Lab RF

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Contents

1 Prerequisites: Installation of QUCS on host

Downloading the tarballs for QUCS and for the Verilog-AMS to SPICE-translator ADMS also cloning QUCS and ADMS from github in other folders and working on the installation from multiple folders resulted in many problems related to build process of `qucs-core`.

Giving up on the tarball installation folder, I went to the git-clone folder and worked until I gave up and went back again to the tar-ball folder then gave up and went so back and forth. Errors from not following installation instructions correctly, errors from uncomplete instructions which assumes a pre-knowledge in operating system internals (when super-user mode is supposed be used and when not, the necessity of the setting of enivroment varibles related to `LDPATH`).

Errouneously mixing installation commands for ADMS and for Qucs but also perhaps because of corruption of tar-balls due to limited and varying internet bandwith. Read on unix.stackexchange that a certain system administrator often encountered corrupted tarballs. Unfourtunately the reference to particular thread is lost but it does not seem unlikely to me given the fact that the `tar` command is old and might not have the same level of sofisticated error correction code as in modern git code, but I am speaking as layman I don't really know.

I also gave up on trying to get the Linux installation to work and tried running the Windows version "Qucs studio" on the Wine-platform for Linux but noticed differences in the display compared to what was shown in the document "Getting started with Qucs" and with Wine having the rumor of being buggy I did not dare to proceed. Besides that, I could not accept defeat from a "simple" installation. Done this before.

To be honest I don't know for sure why the installation suddenly worked. Did I finally follow the instructions or was it that the tarballs were corrupted or was it the enivroment variable or a combination of all these factors? Cloning the github repositories did not give the same errors which could be attributed to an possible existence of a check-sum verfification agorithm in git which possibly is not existing in the `tar` unpacking command or it could be the case that the build commands are different with respect to the git clone installation and therefore caught more attention and care.

Posted on Sourceforge[1] about the build related problems which was ignored because it contained a glaring mistake about the paths provided where I mixed up `/usr/local/bin` and `/usr/local/lib` and also posted another question on unix.stackexchange [2]. related to the Autotools build-tools suite.

According to the ADMS section of QUCS on Github [3] one is supposed to do the following when installing from a tar-ball (which is different from when installing from the github clone)

```
tar xvfz adms-x.x.x.tar.gz
cd adms-x.x.x
./configure --prefix=[/install/location/]
make install
```

The command `make install` does not work if one is not issuing it as a superuser, which is implied for those who know but is nowhere written in the instructions so under Linux Ubuntu one should issue `sudo make install`. It might be obvious to use super-user mode for someone with who is a complete novice but this is not always true. In some instances when installing in super-user mode the program is not availible in normal user-mode and one has to issue a `chmod` command to allow non super-users access.

The descision was for some reason to install ADMS in the location `/usr/local/lib` from reading on peoples problems who had succeeded with the installation on the net.

[1] https://sourceforge.net/p/qucs/discussion/311049/thread/88cc152f64/?limit=25
[2] https://unix.stackexchange.com/questions/715523/what-does-it-mean-when-aclocal-cannot-open-version
[3] https://github.com/Qucs/ADMS/blob/develop/README.md

sudo make install returned stating that the library admsXml could not be found which prompted me to run the configure portion of QUCS as `./configure --with-mkadms =/usr/local/lib/admsXml` but was still not found or not working.
It is also possible that I out of being carless have run `./configure --with-mkadms= /usr/local/bin/admsXml` which I stated on the Sourceforge site thus making a fool of myself. The tip using `--with-mkadms` was not from the official documentation but found on Stackexchange[4].

The ADMS build report which is dispalyed on the terminal during the installation also alludes to that it in some occations might be necessary to point to `LIBDIR` but not knowing what `LIBDIR` was meant to be, I decided to ignore until I realized that this might be crucial point. Found this discussion on Stackexchange[5] relating to how to set enivroment varibles in Linux. I did not do it the official sanctioned Ubuntu way which according to a writer on Stackexchange is the adding of a `.conf` file to `/etc/ld.so.conf.d` and then in the `.conf` file write the path to the ADMS-library because it was not obvious where the `.conf` file then should be stored, so I resorted to editing `.bash_profile` of which I had a vague memory of having done some years ago and it worked then, but not the recommended way according to what someone said on Stackexchange... Using the rather difficult built in vi-editor `vi ~/.bash_profile` the addition was

```
LD_LIBRARY_PATH=/usr/local/lib
export LD_LIBRARY_PATH
```

The `.bash_profile` which I had edited before was empty when opened and information being lost from there after restart is discussed on mentioned thread but the work-around suggested results in an error if tried. After editing `.bash_profile` one is supposed to issue (at the terminal):

```
source ~/.bash_profile
sudo ldconfig
```

Finally, I still could not build the documentation because QUCS uses now obsolete libraries related to TexLive which are `texlive-math-extra`, and`pgf`. The `pgf` library is nowhere to be found because it is an instrinsic part of the standard TexLive distribution since years ago. Luckily QUCS provide the option to not build the documentation using the switch `--disable-doc` in the top-level configure script. QUCS also uses an obsololete, at least with respect to Ubuntu, Octave package for converting encapsulated postscript files `octave-epstk` which is nowhere to be found except for Debian Linux it seems.

Being still at the novice level in installing from source code and not having completed the full verfication of the installation (`make check`) before throwing out the code folder, the installation is not to be trusted. Dr. Dancila helped me with some tests in a Zoom-session but after the session ended disturbing warnings were detected which were not brought to Dr. Dancila's attention.

2 Verification of the installation with respect to RF-theory and with respect to performance as stated in the QUCS manual

Doing so many mistakes during the installation there can be no trust that something was not broken or corrupted working with both the tarball folder and the git cloned folder for both ADMS and QUCS. Due to the mistakes during the installation and the many problems encounterd also because of the fact that full built in checks of the installation were not performed before deleting the code folder, it feels necessary to evaluate the installation with respect to theory and record warnings and errors for evaluation before proceeding with the lab so to have some level of confidence that the QUCS installation will actually perform as the authors designed it. Especially concerning was the

[4] https://stackoverflow.com/questions/36102809/qucs-core-configure-error-needs-admsxml
[5] https://stackoverflow.com/questions/13428910/how-to-set-the-environmental-variable-ld-library-path-in-linux

following meassage after completing a zoom-session with Dr. Dancila and which was not shown to him.

```
Errors and Warnings:
--------------------

line 33: no trailing end-of-line found, continuing...
```

Does this impact the accuracy of the calculations? I certainly would not bet my life on that it does not!
Futher more we also want to verify that simple processes which will be used in the lab was not corrupted such that reading the touchstone file provided by Dr. Dancila and some of the built in functions which will be used in the lab before actually proceeding to the lab task. We also repeat the tests done together with the Dr. Dancila for full reference.

The following tests feels suitable to perform at minimum before proceeding with the actual lab-work.

(a) Reflection coefficient of load without transmission line should be transformed to the correct position in the Smith-chart.

(b) Calculation of Return Loss and Mismatch Loss should conform with Matlab/Octave.

(c) Reflection coefficient of load with 50 ohm transmission line should be transformed to correct position in the Smith chart

(d) Sweeping the length of the transmission line one half wave length shoul result in a plot of the reflection coefficient describing a full circle.

(e) Example 9-13 in David K. Cheng "'Field and wave Electromagnetics"' short circuited line is transformed to a certain impedance with a transmission line of 0.1λ.

(f) Plot of S-parameters verifying that the particular part of the touchstone file is read correctly.

(g) Plot of the built in function $Rolett()$ to verify that it accurately performs the calculation of the Rolett stability condition.

(h) Plot of built-in functions $stabL()$ and $stabS()$ verifying input- and output-stability circles with respect to theory.

(i) Plot of the Noise-figure data verifying that the particular part of the Touchstone file is read correctly.

The Toucstone file is as follows

```
! Filename:       BFG520I.S2P       Version:   2.1
! NXP part #: BFG520                                     Date: Feb 1992
! Bias condition: Vce=6V, Ic=10mA
! IN LINE PINNING: same data as with cross emitter pinning.
#  MHz  S  MA  R  50
! Freq        S11            S21             S12             S22      !GUM [dB]
   40     .711  -14.1  23.473  170.8    .007   82.3    .974   -6.8 !    43.5
  100     .690  -34.5  22.218  158.5    .016   74.0    .931  -16.3 !    38.5
  200     .640  -64.3  19.183  141.5    .029   62.4    .816  -28.8 !    32.7
  300     .597  -88.3  16.207  128.8    .037   54.2    .703  -37.1 !    29.1
  400     .569 -106.1  13.627  119.5    .042   50.9    .613  -42.4 !    26.4
  500     .553 -119.9  11.673  112.5    .046   48.6    .544  -45.7 !    24.5
  600     .538 -130.7  10.152  107.0    .050   47.2    .493  -47.6 !    22.8
  700     .526 -139.2   8.947  102.5    .052   47.1    .453  -48.8 !    21.4
  800     .514 -146.7   7.968   98.6    .055   47.9    .422  -49.5 !    20.2
  900     .503 -154.1   7.148   95.1    .057   48.8    .396  -50.2 !    19.1
```

```
 1000     .495 -160.2   6.488   92.4     .059   49.7     .374  -51.1 !     18.1
 1200     .494 -171.7   5.468   87.0     .065   51.3     .344  -53.4 !     16.5
 1400     .500 -131.8   4.748   82.0     .069   52.5     .331  -56.1 !     15.3
 1600     .500  173.3   4.159   77.8     .075   55.2     .326  -57.3 !     14.1
 1800     .494  166.0   3.722   74.3     .082   56.4     .317  -57.9 !     13.1
 2000     .496  158.4   3.368   71.0     .087   58.2     .297  -59.5 !     12.2
 2200     .515  151.9   3.080   67.9     .093   59.0     .279  -64.4 !     11.5
 2400     .535  147.6   2.798   64.2     .097   60.7     .280  -71.4 !     10.8
 2600     .540  144.2   2.579   61.6     .107   61.3     .298  -76.3 !     10.1
 2800     .534  139.8   2.428   58.2     .114   60.9     .313  -77.7 !      9.6
 3000     .541  133.9   2.259   55.3     .120   62.3     .309  -78.3 !      9.0
! Noise data:
! Freq.      Fmin        Gamma-opt         rn
  500        1.10     .330      27.0       .250
  900        1.25     .294      48.0       .260
 1000        1.30     .298      52.0       .270
 2000        1.90     .242     134.0       .160
```

2.1 Tests

(a) Reflection coefficient of load without transmission line should be transformed to the correct position in the Smith-chart.

When the dimensions of the circuit d are much less than the length of the wave-length λ the voltage depends only on their configuraion in the circuit but geometric distances from the source are of no importance, therefore sweeping the frequency when there is no difference in the voltage along the line should result in a constant reflection coefficient for all frequencies.

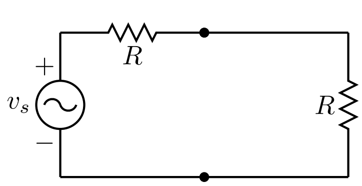


Figure 1: When $d << \lambda$ the dimensions does not impact any behaviour of the circuit

The formula for the reflection coefficient Γ is

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0}$$

We are using normalized impedances so the formula is

$$\begin{aligned}\Gamma &= \frac{Z_L/Z_0 - 1}{Z_L/Z_0 + 1} \\ &= \frac{z_L - 1}{z_L + 1}\end{aligned}$$

We recieve

$$\begin{aligned}\Gamma &= \frac{z_L - 1}{z_L + 1} \\ &= \frac{25/50 - 1}{25/50 + 1} \\ &= -0.3333\end{aligned}$$

QUCS gives the following results

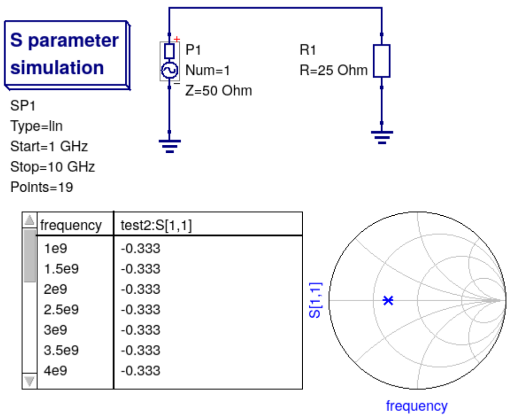


Figure 2: $R = 25$ Ohm

with no warnings or errors and thus we did not break anything during the installation here.

(b) Reflection coefficient of load with 50 ohm transmission line should be transformed to correct position in the Smith chart

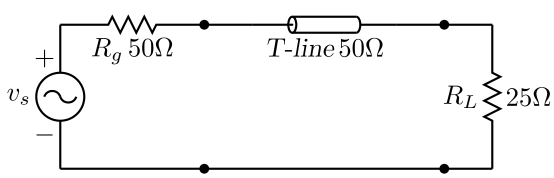


Figure 3: When $d << \lambda$ the dimensions does not impact any behaviour of the circuit

If we have $Z_L = 25$ Ohm and 20 mm lossless transmissionline of 50 Ohm applying a voltage of 2 GHz, the theory states that the impedance seen at distance l from the load looking towards the load is

$$Z_i = R_0 \frac{Z_L + jR_0 tan\beta l}{R_0 + jZ_L tan\beta l}$$

What is βl ? βl is supposed to be $2\pi l/\lambda$ and we arrive to this because

$$\beta = \omega\sqrt{LC} \quad u_p = \frac{1}{\sqrt{LC}}$$
$$\omega = 2\pi f \qquad u_p = f\lambda$$

So

$$\begin{aligned}\beta l &= \omega\sqrt{LC}\, l = 2\pi f\sqrt{LC}\, l \\ &= 2\pi\frac{u_p}{\lambda}\sqrt{LC}\, l = 2\pi\frac{1}{\cancel{\sqrt{LC}}\lambda}\cancel{\sqrt{LC}}\, l \\ &= 2\pi l/\lambda\end{aligned}$$

We apparently need a number for βl at 2 GHz where $l = 1$mm, the wavelength l is

$$\lambda = \frac{u_p}{f} = \frac{3\cdot 10^8}{2\cdot 10^9} = 0.15 \text{ m} = 150 \text{ mm}$$

then l/λ is

$$\frac{l}{\lambda} = \frac{20}{150} = 0.13333$$

so

$$\begin{aligned}Z_i &= R_0\frac{Z_L + jR_0 tan\beta l}{R_0 + jZ_L tan\beta l} \\ &= 50\frac{25 + j50tan0.133333}{50 + j25tan0.13333} \\ &= 42.677 + 31.832i\end{aligned}$$

Normalized to the transmissionline impedance Z_i becomes z_i as

$$\begin{aligned}z_i &= Z_i/50 \\ &= \frac{42.677 + 31.832i}{50} \\ &= 0.8535 + 0.6366i\end{aligned}$$

To find out which reflection coefficient this represents we could either calculate Γ at the load and realize that $\Gamma = \Gamma(z' = 20mm)$ is Γ measured at the load multiplied with a phase factor e^{θ_γ} corresponding to 20mm or we can simply calculate

$$\Gamma_i = \frac{z_i - 1}{z_i + 1}$$

We will do both starting with the last stated equation

$$\begin{aligned}\Gamma_i &= \frac{z_i - 1}{z_i + 1} \\ &= \frac{0.8535 + 0.6366i - 1}{0.8535 + 0.6366i + 1} \\ &= 0.034843 + 0.331507i \\ &= 0.33\angle\underline{84}^\circ\end{aligned}$$

We also use We will do both starting with the last stated equation

$$
\begin{aligned}
\Gamma_i &= \Gamma_L e^{-2\gamma z'} \\
&= \Gamma_L e^{-j2\beta z'} \\
&= \Gamma_L e^{-j2\frac{2\pi z'}{\lambda}} \\
&= \Gamma_L e^{-j2\frac{2\pi 20}{150}} \\
&= 0.034843 + 0.331507i \qquad = 0.33\underline{/84}^{\circ}
\end{aligned}
$$

which is the same result as the first method of calculation. The last calculations were obtained using the followong Ocatve script

```
l=0.020
c = 3E+8
f=2E+9
lambda = c/f
beta=2*pi/lambda
R0=50;
ZL=25;

Zi=R0*((ZL+i*R0*tan(beta*l))...
   /(R0+i*ZL*tan(beta*l)))

zi=Zi/R0

Gamma_i = (zi-1)/(zi+1)

[THETA, R] = cart2pol...
  (real(Gamma_i), imag(Gamma_i))

THETA_GRAD=THETA*180/pi

zL=ZL/R0

Gamma_L = (zL-1)/(zL+1)
Gamma_i2 = Gamma_L*exp(-i*2*beta*l)
```

The QUCS installation also gives the same results. We also notice that the point seems to be correct located on the Smith Chart which is almost straight above the centerpoint of the Smithchart $(1, 0)$.

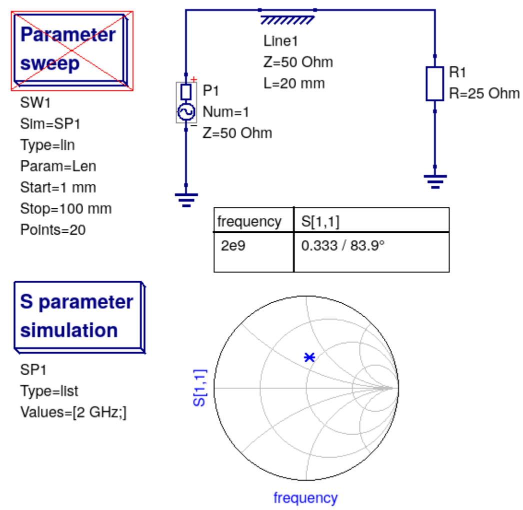
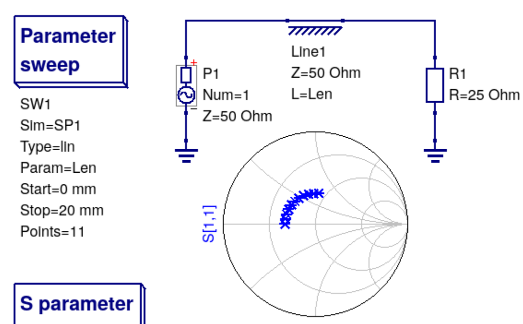


Figure 4: $R = 25$ Ohm, 20 mm 50 Ohm transmissionline

It is easier to verify the correctness of the point if the length of the line is swept from 0 to 20 mm so that ones sees on the Smithchart how the reflection coefficient changes when the length of the transmissionline varies. We expect to see a clockwise arc.



Len	frequency	S[1,1]
0	2e9	-0.333
0.002	2e9	0.333 / 170°
0.004	2e9	0.333 / 161°
0.006	2e9	0.333 / 151°
0.008	2e9	0.333 / 142°
0.01	2e9	0.333 / 132°
0.012	2e9	0.333 / 122°
0.014	2e9	0.333 / 113°
0.016	2e9	0.333 / 103°
0.018	2e9	0.333 / 93.5°
0.02	2e9	0.333 / 83.9°

Figure 5: Clockwise rotation of Γ is “Towards the generator”

with no warnings or errors and thus we did not break anything during the installation here but we found the transient simulation to be broken.

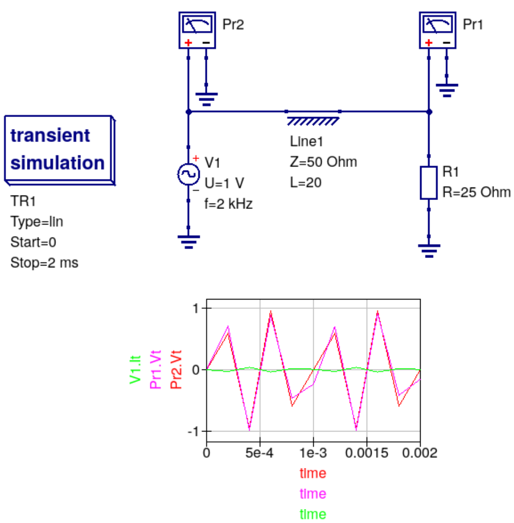


Figure 6: Transient simulation is broken

(e) Calculation of Return Loss and Mismatch Loss should conform with Matlab/Octave.

Here it looks like that QUCS is broken. The result does not conform with Matlab/Octave

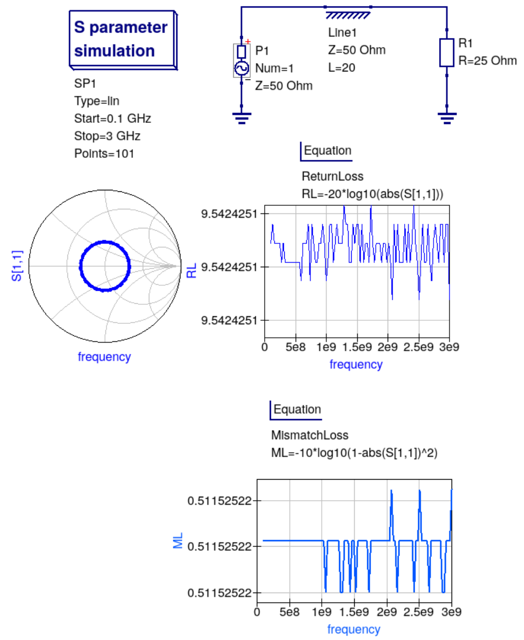


Figure 7: log-calculations are rubbish. Installation is broken

Octave gives the following Return Loss

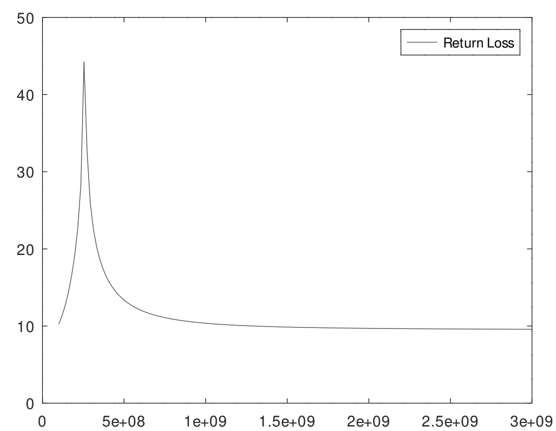


Figure 8: Return Loss

and the following Mismatch Loss

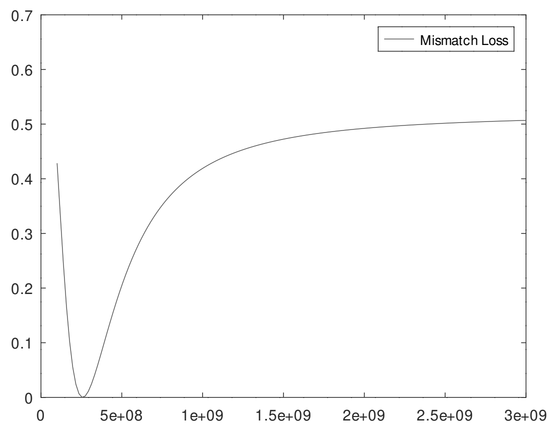


Figure 9: Mismatch Loss

From the following Octave code

```
l=0.020
c = 3E+8
f_start=0.1E+9;
f_stop=3.0E+9;
n_steps = 150;
step_size=(f_stop-f_start)/n_steps;
f=f_start:step_size:f_stop
GammaIn =zeros(size(f));

R0=50;
ZL=25;

for i=1:n_steps+1
      lambda = c/f(i)
      beta=2*pi/lambda

      Zi=R0*((ZL+i*R0*tan(beta*l))...
         /(R0+i*ZL*tan(beta*l)))

      zi=Zi/R0

      Gamma_i = (zi-1)/(zi+1);
      GammaIn(i)=Gamma_i;

 endfor

RL = -20*log10(abs(GammaIn));

ML = -10*log10(1-abs(GammaIn).^2);
figure(1)
plot(f,RL,";Return Loss;");

figure(2)
plot(f, ML, ";Mismatch Loss;")
```

(f) Sweeping the length of the transmission line one half wave length shoul result in a plot of the reflection coefficient describing a full circle.

If we apply 3 GHz we will have a wave length of 100 mm, so sweeping the length of the transmission line from 0 mm to $\lambda/2 = 50$mm must result in a full circle drawn in clockwise direction. Here we sweep to 75% of $\lambda/2$ to verify that the circle grows clockwise “towards the generator”.

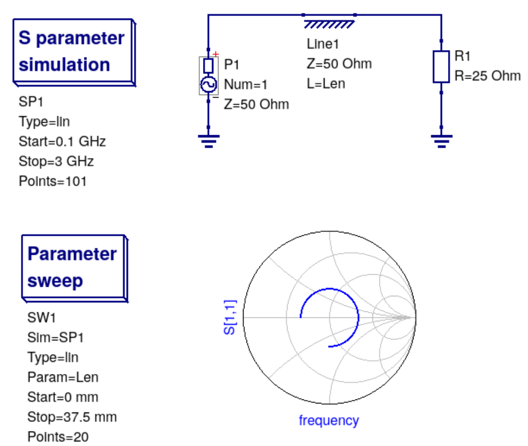


Figure 10: Sweeping the lenght of the transmission line from 0 mm to 75% of $\lambda/2$

(g) Example 9-13 in David K. Cheng “‘Field and wave Electromagnetics”’ short circuited line is transformed to a certain impedance with a transmission line of 0.1λ.

The example states - use the Smith chart to find the imput impedance of a section of 50 Ω lossless transmission line that is 0.1 wavelength long and is terminated in a short circuit.

The formula for the input impedance is

$$Z_i = R_0 \frac{Z_L + jR_0 tan\beta l}{R_0 + jZ_L tan\beta l}$$

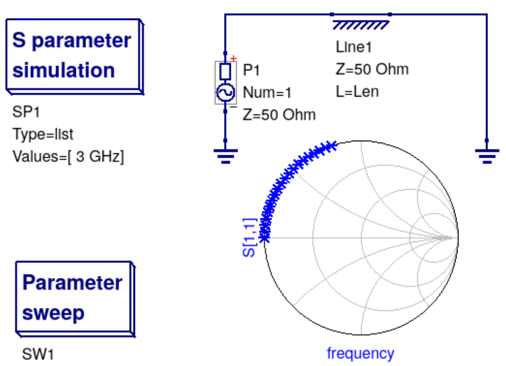
and setting $Z_L = 0$ and $l = 0.1\lambda$ gives

$$
\begin{aligned}
Z_i &= R_0 \frac{0 + jR_0 tan\beta l}{R_0 + j \cdot 0 \cdot tan\beta l} \\
&= jR_0 tan\beta l = jR_0 tan 2\pi \cdot 0.1 \\
&= 0 + 36.3271i
\end{aligned}
$$

which means that the normalized impedance $z_i = i0.726542528$ The corresponding reflection coefficient should read

$$
\begin{aligned}
\Gamma_i &= \frac{z_1 - 1}{z_1 + 1} \\
&= -0.3090 + 0.9511i
\end{aligned}
$$

Which also conforms to the QUCS calculation

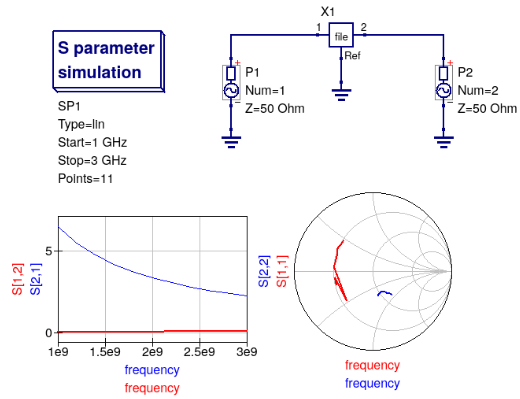


Len	frequency	S[1,1]
0.00474	3e9	-0.828+j0.561
0.00526	3e9	-0.789+j0.615
0.00579	3e9	-0.746+j0.665
0.00632	3e9	-0.701+j0.713
0.00684	3e9	-0.652+j0.758
0.00737	3e9	-0.601+j0.8
0.00789	3e9	-0.546+j0.838
0.00842	3e9	-0.49+j0.872
0.00895	3e9	-0.431+j0.902
0.00947	3e9	-0.37+j0.929
0.01	3e9	-0.308+j0.951

Figure 11: Sweeping the lenght of shortcircuited transmission line from 0 mm to 10mm at 3GHz

(h) Plot of S-parameters verifying that the particular part of the touchstone file is read correctly.

The reading is correct



Excerpt from Touhcstone for 1 GHz
Frequency in MHz/S11/S21/S12/S22
1000 .495 -160.2 6.488 92.4 .059 49.7 .374 -51.1 ! 18.1

frequency	S[1,1]	S[1,2]	S[2,1]	S[2,2]
1e9	0.495 / -160°	0.059 / 49.7°	6.49 / 92.4°	0.374 / -51.1°
1.2e9	0.494 / -172°	0.065 / 51.3°	5.47 / 87°	0.344 / -53.4°
1.4e9	0.5 / -132°	0.069 / 52.5°	4.75 / 82°	0.331 / -56.1°
1.6e9	0.5 / 173°	0.075 / 55.2°	4.16 / 77.8°	0.326 / -57.3°
1.8e9	0.494 / 166°	0.082 / 56.4°	3.72 / 74.3°	0.317 / -57.9°
2e9	0.496 / 158°	0.087 / 58.2°	3.37 / 71°	0.297 / -59.5°
2.2e9	0.515 / 152°	0.093 / 59°	3.08 / 67.9°	0.279 / -64.4°
2.4e9	0.535 / 148°	0.097 / 60.7°	2.8 / 64.2°	0.28 / -71.4°
2.6e9	0.54 / 144°	0.107 / 61.3°	2.58 / 61.6°	0.298 / -76.3°
2.8e9	0.534 / 140°	0.114 / 60.9°	2.43 / 58.2°	0.313 / -77.7°

Figure 12: S-parameter sweep with excerpt of .s2p-file at 1GHz

though errors were detected by the simulator according to the log

```
Errors occurred during simulation
on sön 04. sep 2022 at 22:19:18:426
Aborted.
```

```
Errors and Warnings:
--------------------

checker error, no actions defined: nothing to do
```

(i) Plot of the built in function $Rollet()$ to verify that it accurately performs the calculation of the Rollet stability condition.

We notice that the transistor is unstable for low frequencies $(K < 1)$ Looks the same as the one calculated in Matlab

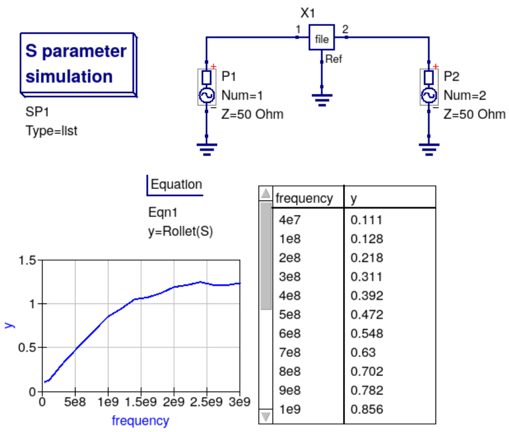


Figure 13: Rollet stability factor K

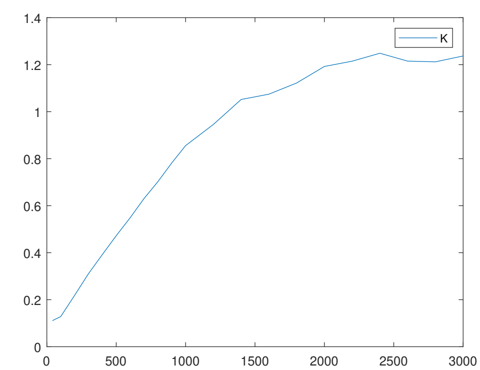


Figure 14: Rollet stability factor K calculated with Matlab

Cleand the .s2p file of the comments and the noisefigeure data and adapted code from Mathworks[6]

```
fid  = fopen('Spar.s2p', 'r') ;
 data = cell(1e6, 9) ;          % Prealloc.
 rCnt = 0 ;                     % Row counter.
 while ~feof(fid)
    rCnt = rCnt + 1 ;
    data{rCnt,1} = fscanf(fid, '%f', 1);
    data{rCnt,2} = fscanf(fid, '%f', 1);
    data{rCnt,3} = fscanf(fid, '%f', 1);
    data{rCnt,4} = fscanf(fid, '%f', 1);
    data{rCnt,5} = fscanf(fid, '%f', 1);
    data{rCnt,6} = fscanf(fid, '%f', 1);
    data{rCnt,7} = fscanf(fid, '%f', 1);
    data{rCnt,8} = fscanf(fid, '%f', 1);
    data{rCnt,9} = fscanf(fid, '%f', 1);
 end
 fclose(fid) ;
 data = data(1:rCnt,:) ;   % Truncate.
A=cell2mat(data);
f=A(:,1);
S11=zeros(size(f))
S21=zeros(size(f))
S12=zeros(size(f))
S22=zeros(size(f))
j=sqrt(-1);
for i=1:size(f)
 phi = deg2rad(A(i,3))
 phi1 = A(i,3)*pi/180
 S11(i)=A(i,2)*(cos(phi)+j*sin(phi))
 phi= deg2rad(A(i,5))
 S21(i)=A(i,4)*(cos(phi)+j*sin(phi))
 phi= deg2rad(A(i,7))
 S12(i)=A(i,6)*(cos(phi)+j*sin(phi))
 phi= deg2rad(A(i,9))
 S22(i)=A(i,8)*(cos(phi)+j*sin(phi))
end
Delta = abs(S11.*S22-S12.*S21)
K=(1-abs(S11).^2-abs(S22).^2+abs(Delta).^2)...
  ./(2*abs(S12.*S21))
plot(f,K)
legend('K')
```

(j) Plot of built-in functions *stabL*() and *stabS*() verifying input- and output-stability circles with respect to theory.

[6] https://se.mathworks.com/matlabcentral/answers/76197-how-to-read-strings-from-file-with-fscanf-or-sscanf-not-textscan

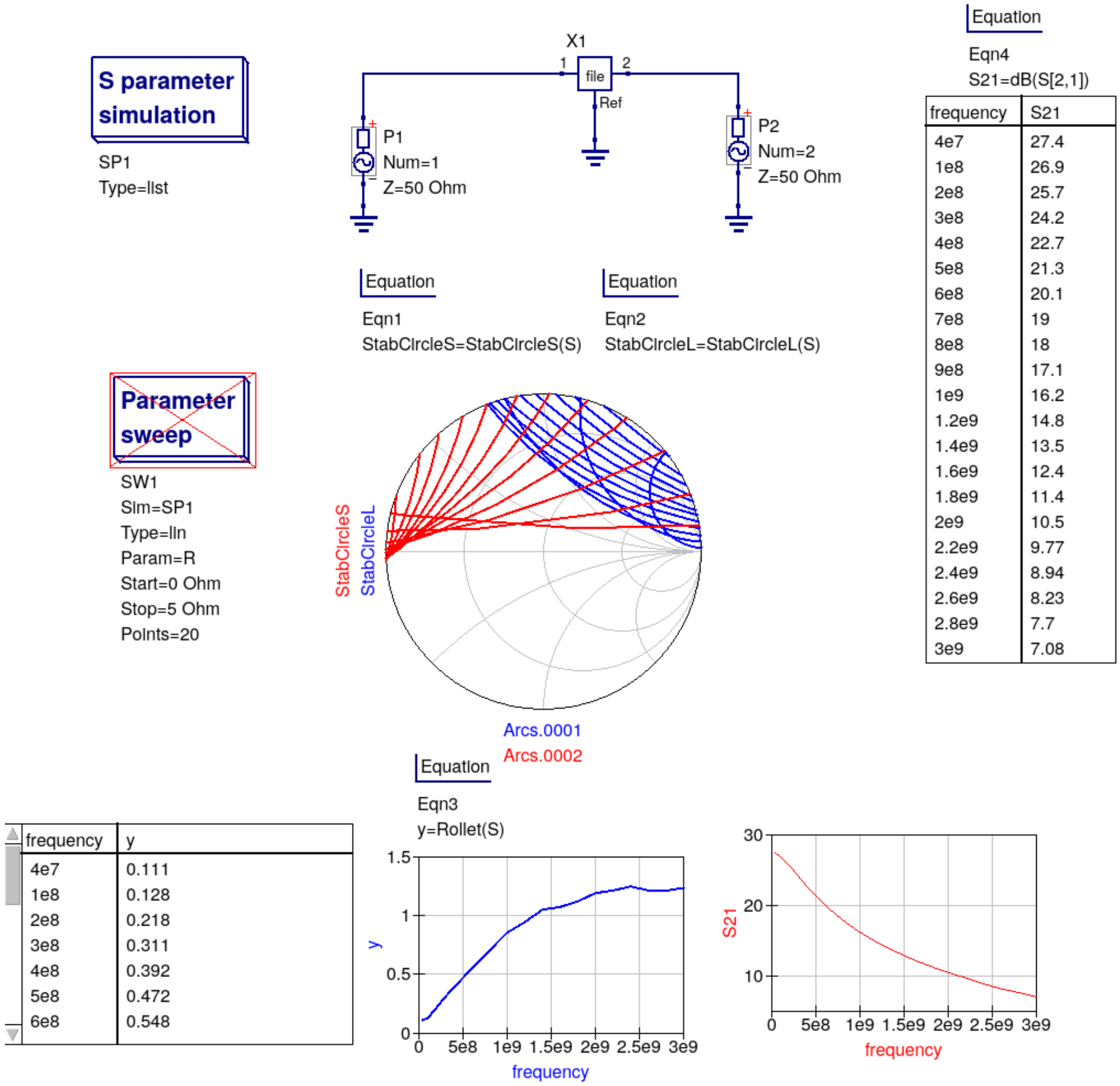


Figure 15: Input and output stability circles

3 Amplifier Theory, Chap. 12 in Pozar

In this section we will verify the equations presented in Pozar in the chapter about Microwave amplifiers. Pozar posts the following figure

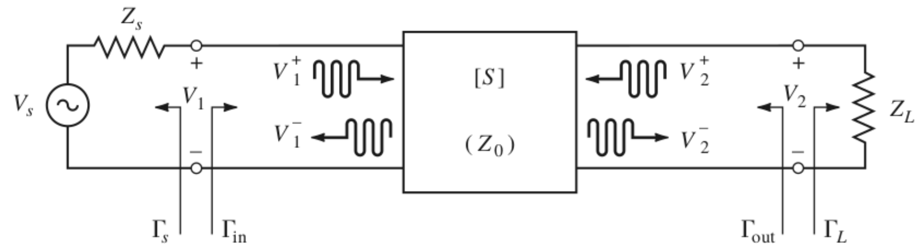


FIGURE 12.1 A two-port network with arbitrary source and load impedances.

Figure 16: Pozar figure Chap. 12

3.1 Γ_L and Γ_S

Pozar starts with the presentation of standard well known basic Electromagnetic Theory of the load reflection coefficient Γ_L

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0}$$

The source reflection coefficient looking into the load

$$\Gamma_S = \frac{Z_S - Z_0}{Z_S + Z_0}$$

Then he defines voltage scattering parameters which according to what I've seen differ between authors.

$$\begin{aligned} V_1^- &= S_{11}V_1^+ + S_{12}V_2^+ \\ V_2^- &= S_{21}V_1^+ + S_{22}V_2^+ \end{aligned}$$

This definition is apparently only valid when the source resistor and the load is equal to the characteristic impedance[7] and is in its general way defined as

$$\begin{aligned} b_1 &= S_{11}a_1 + S_{12}a_2 \\ b_2 &= S_{21}a_1 + S_{22}a_2 \end{aligned}$$

where

$$\begin{aligned} a &= \frac{1}{2}\frac{V + Z_oI}{\sqrt{|Re(Z_0)|}} \\ b &= \frac{1}{2}\frac{V + Z_o^*I}{\sqrt{|Re(Z_0)|}} \end{aligned}$$

See footnote[8]. It is unclear how one arrives at the Pozar expression of the scattering parameters from the general definition and for now will not try to show this. We think it is very unfourtunate that Pozar does not show this.

Since the load voltage coefficient Γ_L, looking into the load is the quotient between the backward travelling wave V_2^+ and the forward travelling wave V_2^-, which is

$$\Gamma_L = \frac{V_2^+}{V_2^-}$$

and the source reflection coefficient Γ_S looking into the generator also is the backward wave (to the load) divided with the forward wave (into the generator)

$$\Gamma_S = \frac{V_1^+}{V_1^-}$$

Pozar rewrites equation (1) where he replaces V_2^+ with $\Gamma_L V_2^-$

$$V_1^- = S_{11}V_1^+ + S_{12}\Gamma_L V_2^- \quad (1)$$
$$V_2^- = S_{21}V_1^+ + S_{22}\Gamma_L V_2^- \quad (2)$$

[7] https://en.wikipedia.org/wiki/Scattering_parameters
[8] https://se.mathworks.com/discovery/s-parameter.html

3.2 Γ_{in} -Reflection coefficient looking into the network

Pozar arrives at Γ_{in} the reflection coefficient looking into the network, which is the backward wave divided with the forward wave.

$$\Gamma_{in} = \frac{V_1^-}{V_1^+}$$

and finds an expression for Γ_{in} as a function of the S-parameters by solving for V_2^- in equation (1) and inserting it into equation (2) and then solving for the quotient $\frac{V_1^-}{V_1^+}$ thus obtaining a new expression for Γ_{in}

$$\begin{aligned} V_1^- &= S_{11}V_1^+ + S_{12}\Gamma_L V_2^- \iff \\ V_2^- &= \frac{V_1^- - S_{11}V_1^+}{S_{12}\Gamma_L} \end{aligned}$$

Inserting the expression of V_2^- into (2) gives

$$\begin{aligned} V_2^- &= S_{21}V_1^+ + S_{22}\Gamma_L V_2^- \qquad (2) \\ \frac{V_1^- - S_{11}V_1^+}{S_{12}\Gamma_L} &= S_{21}V_1^+ + S_{22}\Gamma_L \frac{V_1^- - S_{11}V_1^+}{S_{12}\Gamma_L} \end{aligned}$$

We collect the terms with V_1^- to the left and the terms with V_1^+ to the right but first we clean the denominator on the left hand side

$$\begin{aligned} V_1^- - S_{11}V_1^+ = S_{12}\Gamma_L S_{21}V_1^+ \\ + S_{12}\Gamma_L S_{22}\Gamma_L \frac{V_1^- - S_{11}V_1^+}{S_{12}\Gamma_L} \end{aligned}$$

We see that $S_{12}\Gamma_L$ cancels up and down on the second term of the right hand side

$$\begin{aligned} V_1^- - S_{11}V_1^+ = S_{12}\Gamma_L S_{21}V_1^+ \\ + S_{22}\Gamma_L V_1^- - S_{22}\Gamma_L S_{11}V_1^+ \end{aligned}$$

Now we are ready to collect the terms with V_1^- and V_1^+ left and right

$$\begin{aligned} V_1^- - S_{22}\Gamma_L V_1^- = S_{11}V_1^+ + S_{12}\Gamma_L S_{21}V_1^+ \\ - S_{22}\Gamma_L S_{11}V_1^+ \end{aligned}$$

We break out

$$V_1^-(1 - S_{22}\Gamma_L) = V_1^+(S_{11} + S_{12}\Gamma_L S_{21} - S_{22}\Gamma_L S_{11})$$

and solve for V_1^-/V_1^+

$$\frac{V_1^-}{V_1^+} = \frac{S_{11} + S_{12}\Gamma_L S_{21} - S_{22}\Gamma_L S_{11}}{1 - S_{22}\Gamma_L}$$

This does not look like Pozar's expression but if we break out S_{11} from the numerator we get

$$\frac{V_1^-}{V_1^+} = \frac{S_{11}(1 - S_{22}\Gamma_L) + S_{12}\Gamma_L S_{21}}{1 - S_{22}\Gamma_L}$$

and we write the right hand side as two terms

$$\frac{V_1^-}{V_1^+} = \frac{S_{11}(1 - S_{22}\Gamma_L)}{1 - S_{22}\Gamma_L} + \frac{S_{12}\Gamma_L S_{21}}{1 - S_{22}\Gamma_L}$$

Now we see that a common term is cancelled up and down at the first term on the right hand side and we arrive at Pozar's expression

$$\frac{V_1^-}{V_1^+} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L}$$

Thus Pozar arrived at an expression for Γ_{in} only involving S-parameters and Γ_L

$$\Gamma_{in} = \frac{V_1^-}{V_1^+} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L}$$

3.3 Γ_{out} -Reflection coefficient looking into the network from the load side

Obviously Γ_{out} looking into the network from the load side must be (by symmetry)

$$\Gamma_{out} = \frac{V_2^-}{V_2^+} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S}$$

which Pozar also states without evidence, but we can show this also for completeness.

The reflection coefficient looking into the source Γ_S having the network at the back is the backward moving wave V_1^+ going into the network divided by the forward going wave travelling into the source V_1^-

$$\Gamma_S = \frac{V_1^+}{V_1^-} \iff$$
$$V_1^+ = \Gamma_S V_1^-$$

which means that we can re-write Pozar's S-parameter definition

$$V_1^- = S_{11}V_1^+ + S_{12}V_2^+$$
$$V_2^- = S_{21}V_1^+ + S_{22}V_2^+$$

as

$$V_1^- = S_{11}\Gamma_S V_1^- + S_{12}V_2^+$$
$$V_2^- = S_{21}\Gamma_S V_1^- + S_{22}V_2^+$$

We are hunting for the expression Γ_{out} looking into the network from the load side so we obviously must solve for V_1^- from the first equation above and insert it into the second

$$V_1^- = S_{11}\Gamma_S V_1^- + S_{12}V_2^+ \iff$$
$$V_1^-(1 - S_{11}\Gamma_S) = S_{12}V_2^+ \iff$$
$$V_1^- = \frac{S_{12}V_2^+}{1 - S_{11}\Gamma_S}$$

Inserting the expression of V_1^- into the second equations

$$V_2^- = S_{21}\Gamma_S V_1^- + S_{22}V_2^+$$
$$= S_{21}\Gamma_S \frac{S_{12}V_2^+}{1 - S_{11}\Gamma_S} + S_{22}V_2^+$$

Solving for V_2^-/V_2^+

$$\frac{V_2^-}{V_2^+} = S_{21}\Gamma_S \frac{S_{12}}{1 - S_{11}\Gamma_S} + S_{22}$$

Which is the wanted expression if we just write it in order

$$\Gamma_{out} = \frac{V_2^-}{V_2^+} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S}$$

3.4 Availible power to the network P_{in}

Pozar want to derive an expression for the power gain $G = P_L/P_{in}$ further down the road. It is is the quotient of the power delivered to the load to the power availible to the network. Pozar is not clear, I thought that he wanted to establish a power-wave

$$P_{in} = \frac{1}{Z_0}\frac{V_1^+}{\sqrt{2}}\frac{V_1^{+*}}{\sqrt{2}}$$
$$= \frac{1}{2Z_0}|V_1^+|^2$$

but this is not the case. He defines P_{in} as

$$P_{in} = \frac{1}{2Z_0}V_1 \cdot V_1^*$$

where the factor an half comes from scaling to effective values.

He starts with establishing that the voltage $V_1 = V_1^+ + V_1^-$ is simply the voltage divder expression Because

$$V_1 = V_S \frac{Z_{in}}{Z_{in} + Z_S} = V_1^+ + V_1^- = V_1^+(1 + \Gamma_{in})$$

so he will be able to write

$$\begin{aligned} P_{in} &= \frac{1}{2Z_0} V_1 \cdot V_1^* \\ &= \frac{1}{2Z_0} (V_1^+(1+\Gamma_{in}))(V_1^+(1+\Gamma_{in})^* \\ &= \frac{1}{2Z_0} (V_1^+(1+\Gamma_{in}))(V_1^{+*}(1+\Gamma_{in}))^* \end{aligned}$$

because Γ_{in} is a complex number so

$$\begin{aligned} P_{in} &= \frac{1}{2Z_0} V_1^+ V_1^{+*}(1+\Gamma_{in})(1+\Gamma_{in}^*) \\ &= \frac{|V_1^+|^2}{2Z_0}(1+\Gamma_{in})((1+\Gamma_{in}^*) \\ &= \frac{|V_1^+|^2}{2Z_0}(1+|\Gamma_{in}|^2) \end{aligned}$$

because

$$\begin{aligned} (1+jb)(1-jb) &= 1^2 - jb + jb - j^2 b^2 \\ &= 1^2 + b^2 = 1 + |jb|^2 \end{aligned}$$

but Pozar claims

$$P_{in} = \frac{|V_1^+|^2}{2Z_0}(1-|\Gamma_{in}|^2)$$

What is he using? Does he not take the complex conjugate? Does he perhaps take

$$\begin{aligned} P_{in} &= \frac{1}{2Z_0} V_1^2 \\ &= \frac{1}{2Z_0} (V_1^+(1+\Gamma_{in}))(V_1^+(1+\Gamma_{in})) \end{aligned}$$

but Γ_{in} is still complex

$$\begin{aligned} (1+jb)(1+jb) &= 1^2 + jb + jb + j^2 b^2 \\ &= 1 - b^2 + 2jb \end{aligned}$$

so

$$\begin{aligned} P_{in} &= \frac{1}{2Z_0} V_1^2 \\ &= \frac{(V_1^+)^2}{2Z_0}(1-\Gamma_{in}^2 - 2j\Gamma_{in}) \end{aligned}$$

No Pozar is not taking the square ——————————-
Then he is using

$$\Gamma_{in} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0}$$

which is recasted to

$$Z_{in} = Z_0 \frac{1 + \Gamma_{in}}{1 - \Gamma_{in}}$$

because

$$\begin{aligned} \Gamma_{in} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} \iff \\ \Gamma_{in}(Z_{in} + Z_0) = Z_{in} - Z_0 \end{aligned}$$

Collecting terms with Z_{in} to the right and terms with Z_0 on the left

$$Z_0 + \Gamma_{in} Z_0 = Z_{in} - \Gamma_{in} Z_{in}$$

We break out Z_0 on the left and we break out Z_{in} on the right

$$Z_0(1 + \Gamma_{in}) = Z_{in}(1 - \Gamma_{in})$$

We solve for Z_{in}

$$Z_{in} = Z_0 \frac{1 + \Gamma_{in}}{1 - \Gamma_{in}}$$

He replaces Z_{in} with this expression in

$$\begin{aligned} V_S \frac{Z_{in}}{Z_{in} + Z_S} = V_1^+(1 + \Gamma_{in}) \\ V_S \frac{Z_0 \frac{1+\Gamma_{in}}{1-\Gamma_{in}}}{Z_0 \frac{1+\Gamma_{in}}{1-\Gamma_{in}} + Z_S} = V_1^+(1 + \Gamma_{in}) \end{aligned}$$

which becomes

$$V_S \frac{Z_0(1 + \Gamma_{in})}{Z_0 \frac{(1-\Gamma_{in})1+\Gamma_{in}}{1-\Gamma_{in}} + (1 - \Gamma_{in})Z_S} = V_1^+(1 + \Gamma_{in})$$

$1 + \Gamma_{in}$ cancels right and left and $1 - \Gamma_{in}$ cancels in the left most term of the denominator on the left hand side.

$$V_S \frac{Z_0}{Z_0(1 + \Gamma_{in}) + (1 - \Gamma_{in})Z_S} = V_1^+$$

He also replaces Z_S with the recasted version of Γ_S which is the reflection coefficient looking into the source

$$\Gamma_S = \frac{Z_S - Z_0}{Z_S + Z_0} \iff$$
$$Z_S = Z_0 \frac{1+\Gamma_S}{1-\Gamma_S}$$

Inserting the expression of Z_S

$$V_1^+ = V_S \frac{Z_0}{Z_0(1+\Gamma_{in}) + (1-\Gamma_{in})Z_S}$$
$$V_1^+ = V_S \frac{Z_0}{Z_0(1+\Gamma_{in}) + (1-\Gamma_{in})Z_0 \frac{1+\Gamma_S}{1-\Gamma_S}}$$

We multiply and divide the left term in the denominator with $1-\Gamma_S$ and invert this bottom factor to the numerator.

$$V_1^+ = V_S \frac{Z_0(1-\Gamma_S)}{Z_0(1+\Gamma_{in})(1-\Gamma_S) + (1-\Gamma_{in})Z_0(1+\Gamma_S)}$$

Z_0 cancels up and down. We exapand the denominator

$$V_1^+ = V_S \frac{Z_0(1-\Gamma_S)}{Z_0(1+\Gamma_{in})(1-\Gamma_S) + (1-\Gamma_{in})Z_0(1+\Gamma_S)}$$
$$= V_S \frac{(1-\Gamma_S)}{1 - \cancel{\Gamma_S} + \cancel{\Gamma_{in}} - \Gamma_{in}\Gamma_S + 1 + \cancel{\Gamma_S} - \cancel{\Gamma_{in}} - \Gamma_{in}\Gamma_S}$$

remains

$$V_1^+ = V_S \frac{(1-\Gamma_S)}{1 - \Gamma_{in}\Gamma_S + 1 - \Gamma_{in}\Gamma_S}$$
$$= V_S \frac{(1-\Gamma_S)}{2 - 2\Gamma_{in}\Gamma_S}$$

Breaking out 1/2

$$V_1^+ = \frac{V_S}{2} \frac{1-\Gamma_S}{1-\Gamma_{in}\Gamma_S}$$

so P_{in} becomes if we use Pozar's definition

$$P_{in} = \frac{|V_1^+|^2}{2Z_0}(1-|\Gamma_{in}|^2)$$
$$= \frac{|\frac{V_S}{2}\frac{1-\Gamma_S}{1-\Gamma_{in}\Gamma_S}|^2}{2Z_0}(1-|\Gamma_{in}|^2)$$
$$= \frac{|V_S|^2|1-\Gamma_S|^2}{8|1-\Gamma_{in}\Gamma_S|^2 Z_0}(1-|\Gamma_{in}|^2)$$

Pozar also says that P_{in} becomes this with his definition of P_{in}

$$P_{in} = \frac{|V_S|^2}{8Z_0}\frac{|1-\Gamma_S|^2(1-|\Gamma_{in}|^2)}{|1-\Gamma_{in}\Gamma_S|^2}$$

3.5 Power delivered to the load P_L

Pozar states that P_L is

$$P_L = \frac{|V_2^-|^2}{2Z_0}(1-|\Gamma_L|^2)$$

possibly by a symmetry argument because he is saying that

$$P_{in} = \frac{|V_1^+|^2}{2Z_0}(1-|\Gamma_{in}|^2)$$

For a more involved expression of P_L He solves for V_2^- of the scattering matrix definition to elimate V_2^- in expression of P_L

$$\begin{aligned}
V_2^- &= S_{21}V_1^+ + S_{22}V_2^+ \\
&= S_{21}V_1^+ + S_{22}\Gamma_L V_2^- \\
V_2^-(1-S_{22}\Gamma_L) &= S_{21}V_1^+ \\
V_2^- &= \frac{S_{21}V_1^+}{1-S_{22}\Gamma_L}
\end{aligned}$$

Inserting V_2^- in the expression of P_L becomes

$$\begin{aligned}
P_L &= \frac{|V_2^-|^2}{2Z_0}(1-|\Gamma_L|^2) \\
&= \frac{\left|\frac{S_{21}V_1^+}{1-S_{22}\Gamma_L}\right|^2}{2Z_0}(1-|\Gamma_L|^2) \\
&= \frac{\left|S_{21}V_1^+\right|^2}{|1-S_{22}\Gamma_L|^2 2Z_0}(1-|\Gamma_L|^2)
\end{aligned}$$

Then he is inserting

$$V_1^+ = \frac{V_S}{2}\frac{1-\Gamma_S}{1-\Gamma_{in}\Gamma_S}$$

$$
\begin{aligned}
P_L &= \frac{\left|S_{21}V_1^+\right|^2}{|1-S_{22}\Gamma_L|^2 2Z_0}(1-|\Gamma_L|^2) \\
&= \frac{\left|S_{21}\frac{V_S}{2}\frac{1-\Gamma_S}{1-\Gamma_{in}\Gamma_S}\right|^2}{|1-S_{22}\Gamma_L|^2 2Z_0}(1-|\Gamma_L|^2) \\
&= \frac{|S_{21}V_S(1-\Gamma_S)|^2}{4|1-\Gamma_{in}\Gamma_S|^2|1-S_{22}\Gamma_L|^2 2Z_0}(1-|\Gamma_L|^2) \\
&= \frac{|S_{21}|^2|V_S|^2|1-\Gamma_S|^2}{8|1-\Gamma_{in}\Gamma_S|^2|1-S_{22}\Gamma_L|^2 Z_0}(1-|\Gamma_L|^2)
\end{aligned}
$$

asserting Pozar's expression of P_L Equation 12.7

$$
P_L = \frac{|V_S|^2}{8Z_0}\frac{|S_{21}|^2(1-|\Gamma_L|^2)|1-\Gamma_S|^2}{|1-S_{22}\Gamma_L|^2|1-\Gamma_S\Gamma_{in}|^2}
$$

3.6 Power gain P_L/P_{in}

If we use Pozar's expression of P_{in} and divide P_L/P_{in}

$$
\begin{aligned}
G = \frac{P_L}{P_{in}} &= \frac{\frac{|S_{21}|^2\cancel{|V_S|^2}\cancel{|1-\Gamma_S|^2}}{\cancel{8}\cancel{|1-\Gamma_{in}\Gamma_S|^2}|1-S_{22}\Gamma_L|^2\cancel{Z_0}}(1-|\Gamma_L|^2)}{\frac{\cancel{|V_S|^2}}{\cancel{8}\cancel{Z_0}}\frac{\cancel{|1-\Gamma_S|^2}(1-|\Gamma_{in}|^2)}{\cancel{|1-\Gamma_{in}\Gamma_S|^2}}} \\
&= \frac{|S_21|^2(1-|\Gamma_L|^2)}{|1-S_{22}\Gamma_L|^2(1-|\Gamma_{in}|^2)}
\end{aligned}
$$

3.7 P_{avs} - the maximum power availible from the source

Pozar continues defining P_{avs} as the maximum power availible from the source

$$
\begin{aligned}
P_{avs} &= P_{in}\Big|_{\Gamma_{in}^*=\Gamma_S} \\
&= \frac{|V_S|^2}{8Z_0}\frac{|1-\Gamma_S|^2(1-|\Gamma_{in}|^2)}{|1-\Gamma_{in}\Gamma_S|^2}\Bigg|_{\Gamma_{in}^*=\Gamma_S} \\
&= \frac{|V_S|^2}{8Z_0}\frac{|1-\Gamma_S|^2(1-|\Gamma_S^*|^2)}{|1-\Gamma_S^*\Gamma_S|^2}
\end{aligned}
$$

We have that $\Gamma_S^*\Gamma_S = |\Gamma_S|^2$

$$
\begin{aligned}
P_{avs} &= \frac{|V_S|^2}{8Z_0}\frac{|1-\Gamma_S|^2(1-|\Gamma_S^*|^2)}{|1-|\Gamma_S|^2|^2} \\
&= \frac{|V_S|^2}{8Z_0}\frac{|1-\Gamma_S|^2}{|1-|\Gamma_S|^2|}
\end{aligned}
$$

which Pozar writes as

$$
P_{avs} = \frac{|V_S|^2}{8Z_0}\frac{|1-\Gamma_S|^2}{(1-|\Gamma_S|^2)}
$$

Apparently $|1-|\Gamma_S|^2| = (1-|\Gamma_S|^2)$ which is true if $|\Gamma_S|^2 \leq 1$.

3.8 P_{avn} - the maximum power availible from the network

Pozar defines P_{avn} ss the maximum power availible from the network which is when $\Gamma_{out}^* = \Gamma_L$

$$
\begin{aligned}
P_{avn} &= P_L\Big|_{\Gamma_L=\Gamma_{out}^*} \\
&= \frac{|V_S|^2}{8Z_0}\frac{|S_{21}|^2(1-|\Gamma_L|^2)|1-\Gamma_S|^2}{|1-S_{22}\Gamma_L|^2|1-\Gamma_S\Gamma_{in}|^2}\Bigg|_{\Gamma_L=\Gamma_{out}^*} \\
&= \frac{|V_S|^2}{8Z_0}\frac{|S_{21}|^2(1-|\Gamma_{out}^*|^2)|1-\Gamma_S|^2}{|1-S_{22}\Gamma_{out}|^2|1-\Gamma_S\Gamma_{in}|^2}
\end{aligned}
$$

Pozar the states that Γ_{in} must be evaluated for $\Gamma_L = \Gamma_{out}^*$ and that it can be shown that

$$
|1-\Gamma_S\Gamma_{in}|^2\Bigg|_{\Gamma_L=\Gamma_{out}^*} = \frac{|1-S_{11}\Gamma_S|^2(1-|\Gamma_{out}|^2)^2}{|1-S_{22}\Gamma_{out}^*|^2}
$$

I've tried to show this but cannot. But we insert this and see what happens

$$
\begin{aligned}
P_{avn} &= \frac{|V_S|^2}{8Z_0}\frac{|S_{21}|^2(1-|\Gamma_{out}^*|^2)|1-\Gamma_S|^2}{|1-S_{22}\Gamma_{out}|^2|1-\Gamma_S\Gamma_{in}|^2} \\
&= \frac{|V_S|^2}{8Z_0}\frac{|S_{21}|^2(1-|\Gamma_{out}^*|^2)|1-\Gamma_S|^2}{|1-S_{22}\Gamma_{out}|^2\frac{|1-S_{11}\Gamma_S|^2(1-|\Gamma_{out}|^2)^2}{|1-S_{22}\Gamma_{out}^*|^2}}
\end{aligned}
$$

The expression $(1-|\Gamma_{out}^*|^2)$ in the numerator seems to cancel and we can invert the expression $|1-S_{22}\Gamma_{out}^*|^2$ from the denominator to the numerator but that's it

$$
P_{avn} = \frac{|V_S|^2}{8Z_0}\frac{|S_{21}|^2\cancel{|1-S_{22}\Gamma_{out}^*|^2}|1-\Gamma_S|^2}{\cancel{|1-S_{22}\Gamma_{out}|^2}|1-S_{11}\Gamma_S|^2(1-|\Gamma_{out}|^2)}
$$

Cancellation up and down

$$
P_{avn} = \frac{|V_S|^2}{8Z_0}\frac{|S_{21}|^2|1-\Gamma_S|^2}{|1-S_{11}\Gamma_S|^2(1-|\Gamma_{out}|^2)}
$$

3.9 G_A - Availible power gain

Pozar defines G_A as the maximum power possible delivered from the network P_{avn} to the maximum power availibe from the source which he derived from conjugately matched reflection coefficients

$$
\begin{aligned}
G_A = \frac{P_{avn}}{P_{avs}} &= \frac{\frac{|V_S|^2}{8Z_0}\frac{|S_{21}|^2|1-\Gamma_S|^2}{|1-S_{11}\Gamma_S|^2(1-|\Gamma_{out}|^2)}}{\frac{|V_S|^2}{8Z_0}\frac{|1-\Gamma_S|^2}{(1-|\Gamma_S|^2)}} \\
&= \frac{|S_{21}|^2(1-|\Gamma_S|^2)}{|1-S_{11}\Gamma_S|^2(1-|\Gamma_{out}|^2)}
\end{aligned}
$$

which confirms Pozar's equation 12.12

3.10 G_{TU} Tranceducer power gain

Pozar defines the tranceducer power gain G_{TU} as the power delivered to the load P_L to the maximum power avialible from the source P_{avs}

$$
\begin{aligned}
G_{TU} = \frac{P_L}{P_{avs}} &= \frac{\frac{|V_S|^2}{8Z_0}\frac{|S_{21}|^2(1-|\Gamma_L|^2)|1-\Gamma_S|^2}{|1-S_{22}\Gamma_L|^2|1-\Gamma_S\Gamma_{in}|^2}}{\frac{|V_S|^2}{8Z_0}\frac{|1-\Gamma_S|^2}{(1-|\Gamma_S|^2)}} \\
&= \frac{|S_{21}|^2(1-|\Gamma_L|^2)(1-|\Gamma_S|^2)}{|1-S_{22}\Gamma_L|^2|1-\Gamma_S\Gamma_{in}|^2}
\end{aligned}
$$

A special case, writes Pozar, arises when $\Gamma_L = 0$ and $\Gamma_S = 0$ which would correspond to a non-resonant network which reduces $G_{TU} = S_{21}^2$ in contrast to a resonant network. He separates G_{TU} as

$$
\begin{aligned}
G_{TU} &= \frac{|S_{21}|^2(1-|\Gamma_L|^2)(1-|\Gamma_S|^2)}{|1-S_{22}\Gamma_L|^2|1-\Gamma_S\Gamma_{in}|^2} \\
&= \underbrace{\frac{1-|\Gamma_S|^2}{|1-\Gamma_S\Gamma_{in}|^2}}_{G_S}\cdot\underbrace{|S_{21}|^2}_{G_0}\cdot\underbrace{\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}}_{G_L}
\end{aligned}
$$

4 Stability

Stability requires $|\Gamma_{in}| < 1$ and $|\Gamma_{out}| < 1$ We verified Pozar's expression of the above. We have the backward voltage wave to the forward volatage wave looking into the network from the source side.

$$\Gamma_{in} = \frac{V_1^-}{V_1^+} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L}$$

and similarly the backwardmoving voltage wave looking into the network from the load side (moving into the load) to the forward moving voltage wave (reflected from the load).

$$\Gamma_{out} = \frac{V_2^-}{V_2^+} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_S}$$

so stability requires

$$
\begin{aligned}
|\Gamma_{in}| &= \left|S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L}\right| < 1 \\
|\Gamma_{out}| &= \left|S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_S}\right| < 1
\end{aligned}
$$

Of the requirement for Γ_{in} Pozar derives the so called "output stability circle" which poses restrictions on Γ_L for stability. We've checked the input and output stability circles already and they don't look correct comparing with the calculated Rollet stabilty graf.

5 Design for maximum gain

We know that for maximum power transfer which also means resonance in passive networks $\Gamma_{in}^* = \Gamma_S$ and $\Gamma_{out}^* = \Gamma_L$

$$
\begin{aligned}
G_{TU} &= \underbrace{\frac{1-|\Gamma_S|^2}{|1-\Gamma_S\Gamma_{in}|^2}}_{G_S}\cdot\underbrace{|S_{21}|^2}_{G_0}\cdot\underbrace{\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}}_{G_L} \\
&= \underbrace{\frac{1-|\Gamma_S|^2}{|1-|\Gamma_S|^2|^2}}_{G_S}\cdot\underbrace{|S_{21}|^2}_{G_0}\cdot\underbrace{\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}}_{G_L}
\end{aligned}
$$

Up and down factors cancels and we get

$$G_{TU} = \underbrace{\frac{1}{|1-|\Gamma_S|^2|}}_{G_S} \cdot \underbrace{|S_{21}|^2}_{G_0} \cdot \underbrace{\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}}_{G_L}$$

To achive this we solve

$$\Gamma_S^* = \Gamma_{in} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L}$$
$$\Gamma_L^* = \Gamma_{out} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_S}$$

which gives the solutions

$$\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|}}{2C_1}$$

where

$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - \Delta$$
$$C_1 = S_{11} - \Delta \cdot S_{22}^*$$

and Γ_L is given by

$$\Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|}}{2C_2}$$

where

$$B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - \Delta$$
$$C_2 = S_{22} - \Delta \cdot S_{11}^*$$

We are requested to provide a design for 2.45GHz but we don't have S-parameters for that frequecy so we decided on 2.4 GHz and hoping that the design would have enough bandwidth. For the desing we used Octave and verified each step with a Matlab library[9]

```
clear all
% addpath("/home/lasse/ewa")
%Add above statement to console input
i=sqrt(-1);
disp("S-Params at 1.8GHz BFG520 ...
       Common Emitter 6V10mA")
S11_r = 0.494
Theta_S11_grad=166
S21_r=3.722
Theta_S21_grad=74.3
S12_r=0.082
Theta_S12_grad= 56.4
S22_r=0.317
Theta_S22_grad= -57.9
%Call to the EMW-library
disp("")
S=smat([0.494 166.0 3.722 74.3 0.082 ...
     56.4 0.317 -57.9 ])
disp("")
```

[9] `http://eceweb1.rutgers.edu/~orfanidi/ewa/`

We convert to cartesian coordinates, not shown here, but the full program will in the appendix We did the Rollet-stability test

```
%Stability check
disp('Stability check')
Delta = S11*S22-S12*S21
K_Stab=(1-abs(S11)^2-abs(S22)^2+abs(Delta)^2)...
      /(2*abs(S12*S21))
Abs_Delta = abs(Delta)
if(K_Stab>1 & Abs_Delta<1)
 disp("Stability OK")
else
 disp("Stability not OK")
end
```

Which resulted in $K = 1.1220$ and $|\Delta| = 0.175$ which means that it stable at the frequecy though unstable at lower frequencies as previously shown. Stability circles using the toolbox are using commands

```
[cL,rL] = sgcirc(S,'l');
[cG,rG] = sgcirc(S,'s');
smith;
smithcir(cL, rL, 1.1, 1.5);
smithcir(cG, rG, 1.1, 1.5);
```

Unfourtunately the toolbox doesn't seem to give the option to insert a legend but the upper is the load stability circle and the lower is the source stability cirlce. They are both outside the unit circle and this means that any source and load impedances will not result in instability.

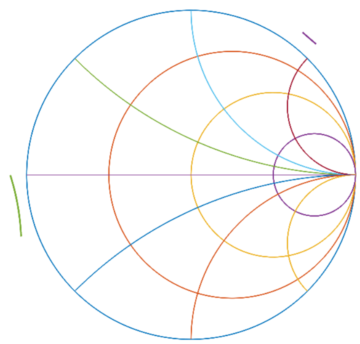


Figure 17: Load stabilty circle (upper). Source stability circle (lower)

We calculated to source and load reflection coefficients Γ_s and Γ_L for a conjugate match which means $\Gamma_s^* = \Gamma_{in}$ and $\Gamma_L^* = \Gamma_{out}$.

```
%Maximum gain calculations
disp("Maximum gain calculations")
B1=1+abs(S11)^2-abs(S22)^2 -abs(Delta)^2
B2=1+abs(S22)^2-abs(S11)^2 -abs(Delta)^2
C1=S11 - Delta*conj(S22)
C2=S22 - Delta*conj(S11)
disp("")
%Library calculation
[K,mu,D,B1,B2,C1,C2,D1,D2] = sparam(S)
disp("")


Gamma_S_plus  = (B1 + sqrt(B1^2
               -4*abs(C1)^2) )/(2*C1);
Gamma_S_minus = (B1 - sqrt(B1^2
               -4*abs(C1)^2) )/(2*C1);
if(abs(Gamma_S_plus)<1)
  Gamma_MS=Gamma_S_plus
  disp("Returned Gamma_S_plus")
else
  Gamma_MS=Gamma_S_minus
  disp("Returned Gamma_S_minus")
endif

Gamma_L_plus  = (B2 + sqrt(B2^2 ....
               -4*abs(C2)^2) )/(2*C2);
Gamma_L_minus = (B2 - sqrt(B2^2....
               -4*abs(C2)^2) )/(2*C2);
if(abs(Gamma_L_plus)<1)
  Gamma_ML=Gamma_L_plus
  disp("Returned Gamma_L_plus")
else
  Gamma_ML=Gamma_L_minus
  disp("Returned Gamma_L_minus")
endif
```

In all situations when we tested different S-parameters to learn we always for some reason had to pick the value which was calculated with the minus sign before the square root. Our Octave print-out was the following

```
Gamma_MS = -0.7395 - 0.1306i
Gamma_ML =  0.4402 + 0.5100i

AbsGamma_MS = 0.7510
Theta_deg = -169.99
AbsGamma_ML = 0.6737
Theta_deg = 49.204

Gammas according to Matlab toolbox
Gamma_S_Match = -0.7395 - 0.1306i
Gamma_L_Match =  0.4402 + 0.5100i
```

Matching is simply done by plotting the corresponding impedance on the smith chart and transforming it to the load impedance/generator impedance, in this case 50 Ohm, but always walk “towards the load”. Everything works precisely the same for as for single stub matching of a 50 Ohm line when you walk “towards the generator” but here instead you walk “towards the load”.

We calculated the tranceducer gain G_T correct, in agreement witht the toolbox result but could not get the correct result for for the operating power gain G and the availibe power gain G_A.

```
%Gains of matching networks
disp("Gains of matching networks")
%Gain source matching network
G_S=1/(1-abs(Gamma_MS)^2)
G_S_dB=10*log10(G_S)
%Intrinsic gain
G_0=abs(S21)^2
G_0_dB=10*log10(G_0)
%Gain load matching network
G_L=(1-abs(Gamma_ML)^2)...
       /(abs(1-S22*Gamma_ML)^2)
G_L_dB=10*log10(G_L)
disp("Gt gain")
Gt = G_0*G_S*G_L
Gt_dB = G_S_dB+G_0_dB+G_L_dB
Gamma_L_unmatched =0;%no-conjugate match
%Power gain
```

```
%Source conj. matched but not the load
G_power=(1/(1-abs(Gamma_MS)^2))*abs(S21)^2.
    *((1-abs(Gamma_L_unmatched)^2)...
     /(abs(1-S22*Gamma_L_unmatched)^2))
G_power_dB=10*log10(G_power)
%Availible power gain
%Load is conj.matched but not source
Gamma_G_unmatched=0
G_availible=((1-abs(Gamma_G_unmatched)^2)...
    /(abs(1-S11*Gamma_G_unmatched)^2))...
    *(abs(S21)^2)*(1/(1-abs(Gamma_ML)^2))
G_availible_dB=10*log10(G_availible)
```

The print out was the following ommitting unneccesary data

```
Gains of matching networks
G_S_dB = 3.6047
G_0_dB = 11.416
G_L_dB = -0.5749
Gt_dB = 14.445
G_power_dB = 15.020
G_availible_dB = 14.043

Gains from toolbox
transducer power gain
   at given optimized GammmaG and GammaL
Gt_dB = 14.445
available power gain at
   given GammaG with GammaL = GammaOut*
Ga_dB = 11.875
operating power gain
     at given GammaL with GammaG = GammaIn*
Gp_dB = 12.631
maximum unilateral gain
Gu_dB = 13.090
maximum available gain (MAG)
Gmag_dB = 14.445
maximum stable gain (MSG)
Gmsg_dB = 16.570
```

I have not understood the gains correctly reagarding how to calculate them. The input and output stubs and the corresponding serial lines were calculated through the toolbox

```
disp("Stubs")
dl_in = stub1 (conj(z_S) ,'po')
dl_out = stub1 (conj(z_L) ,'po')
```

which give two pairs of solutions each The print out is the following

```
dl_in =

   0.315925   0.428694
   0.184075   0.043487

dl_out =

   0.3298   0.1155
   0.1702   0.2478
```

The shortest input stub is $d_{in} = 0.184075\lambda$ and its corresponding serial line is $l_{in} = 0.043487\lambda$. For the output stub we also picked the shortest $d_{in} = 0.1702\lambda$ with series line $l_{in} = 0.2478\lambda$. This was the design

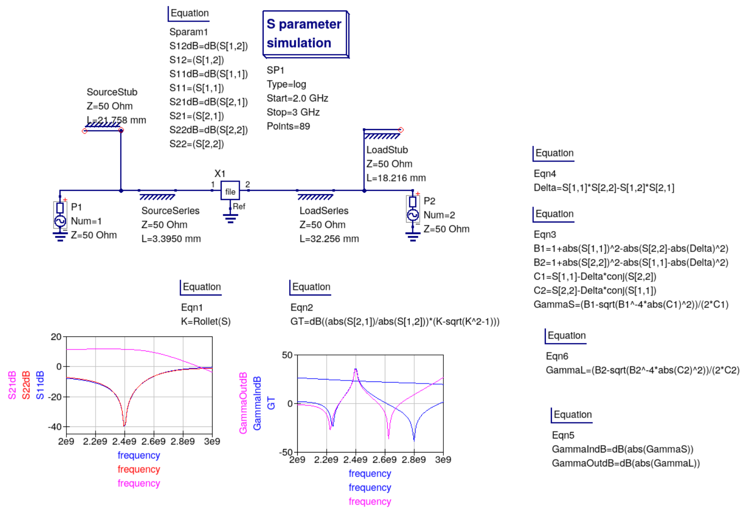


Figure 18: Amplifier with T-lines optimized for 2.4GHz

We see the resonance of Γ_{in} and Γ_{out} which gives the fact that a passive network can have gains larger than one. The matching networks are in fact LC-tanks in resonance. For 2.4GHz we have return loss taking the $\check{+}$dB$+$ of $S11$ of $2 \times 44.6 = 89.2$dB.

frequency	S21dB	S22dB	S11dB
2.39e9	11.6	-34.8	-34.9
2.39e9	11.6	-35	-35.2
2.39e9	11.6	-35.2	-35.4
2.39e9	11.6	-35.5	-35.6
2.39e9	11.6	-35.7	-35.9
2.39e9	11.6	-36	-36.1
2.39e9	11.6	-36.2	-36.4
2.39e9	11.6	-36.5	-36.6
2.39e9	11.6	-36.8	-36.9
2.39e9	11.6	-37.1	-37.2
2.39e9	11.6	-37.4	-37.5
2.39e9	11.6	-37.7	-37.8
2.39e9	11.6	-38	-38.1
2.39e9	11.6	-38.3	-38.5
2.39e9	11.6	-38.7	-38.8
2.39e9	11.6	-39	-39.2
2.39e9	11.6	-39.4	-39.5
2.39e9	11.6	-39.8	-39.9
2.39e9	11.6	-40.2	-40.3
2.39e9	11.6	-40.6	-40.8
2.39e9	11.6	-41.1	-41.2
2.39e9	11.6	-41.6	-41.7
2.39e9	11.6	-42.1	-42.2
2.4e9	11.6	-42.6	-42.7
2.4e9	11.6	-43.2	-43.3
2.4e9	11.6	-43.8	-43.9
2.4e9	11.6	-44.5	-44.6

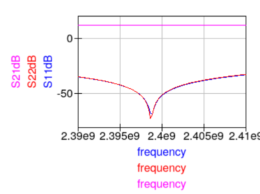


Figure 19: G_T and returnloss for 2.4GHz

For the specification of 2.45GHz it is somewhat lesser good because we had no S-Parameters for this frequency I did not know how to extrapolate the figures to 2.45GHz.

frequency	S21dB	S22dB	S11dB
2.44e9	11.4	-21	-21.4
2.44e9	11.4	-21	-21.3
2.44e9	11.4	-20.9	-21.2
2.44e9	11.4	-20.9	-21.2
2.44e9	11.4	-20.8	-21.1
2.44e9	11.4	-20.8	-21.1
2.44e9	11.4	-20.7	-21.1
2.44e9	11.4	-20.7	-21
2.44e9	11.4	-20.6	-21
2.44e9	11.4	-20.6	-20.9
2.44e9	11.4	-20.5	-20.9
2.44e9	11.4	-20.5	-20.8
2.44e9	11.4	-20.4	-20.8
2.44e9	11.4	-20.4	-20.7
2.44e9	11.4	-20.3	-20.7
2.44e9	11.4	-20.3	-20.6
2.44e9	11.4	-20.2	-20.6
2.44e9	11.4	-20.2	-20.5
2.44e9	11.4	-20.2	-20.5
2.44e9	11.4	-20.1	-20.4
2.44e9	11.4	-20.1	-20.4
2.44e9	11.4	-20	-20.3
2.44e9	11.4	-20	-20.3
2.45e9	11.4	-19.9	-20.2
2.45e9	11.4	-19.9	-20.2
2.45e9	11.4	-19.8	-20.2
2.45e9	11.4	-19.8	-20.1

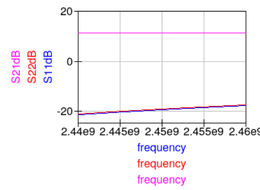


Figure 20: G_T and returnloss for 2.45GHz

We have returnloss of 40 dB and a tranceducer gain of 11.4 dB. For the Microstrip version we calculated the effective wavelength λ_g. We created the following code to calculate ϵ_{eff} and the width of the microstrip

```
f=2.45E+9
c=3E+8
lambda= c/f;
ko=2*pi/lambda
epsilon_r=4.4
Z0=50
d=1.6E-3
loss_tangent=0.0018

%Accordin to Pozar's book
A=(Z0/60)*sqrt((epsilon_r +1)/2)+((epsilon_r-1)/(epsilon_r+1))*(0.23+0.11/epsilon_r)
B=(377*pi)/(2*Z0*sqrt(epsilon_r))

%For a given impedance Z0 and dielectric constant epsilon_r
%W/d-ratio calculated using the two formulas
%and the value is selected depinding if the output corresponds to the right restrictions
%on its origin
W2d_ratio1 =(8*exp(A))/(exp(2*A)-2)
W2d_ratio2 =(2/pi)*(B-1-log(2*B-1)+((epsilon_r-1)/(2*epsilon_r))*(log(B-1)+0.39-(0.61/epsilon_r)))
if(W2d_ratio1 <2)
 W2d_ratio = W2d_ratio1
 disp("Returned the first ratio")
elseif(W2d_ratio2 >2)
 W2d_ratio = W2d_ratio1
 disp("Returned the second ratio")
end
%Need epsilon effective for reduced wavelength
epsilon_eff = ((epsilon_r+1)/2)+((epsilon_r-1)/2)*(1/(sqrt(1+12*1/( W2d_ratio))))
attenuation_dielectric =(ko*epsilon_r*(epsilon_eff-1)*loss_tangent)/(2*sqrt(epsilon_eff)*(epsilon_r-
```

```
Width =  W2d_ratio*d
```

The output is

```
f = 2.4500e+09
c = 3.0000e+08
ko = 51.313
epsilon_r = 4.4000
Z0 = 50
d = 1.6000e-03
loss_tangent = 1.8000e-03
A = 1.5299
B = 5.6463
W2d_ratio1 = 1.9119
W2d_ratio2 = 1.9134
W2d_ratio = 1.9119
Returned the first ratio
epsilon_eff = 3.3302
attenuation_dielectric = 0.076313
Width = 3.0590e-03
```

Noticing now that I am cheating because the optimization is done for 2.4 GHz and the idea was to just own the result of what the simulation gave us at 2.45GHz. It was an error to use the wavelength corresponding to 2.45GHz!

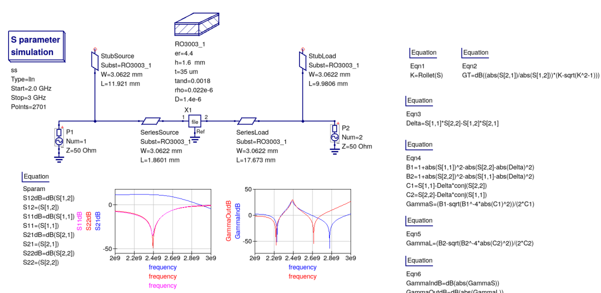


Figure 21: Amplifier with microstrip lines optimized for 2.4GHz

Without Tee-bends

frequency	S11dB	S21dB	S22dB
2.39e9	-34.7	11.5	-38.6
2.39e9	-35	11.5	-42.7
2.39e9	-34.7	11.5	-48.2
2.4e9	-34	11.5	-46.5
2.4e9	-32.9	11.5	-41.1
2.4e9	-31.8	11.5	-37.6
2.4e9	-30.7	11.5	-35
2.41e9	-29.6	11.4	-33
2.41e9	-28.7	11.4	-31.4
2.41e9	-27.7	11.4	-30

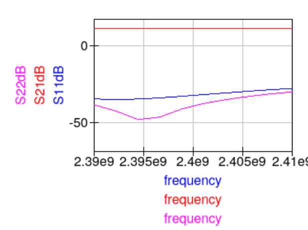


Figure 22: G_T and returnloss for 2.4GHz

At 2.45 GHz we have $G_T = 11.2$dB and a returnloss of around 36dB.

frequency	S11dB	S21dB	S22dB
2.44e9	-19.7	11.3	-20.1
2.44e9	-19.3	11.3	-19.6
2.44e9	-18.9	11.2	-19.2
2.45e9	-18.5	11.2	-18.8
2.45e9	-18.1	11.2	-18.4
2.45e9	-17.8	11.2	-18
2.45e9	-17.4	11.2	-17.7
2.46e9	-17.1	11.2	-17.3
2.46e9	-16.8	11.2	-17
2.46e9	-16.5	11.1	-16.7

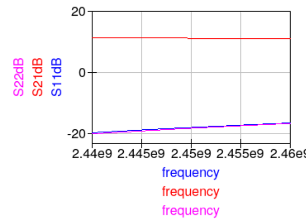
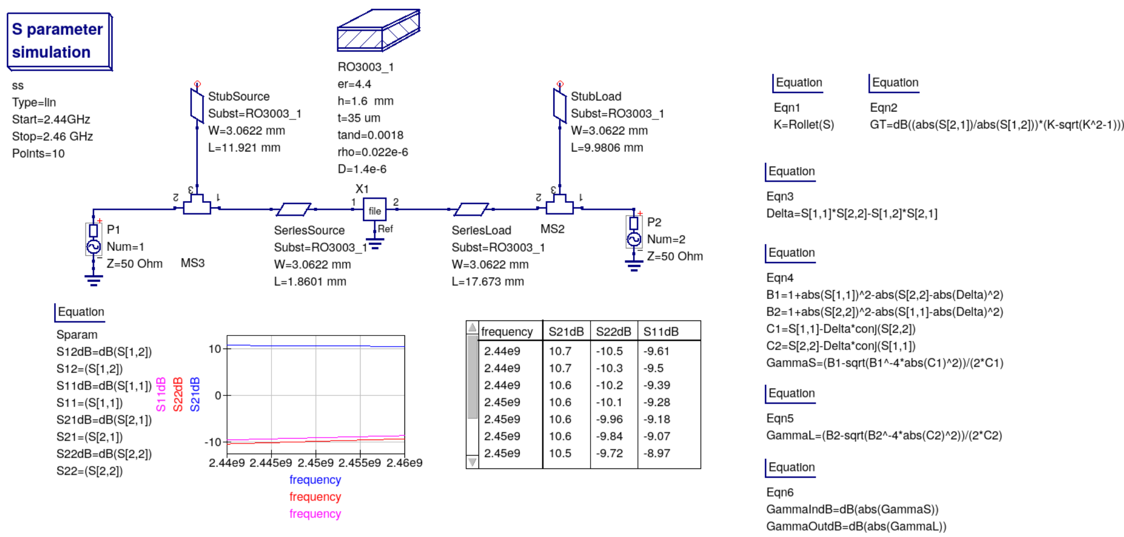


Figure 23: G_T and returnloss for 2.45GHz

There is not much loss including the Tees. Returnloss is 21 dB and and $G_T = 10.6$dB



frequency	S21dB	S22dB	S11dB
2.44e9	10.7	-10.5	-9.61
2.44e9	10.7	-10.3	-9.5
2.44e9	10.6	-10.2	-9.39
2.45e9	10.6	-10.1	-9.28
2.45e9	10.6	-9.96	-9.18
2.45e9	10.6	-9.84	-9.07
2.45e9	10.5	-9.72	-8.97

Figure 24: Amplifier with Tee-sections at 2.45GHz

6 Design for specific gain 13 dB (lowest possible Noise Figure)

We wanted to design for an exact gain of 13 dB at 1.8 MHz with as small Noise-figure as possible. We did not have noise-figure data for 1.8GHz so we used the data for 2.0 GHz. We didn't follow the outline as presented by Pozar who describes the unilateral procedure but followed the path described in "Electromagnetic Waves & Antenna"[10] by Orfanidis using Availibel power -gain circles where the load-side is matched which is also the preferred method for desingning Low Noise Amplifiers. We follwed the path outlined in the book and used the toolbox though little editing had to be used because of older matlab-syntax and a few plot-commands were in the somewhat wrong order and didn't display well. The minimum Noise Figure was found to be 1.96 dB at 13dB of gain where we didn't account for microstrip losses.

[10] http://eceweb1.rutgers.edu/~orfanidi/ewa/

Plotting the components of the trancducer gain we noticed that the output matching network was damping

```
Gains of matching networks
G_S = 2.2933
G_S_dB = 3.6047
G_L = 0.8760
G_L_dB = -0.5749
Gt gain
Total_Gain = 27.831
Total_Gain_dB = 14.445
```

We didn't want to add more damping so we diecided for the availible power gain approach The Noise-figure for a conjugate match The optimal refelction coefficient for the noise figure taken from the touchstone-file was used and the corresponding Γ_{out} calculated with the standard formula

$$\Gamma_{out} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S}$$

The availible gain at optimal nois-figure was recorded and the corresponding $\Gamma_L = \Gamma^*_{out}$ noted

```
disp("Noise figure calculations")
%From Touchstone file at 2 000 MHz
%2000  1.90  .242  134.0   .160
Fmin = 1.90;  rn = 0.16;
%Optimal NF GammaS reflection coeff
gGopt = p2c(0.242 ,134.0 );
% maximum available gain
Gmag = db(sgain(S,'mag'))
% available gain at GammaS opt
Gaopt = db(sgain(S,gGopt,'a'))
% matched load
gLopt = conj(gout(S,gGopt))
figure;
smith;
plot(gGopt,'*r')
plot(gLopt,'xk')
```

The output

```
Noise figure calculations
Gmag = 14.445
Gaopt = 12.304
gLopt =  0.1528 + 0.3450i
```

We see that the reflection coefficient giving the optimal noisefigure of 1.9 dB does not give an availible Gain G_A of 13 dB which is our target. Hence we must allow for a higer noise figure. We plot on the Smith chart the the optimal noise reflection coefficient for the source and its corresponding conjugate to Γ_{out} which means that the output is conjugately matched

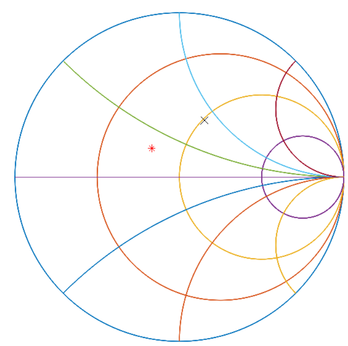


Figure 25: $\Gamma_{S,optN}$ (*) and $\Gamma_L = \Gamma^*_{out}$(x)

We plot Noise figure cirlces for 1.95, 1.96,2.1 and 2.2 dB on the chart

```
[c0,r0]=nfcirc(1.9,Fmin,rn,gGopt);
[c1,r1]=nfcirc(1.95,Fmin,rn,gGopt);
[c2,r2]=nfcirc(1.96,Fmin,rn,gGopt);
[c3,r3]=nfcirc(2.1,Fmin,rn,gGopt);
[c4,r4]=nfcirc(2.2,Fmin,rn,gGopt);

smithcir(c0,r0);%A point only
plot(c0,'*')%Same point as previously
smithcir(c1,r1);
plot(c1,'*')
smithcir(c2,r2);
plot(c2,'*')
smithcir(c3,r3);
plot(c3,'*')
smithcir(c4,r4);
plot(c4,'*')
```

We notice that the circles have almost the same center-point and the fact that these are circles stem for the fact that there are many reflection coefficients that can accomplish a certain Noise figure greater than the minimal NF.

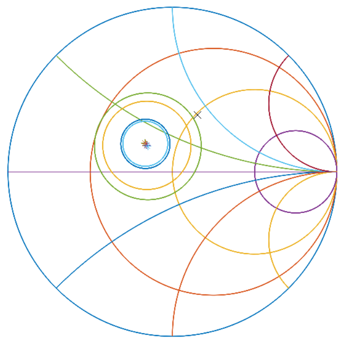


Figure 26: Constant Noise-figure circles at 1.95,1.96, 2.0 and 2.2 dB

What is the noise figure for the conjugate match? We calculate it and we also plot the optimal reflection coefficients for the conjugate match which gives 14.445 dB gain.

```
[gG,gL] = smatch(S)
plot([gG,gL],'*g')
F = nfig(Fmin, rn, gGopt, gG)
```

Output is

```
gG = -0.7395 - 0.1306i
gL =  0.4402 + 0.5100i
F = 3.8037
```

The noise figure for the max gain case (14.445 dB) is 3.8037 dB.

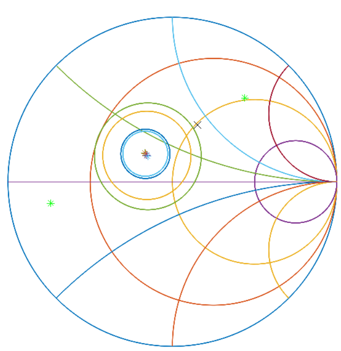


Figure 27: Green stars added for max gain cojugate match reflection coefficients. Γ_S is down left

Next we want to store all source reflection coefficients related to a particular Noise figure circle we are interested in and calculate the gain for each of these circles and pick out the one which gives an availibel power gain of 13 dB. We either know from first hand which noise-figure circle or we tested to see which gains were associated

```
%Create an array of phi-values
%in incremenets of a half degree
%from 0 to 2$\pi$
phi = linspace(0,2*pi,721);
% Record the GammaG's around the c 2 , r 2 circle
gGammas = c2 + r2*exp(j*phi);
% Calculate the available gain in dB
G = db(sgain(S,gGammas,'a'));
%Pick out the index with the
%highest gain
[Ga,i] = max(G)
%If you need to search for a gain on
%the circle use this code
find an index in array of a specific gain
index=1;
sizeG = 721;
for(i=1:1:sizeG)
%G(i)
%Edit the gain values for a different gain
 if (G(i)>13.0 ) && (G(i)<13.1)
   index=i
   G(i)
   disp("found index")
   break;
 endif
end
% GammaS for maximum gain
%if you already know the correct
%Noise-figure circle and just
%want to pick the highest gain
%of corresponding noise-figure circle
gammaS = gGammas(i)
%If you look for a specific gain
%run this line instead
%gammaS = gGammas(index)
```

Output is

```
Ga = 13.001
i = 427
sizeG = 721
gammaG = -0.289592 + 0.088463i
```

so the maximum availible gain G_A on the 1.96 dB Noise figure circle happen to be 13.001 and the corresponding Γ_S value is -0.289592 + 0.088463i.

We want to calculate Γ_{out} for this value of Γ_S with the standard formula and take the conjugate and assign that conjugate value to Γ_L.

$$\Gamma_{out} = S_{22} + \frac{S_{21}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S}$$
$$\Gamma_L = \Gamma_{out}^*$$

We also want to plot the availibe gain circles for 13 dB and see if it has common points with the 1.96 dB Noise figure circle. We also want to plot the the Γ_S that we found and it should lie on the 13 dB availible gain circle as well as on the 1.96 dB Noise-figure circle.

```
gammaL = conj(gout(S,gammaG))
% Available gain circle 13.001 dB
[ca,ra] = sgcirc(S,'a',Ga);
%Center point with a red asterix
plot(ca,'*r')
smithcir(ca,ra);
%GammaG and GammaL added to Smitchart
%with blue asterix
plot([gammaS,gammaL],'*b');
```

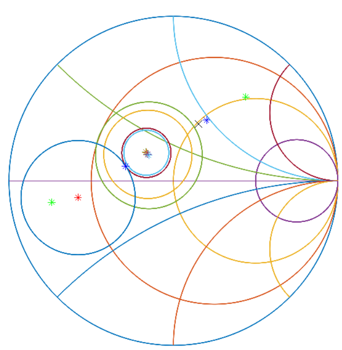


Figure 28: 13 dB Availible gain circle with corresponding Γ_S reflection coefficient added with blue asterix which rests on 13 db availible gain circle and 1.96 dB Noise figure circle. Corresponing Γ_L also added with a blue asterix

We provide a plot to illustrate how the availible power gain varies on the 1.96 dB Noise-figure circle and calculate the stub lengths and the serial lines

```
figure
plot(phi*180/pi, G);
zG=(1+gammaG)/(1-gammaG)
zL=(1+gammaL)/(1-gammaL)
dl_IN_noiseFig = stub1 (conj(zG), 'po' )
dl_OUT_noiseFig = stub1 (conj(zL), 'po' )
lambda_mm=1000*3E8/1.8E9
stub_in_mm = dl_IN_noiseFig(2,1)*lambda_mm
serial_in_mm=dl_IN_noiseFig(2,2)*lambda_mm

stub_out_mm = dl_OUT_noiseFig(2,1)*lambda_mm
serial_out_mm=dl_OUT_noiseFig(2,2)*lambda_mm
```

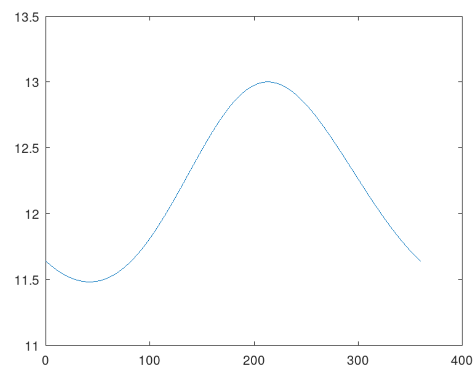


Figure 29: The variation of the availible gain with respect to the $Gamma_S$ reflection coefficients related to the 1.96 dB Noisefigure circle

Remark: It is the conjugate of Z_G and Z_S which are the impedances looking into the DUT which should be tranformed to 50 Ohms. We are always walking "Towards the load" on the Smitchart both for the load side and the source side. The problem is identical to a beginner's load matching problem except there one always traverses the Smitchart "towards the generator". Output

```
lambda_mm = 166.67
stub_in_mm = 15.015
serial_in_mm = 20.685
stub_out_mm = 19.825
serial_out_mm = 42.568
```

We've calculate the width of the microstrip and how to calculate ϵ_{eff} for the formula $\lambda_{eff} = \lambda / \sqrt{\epsilon_{eff}}$

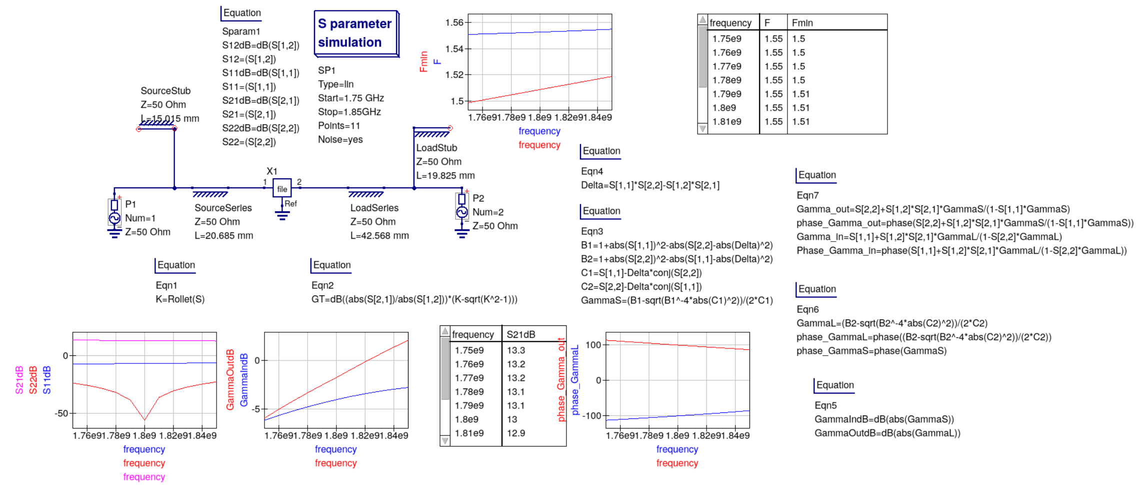


Figure 30: Design with theoretical TLINE transmissionlines

The formula for `GammaIndB` and `GammaOutdB` are not correct but had followed along from the conjugated matched design. They are corrected in the longer sweep below. The Noise-fige plot is not correct but should be higher. Minimum NF should be 1.9 dB at 2 GHz according to the Touchstone file

```
! Noise data:
! Freq.      Fmin         Gamma-opt         rn
  500        1.10      .330       27.0     .250
  900        1.25      .294       48.0     .260
 1000        1.30      .298       52.0     .270
 2000        1.90      .242      134.0     .160
```

The program locks often so we couldn't rearrange the graphics as to not overlap. Here we do a sweep from 1.5 to 3GHz

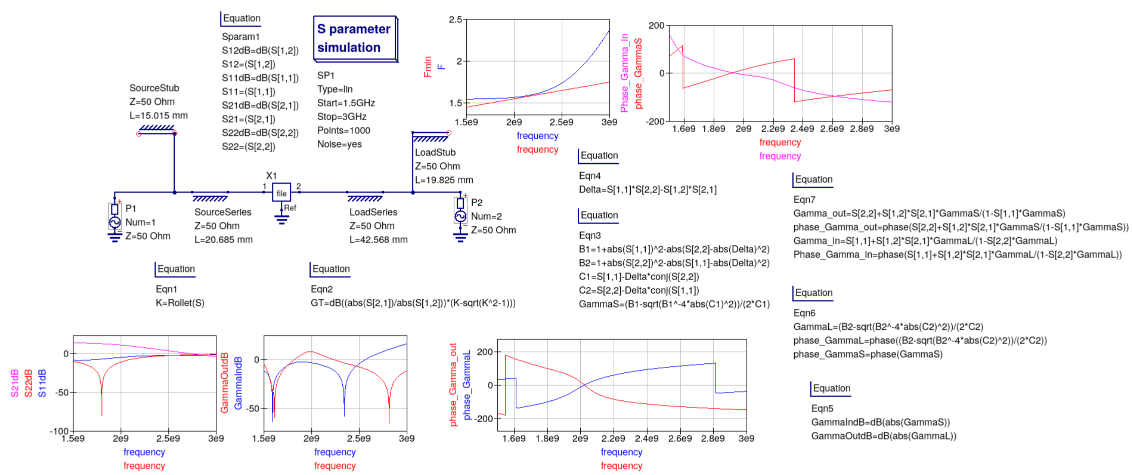


Figure 31: Design with theoretical TLINE transmissionlines. Longer sweep

Calculation of microstrip lengths

```
disp("Microstrips")
d=1.6E-3;
epsilon_r = 4.4;
losstangent = 0.0018
copper_thickness = 34E-6;
u=mstripr(epsilon_r,50);
e_eff= mstripa(epsilon_r,u);
lambda_eff_mm =lambda_mm/sqrt(e_eff)
Width = u*d

stub_in_mm = dl_IN_noiseFig(2,1)*lambda_eff_mm
serial_in_mm=dl_IN_noiseFig(2,2)*lambda_eff_mm

stub_out_mm = dl_OUT_noiseFig(2,1)*lambda_eff_mm
serial_out_mm=dl_OUT_noiseFig(2,2)*lambda_eff_mm
```

Noticing the difference in the wave length!

```
lambda_eff_mm = 91.315
Width = 3.0622e-03
stub_in_mm = 8.2268
serial_in_mm = 11.333
stub_out_mm = 10.862
serial_out_mm = 23.322
```

The Noise figure calculation is still rubbish but we have still have 13 dB gain

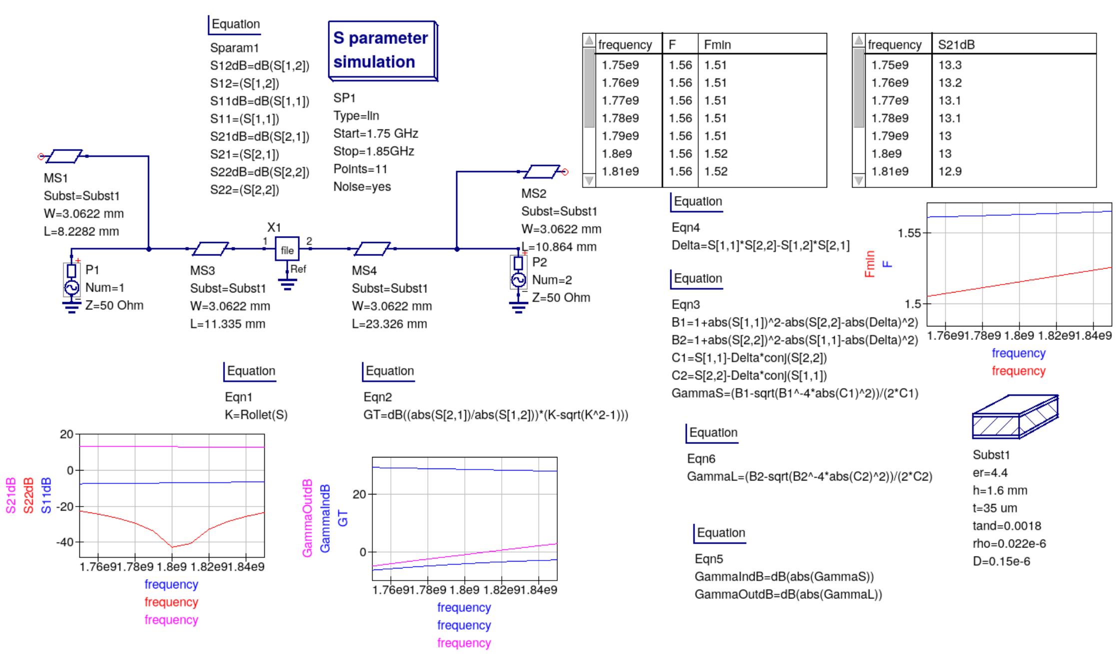


Figure 32: Design with microstrip lines

A longer sweep. Something unphysical happens with Γ_{in} at 2.1 GHz which is not reflected upon the composite S_{11}.

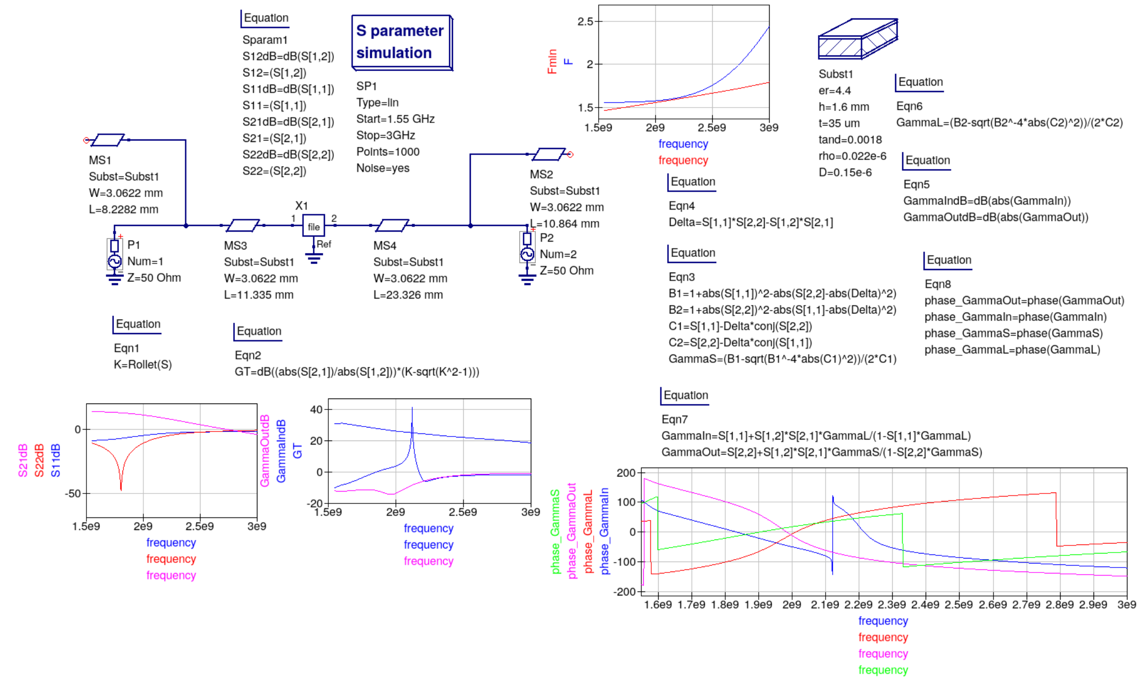


Figure 33: Design with microstrip lines. Longer sweep

Not much losses including Tee-bends

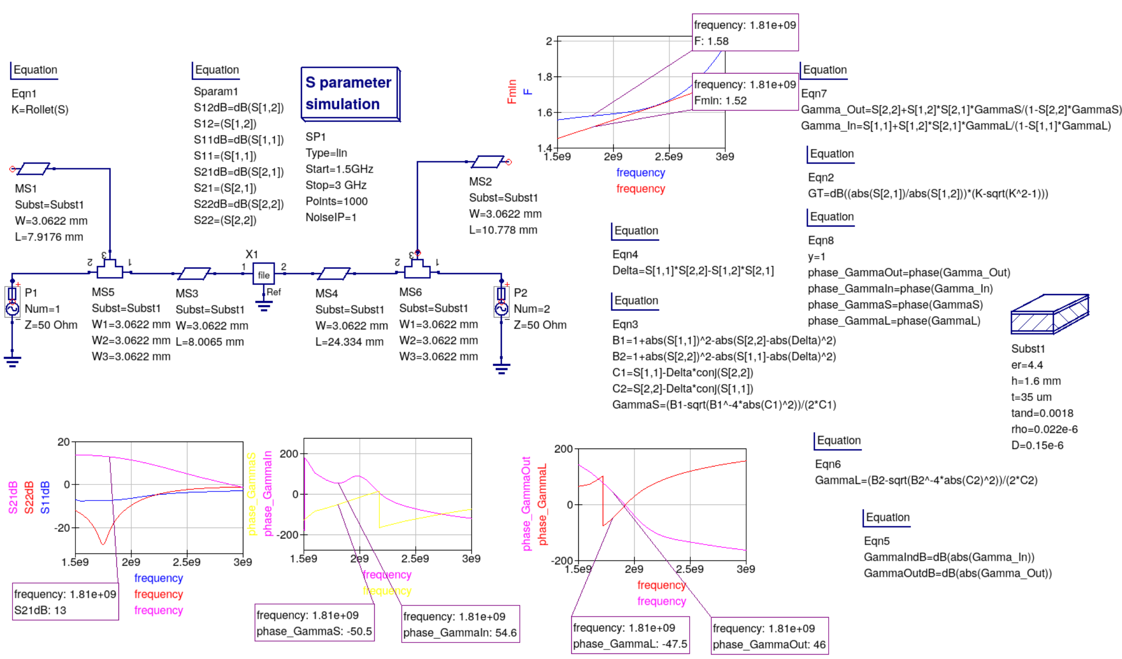


Figure 34: Design with microstrip Tees

What about if we simulate for integer values of the microstrips lengths?

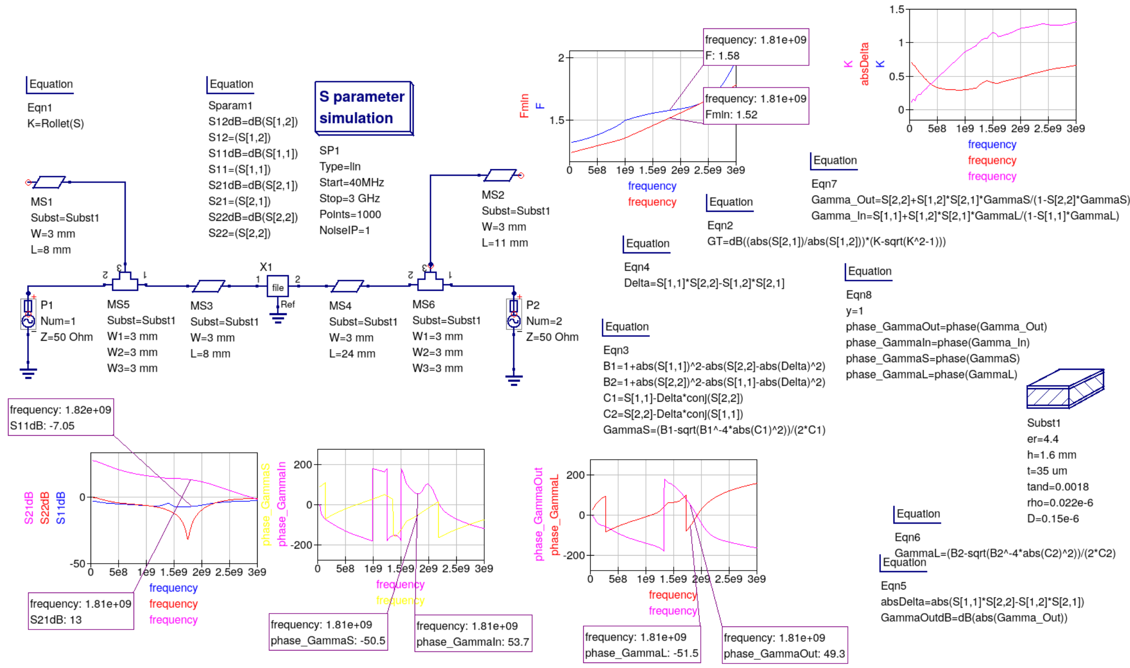


Figure 35: Design with microstrip Tees

Still not stable over all frequencies. Still have the correct gain. Returnloss is around 14 dB at