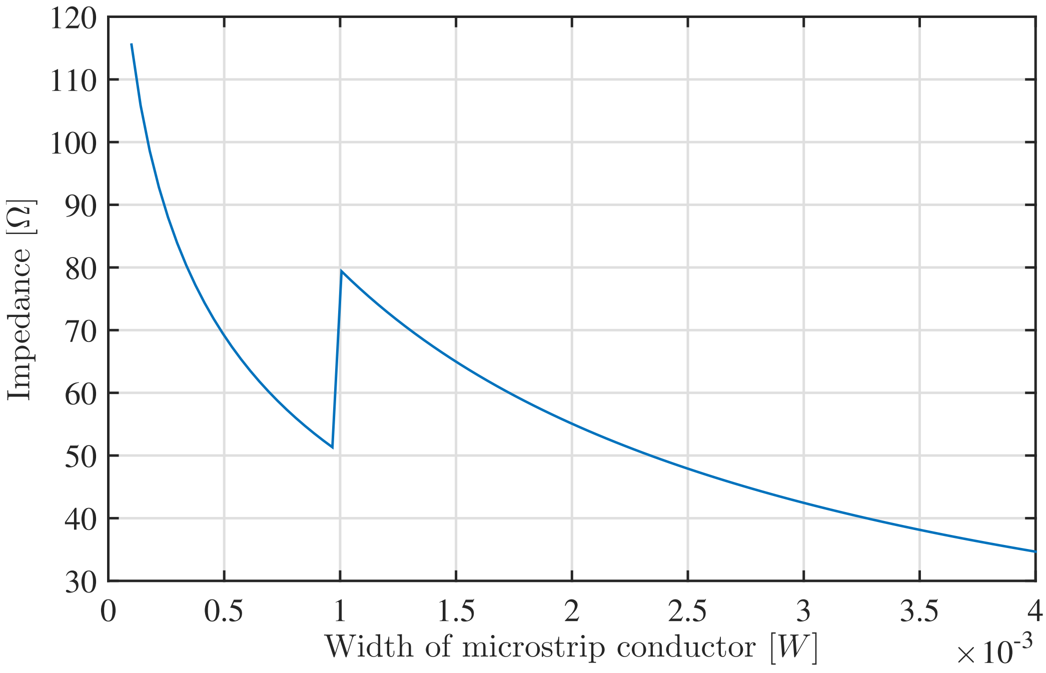

Figure 7: Characteristic impedance of microstrip line.

2.3 S-matrix coefficients

The S-matrix coefficients, transmission and reflection, are given by equations 17 and 18. They are plotted in figure 8.

$$T = \frac{2\sqrt{Z_{\mathrm{c}}^{(2)} Z_{\mathrm{c}}^{(1)}}}{Z_{\mathrm{c}}^{(2)} + Z_{\mathrm{c}}^{(1)}} \tag{17}$$

$$R = \frac{Z_{\mathrm{c}}^{(2)} - Z_{\mathrm{c}}^{(1)}}{Z_{\mathrm{c}}^{(2)} + Z_{\mathrm{c}}^{(1)}} \tag{18}$$

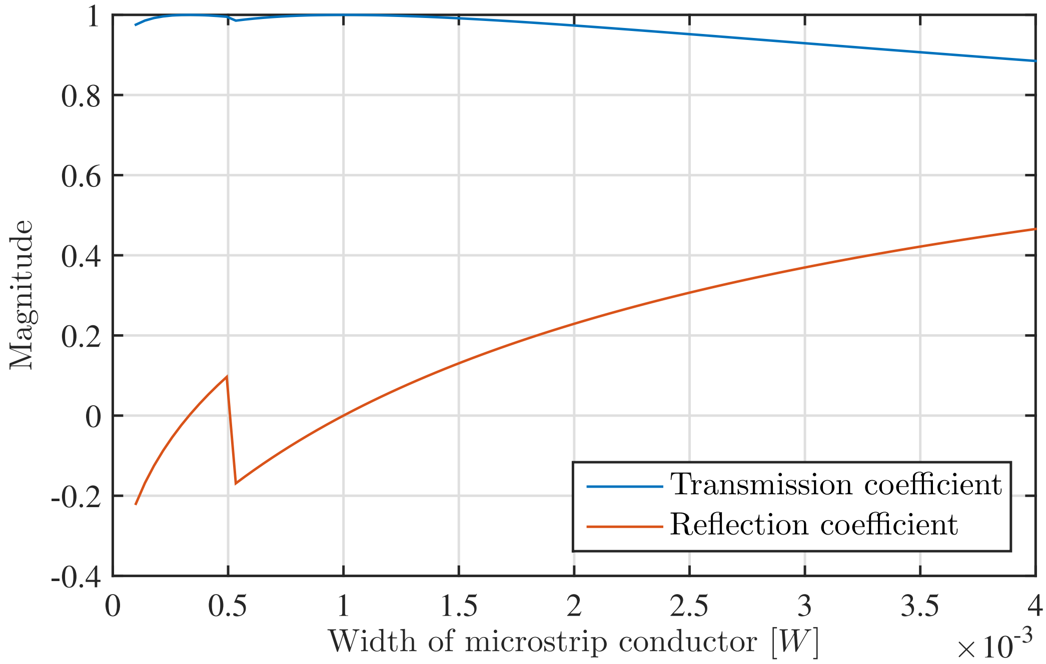

Figure 8: S-matrix coefficients of microstrip line.

3 Microstrip Branch-Line Directional Coupler

3.1 Calculating geometry of MS-BLDC

We want to design a Microstrip branch-line directional coupler with coupling and directivity ratios $-3\,\mathrm{dB}$ at the center frequency. The isolation ratio should be as low as possible.

The *Keysight ADS tools* can be used to calculate the geometry of a microstrip branch-line directional coupler. In order to use the ADS LineCalc utility, we have to obtain some parameters beforehand. These parameters are given by equations 19 and 20.

$$Z_{0,1} = Z_0\sqrt{1-k_c^2} \tag{19}$$

$$Z_{0,2} = \frac{Z_{0,1}}{\sqrt{1-\left(\frac{Z_{0,1}}{Z_0}\right)^2}} \tag{20}$$

Here k_c is given by

$$k_c = \sqrt{\frac{P_c}{P_i}} = \sqrt{\frac{1}{2}} \tag{21}$$

where P_i is the incident power and P_c is the power in the coupled port. We also have to take into account that port 1, port 2 and port 4 are all a quarter wavelength away from each other. Therefore, ideally, the power on port 2 and 4 should be 90° out of phase. We shall examine this closer later.

Inserting the values in table 2 into ADS LineCalc, the results in table 3 are obtained.

Table 2: ADS LineCalc input parameters

Parameter	Size
F_0	1.6 GHz
h	1 mm
ϵ_r	3.32
Z_0	50 Ω
$Z_{0,1}$	35.3553 Ω
$Z_{0,2}$	50.0000 Ω
$\Delta\phi$	90°

Table 3: Branch-line coupler geometry

Parameter	Size
L50	28.8218 mm
W50	2.337 17 mm
L1	28.2112 mm
W1	3.910 89 mm
L2	28.8170 mm
W2	2.346 83 mm

3.2 Simulation of MS-BLDC

Figure 9 shows a model of the MS-BLDC in ADS.

Figure 10 shows a simulation of S(2,1) and S(4,1) in ADS. The markers indicate the peak values. We see that the components are in fact not exactly

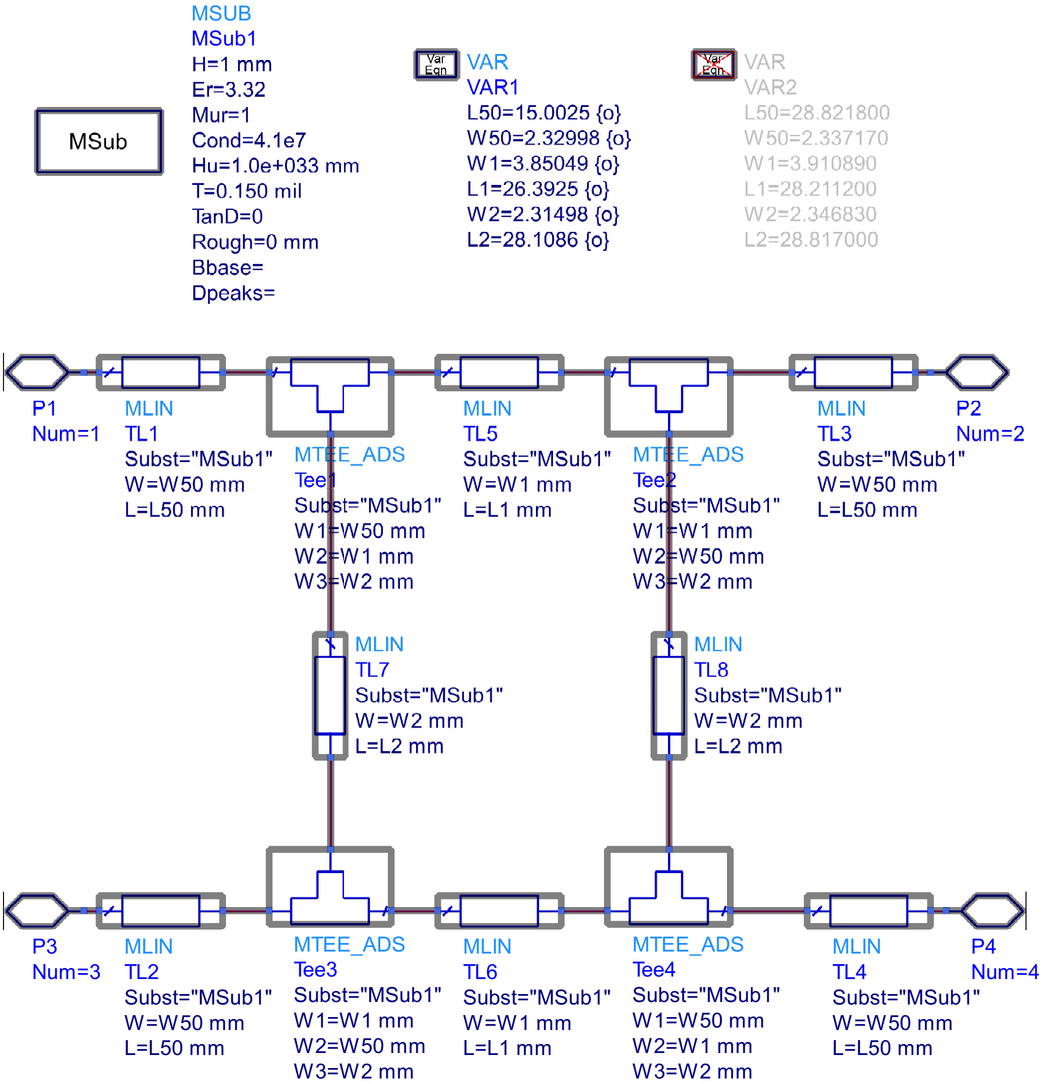

Figure 9: Model of Microstrip branch-line directional coupler in ADS.

−3 dB at 1.6 GHz. As can be seen from figure 11, the two components are approximately 90° out of phase.

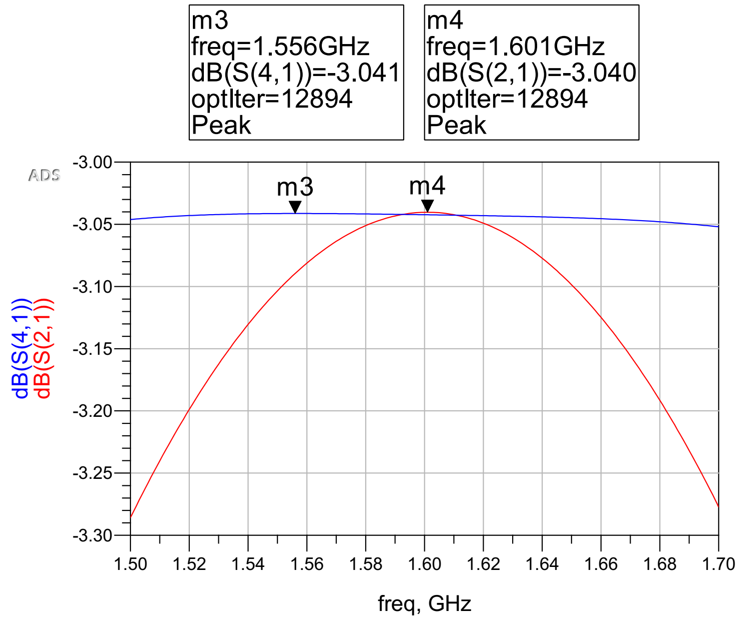

Figure 10: Simulated S-matrix components S(2,1) and S(4,1).

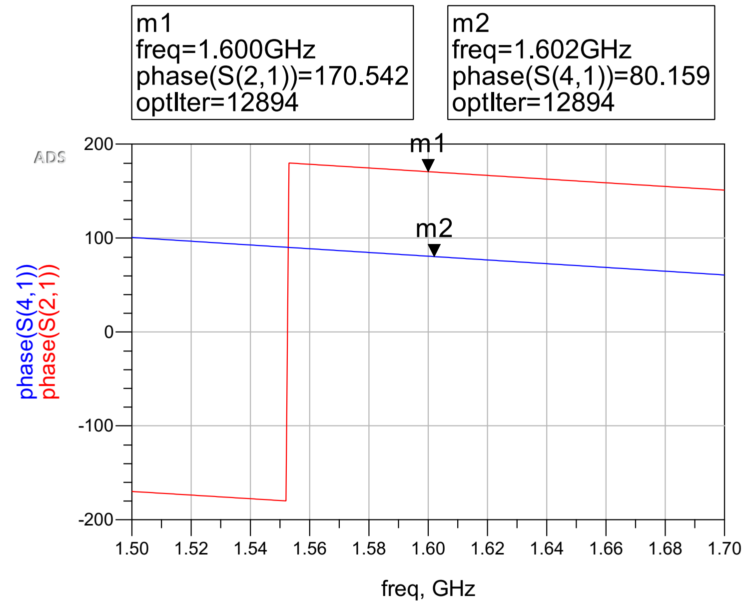

Figure 11: Simulated S(2,1) and S(4,1) phase difference.

References

[1] larebok mikrobolge