

## QFN 4x4 plastic package performance at E-band

*MCCX04 Master's thesis in electrical engineering*

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Göteborg, Sweden, 2011

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Cover:

*QFN 4x4 plastic package, here shown with lid. Silvia has designed this plastic package for E band frequencies of 81-86 GHz. For further information see chapter about earlier studies.*

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## Abstract

E-band has lately turned available for cell phone providers in Europe. Ericsson has today working mini links at these frequencies but they are expensive to manufacture. Here the QFN 4x4 plastic package is examined for E-band performance. This package meets Borås production line and is therefore cost effective to produce. Old measurements and simulations have been validated and compared and new matching networks for two sub frequencies have been designed. Solutions for the 71-76 GHz band and for 81-86 GHz band have been reached. The simulations show that a pass band with return loss less than 15 dB and insertion loss less than 1 dB can be achieved for both sub frequencies. Last these matching networks were simulated on a computer with higher accuracy. However the last validation simulations with higher precision did not confirm the realized pass band performance for the higher sub band.



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## **Aim**

The aim of this master thesis is to investigate the limitations of plastic packaging and air-cavity packaging as well as to validate the measurements done and compare them with simulations in HFSS and ADS. The thesis also describes the design of new matching circuits with improved performance at E-band.

## **Limitations**

The performed work is based on simulations in HFSS and ADS and comparison to available measurements. The variation analysis of the simulated designs is not part of this work.



## Background

In the past years the amount of data traffic in wireless networks, such as cell phone communication and mobile broadband, has grown denser. Today smart-phones, mobile broadband and electronic equipment are used by almost everyone in the western world. Demand for these services is growing rapidly throughout the industrialized world. Also in the developing countries cell phone communication is growing extremely fast. Many cell phone providers are at the moment preparing to launch 4G or LTE nets. The demand for higher download speeds and larger amount of data-traffic have now reached a point where the old cellular networks no longer can handle it. The cell phone carriers now have two options; to confine the amount of data-traffic a wireless equipment/person can use over a period or to invest in the networks to allow larger and higher data transfers in the future. All over the world large investments in cellular networks have started.

One way to increase the availability in the cellular networks is to enhance the data-traffic capacity between base-stations. There are today two ways to send high data bit information between base-stations, one is to use fiber-optic links, and the second is to use high capacity wireless links. The fiber-optic way is extremely expensive and difficult to implement. For example the optic fiber has to be dug down over sometimes great distances and in a city this is extremely difficult and expensive. Microwave links is therefore very beneficial, both in a cost and installation point of view. Ericsson today provides microwave links at up to 38GHz and these can handle data traffic of around 500Mb/s.

Lately a frequency range designated E-band has been made available for cell-phone providers. This frequency range is assigned to be used by microwave links. Earlier these frequencies were dedicated to military use only. The E-band consists of three sub-frequency ranges, 71-76GHz, 82-86GHz and 92-95GHz, see Figure 1. These frequency ranges offer high data transfer rates. Although no standardization has been clarified in Europe for these frequencies all providers aim to have products in these frequency ranges. Ericsson today has made mini-links working at these frequencies with a data rate of 2,5Gb/s, however these do not match Borås production lines and are therefore expensive to manufacture. The aim of this thesis is therefore to evaluate how standard plastic packages can be designed to work at these high frequencies.

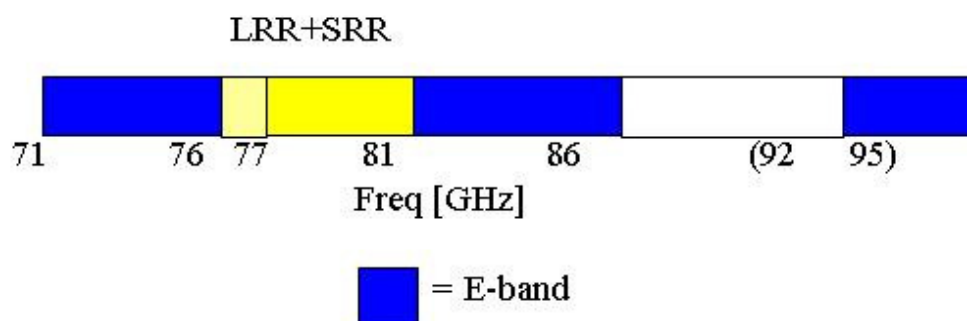


Figure 1. Frequency range of E-band, here shown with frequency spectrum of short range radar (SRR) and long range radar (LRR).

## ***Ericsson standard plastic package***

The Ericsson standard plastic package is used in all mini-links and the production lines in Borås are constructed to handle this type of package. Therefore it is very convenient and cost effective for Ericsson to be able to use the same package and machinery to build the transition at E-band. Since no one in the world has designed transition for plastic packages at these frequencies there is no standard way of doing it and therefore it is a huge competitive advantage to be first.

The plastic package is used in the MINI-link radios by Ericsson with integrated circuits. Different components can be attached on this package to enhance, modulate, and mix etc the ingoing or outgoing signal before transmitting it. Since all transmitters and receivers use high power amplifiers at transmit and receive end it is at utmost importance that this package work at these frequencies. The signal is too weak at receive end and it can not be mixed to a lower frequency before amplification due to losses. The signal has to be amplified at E-band level, 71-86 GHz. Therefore it is crucial that the package meets certain specification. It has to have low losses; a pass band that matches the ingoing and outgoing signal. No self resonance can be allowed when implementing amplifiers. The performance of the signal combined with the package is of utmost importance for the performance of the radio. Since the aim at Ericsson is to have gigabit capacity at E-band, the slightest error will cause the performance to deteriorate significantly.

# Theory

## S-parameters

The backwards reflected voltage from a load or impedance step divided by the incident voltage,  $V^-/V^+$ , is called reflection coefficient  $\Gamma$ . If a device has several ports a specific gamma can be denoted to each port. They are then designated specific names as S11, S22 etc where the numbers stand for the port, i.e. S22 stands for the backward reflected voltage at port 2 incident at port 2. Expanding this notation the voltage transferred from one port to another can be designated. S21 stands for the voltage transferred from port 1 to port 2. This is very convenient since it correlates all the ports of the device and can be put in a matrix form. I.e. the scattering matrix correlates all the incident and reflected waves. Plotting each of these parameters over a frequency range the behavior of a device can be readily shown. The S-parameters or scattering matrix contains all the information needed for a microwave device. The proper impedances for matching network on both source and load side can readily be found. In short the overall performance of a multi port microwave device can easily be displayed in a matrix form. (Collin, 2001)

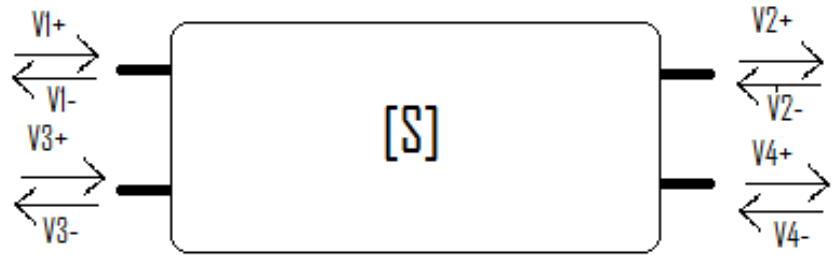


Figure 2. Schematics over S-parameters.

$$\begin{bmatrix} V_1^- \\ \vdots \\ V_N^- \end{bmatrix} = \begin{bmatrix} S_{11} & \cdots & S_{1N} \\ \vdots & \ddots & \vdots \\ S_{N1} & \cdots & S_{NN} \end{bmatrix} \cdot \begin{bmatrix} V_1^+ \\ \vdots \\ V_N^+ \end{bmatrix}$$

Figure 3. Scattering matrix representation.

Here it is very convenient to define insertion loss (IL) and return loss (RL). The insertion loss is the transmission loss when a signal travels from one port to another. In the same way return loss relates to the reflected voltage at a port. These two losses are unwanted over the pass band and they are usually minimized with proper matching networks. However all losses, both insertion and return, can not be removed in reality due to non perfect materials, production variations etc. Although they can be made very small. (Collin 2001)

## Modes

A signal propagating along a waveguide, coaxial line, micro strip line etc is traveling in a certain mode. The number of allowed modes is dependent on the dimensions of the waveguide, coaxial line or the microwave device which the signal is propagating on. Several modes can be present at the same time but usually a circuit is designed to

operate at one mode only. The tangential electrical field has to be zero on a conductive wall and thus a certain set of modes will arise. Two types of modes are present and are determined by if the transversal electric or magnetic field is zero: TE-modes and TM modes

The normal mode is usually the one to consider when designing a circuit. However other modes can be a problem. If the circuit is designed such that other modes are excited these can cause problems, unwanted frequency behavior, worsen performance over pass band etc. (Collin 2001)

## Standing wave

When a wave hits a load or impedance step, it is partly reflected if the load is not matched properly. This reflected wave and incident wave will interfere to create a standing wave pattern. Thus where the reflected wave and incident wave add in phase a voltage maxima occur and where they are out of phase a voltage minima occurs. Since the reflected wave usually has less amplitude than the incident wave the standing wave is never zero in amplitude. The exception is total reflection in for example an open or shorted stub where the reflected wave equals the incident wave in amplitude. According to Collin, 2001 eq 3.43 the standing wave pattern follows the following equation:

$$|V^+| \left[ (1 + \rho)^2 - 4\rho \sin^2\left(\beta l - \frac{\theta}{2}\right) \right]^{1/2} \quad (1)$$

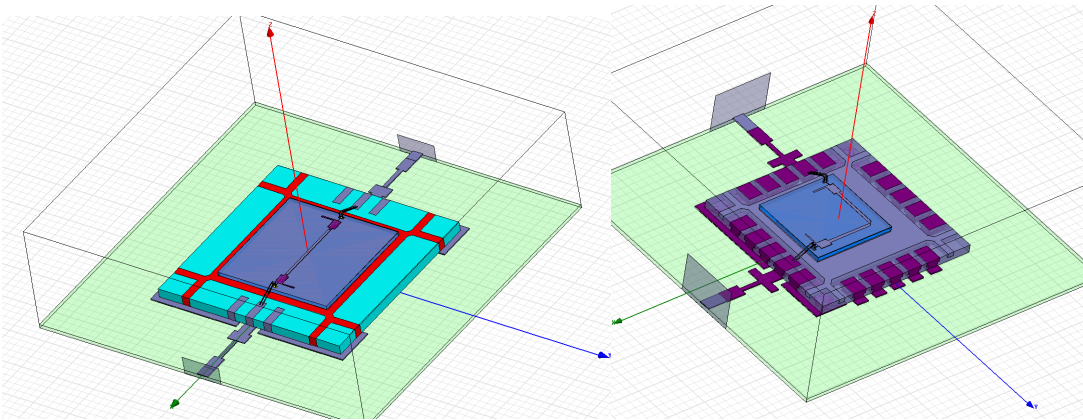
The result in equation 1 states that the total voltage at any point on the line oscillates. Maximum values occur when  $\beta l - \theta/2 = n\pi$  and minimum values when  $\beta l - \theta/2 = n\pi + \pi/2$  where  $n$  is an integer. Thus successive maxima and minima are spaced by  $\lambda/2$  apart where  $\lambda$  is the wavelength in the medium.

Standing waves can occur in almost any microwave device. The slightest impedance step can cause a reflection and thus a standing wave. Standing waves are unwanted since the power delivered to load reduces with the ratio of the standing wave. The larger the standing wave is, i.e. the more of the incident voltage that is reflected, the lower the performance will be. During measurements standing wave patterns are therefore interesting and an easy way to measure the performance of the plastic package. (Collin 2001)

## Earlier studies

The QFN 4x4 is a 20 lead quad flat no-lead plastic package. Each side is 4 mm long and has 5 leads. This package is one of many used plastic packages for integrated circuits. Several of these packages have been investigated for E-band performance in earlier studies. The most beneficial ones were found to be QFN4x4 20 lead and RJR 5x5. RJR5x5 will not be investigated in this thesis.

Silvia Cavalieri D'Oro, Milan, has designed a QFN4x4-package and a RJR5x5-package to operate at the E-band frequencies 71-76 GHz and 81-86GHz. Due to time restraints the main focus of this work will be on QFN4x4. The main part of Silvia's work was to design matching networks on both chip and carrier side. The work ended in a single matching network on chip side optimized for the 81-86 GHz band. Two different matching networks were designed on carrier side. One optimized for 81-86 GHz and one for 71-76 GHz. Thus the lower band of 71-76GHz did not have any own optimized matching network on chip side. Therefore the frequency response for the sub-band of 71-76 GHz will not be as good as for 81-86 GHz.



**Figure 4. The two designed packages. To the left the RJR5x5-package and to the right QFN4x4-package.**

Since Silvia's design on chip side was equal for all packages and sub-bands this design was quite large and broad band. This was very convenient since the only thing that had to be changed was the ingoing matching network on carrier side when switching sub band. The disadvantage of large spacing requirements is limiting the number of components that can be attached to the chip. Therefore it would be more desirable to have a smaller matching network on chip side.

Figure 5 will present most important parts of the QFN 4x4 plastic package. The green substrate layer is a Taclam substrate, a new material used for these high frequencies. It is used since it has a low dielectric constant and it can be manufactured very thin, 100  $\mu\text{m}$ . These plastic packages are usually manufactured with a lid which is not shown here for convenience. This lid will not be used in the simulation either. The dips in thru transmission are present without lid and thereby the cause of these dips is believed can be found without lid. The lid increases the required computer capacity and since the capacity is limited it is beneficial to remove the lid at this point. (Silvia, 2009 [1])

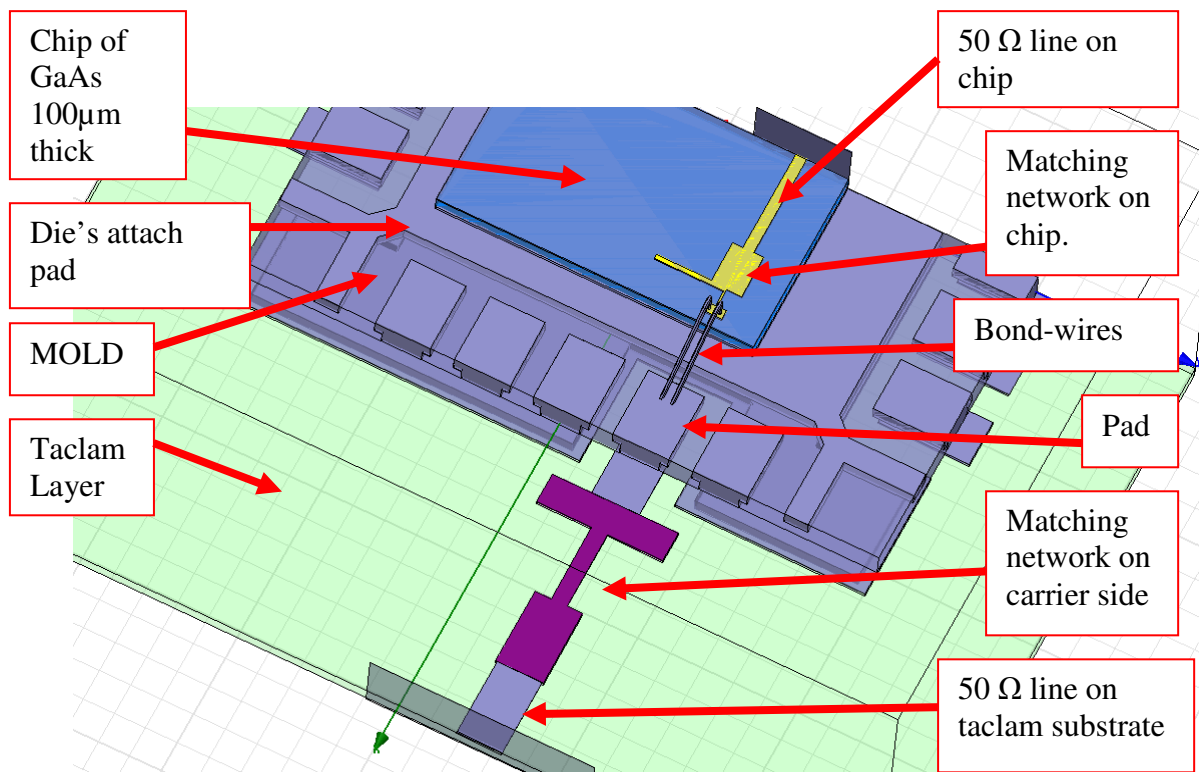


Figure 5. Matching networks designed by Silvia, here shown in a single transition simulation for 71-76 GHz.

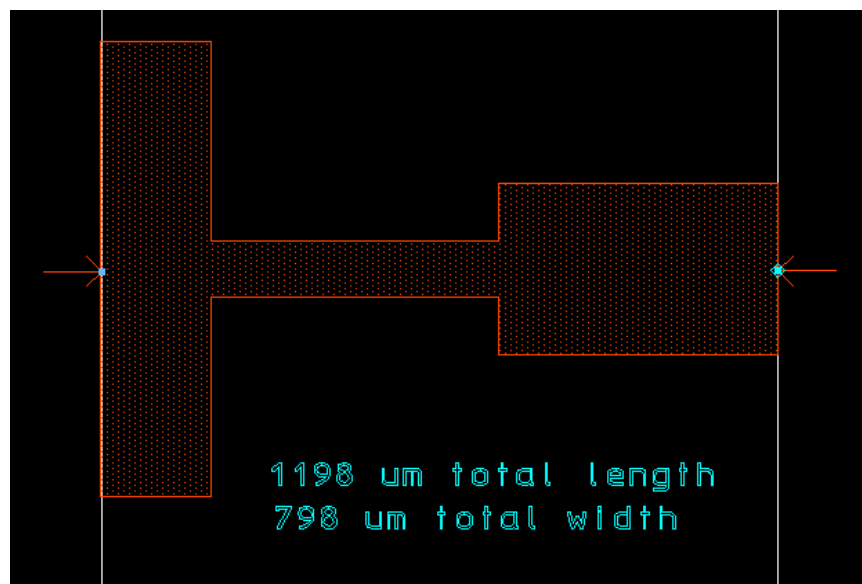


Figure 6. The matching network on carrier for 71-76 GHz designed by Silvia.

Silvia's design of matching network has one major drawback. The matching networks are very angular and therefore very lossy. Sharp corners tend to radiate power and this is highly unwanted. Since a good matching network with small losses is set as one of the design goal, less than 1dB insertion losses, each unmiterred corner will lower the performance.

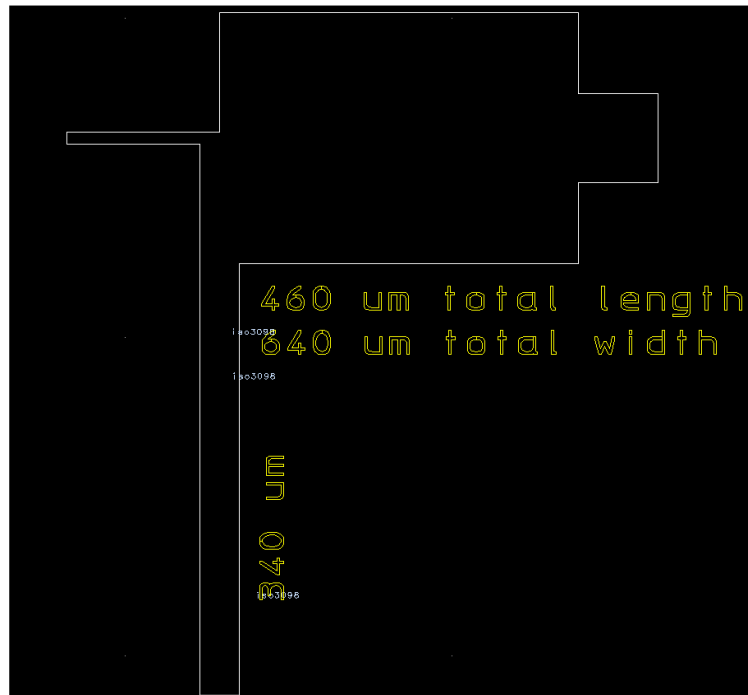


Figure 7. Silvia's design of matching network on chip.

### Differences between simulations and measurements

The simulations and measurements did not match as expected. There were problems with frequency dips and almost complete blockage at some distinct frequencies. Especially the performance at the lower sub band 71-76 GHz was deteriorating. At 70 GHz a dip occurs that causes almost complete blockage and has a strong influence on the pass band. This dip is present in all the designs, in both packages, QFN4x4 and RJR5x5, and both frequency bands. Since this dip occurs even in the single transition it is expected to be a result of the existing design and will therefore be investigated in detail. Note the frequency shift seen in Figure 8 and Figure 9.

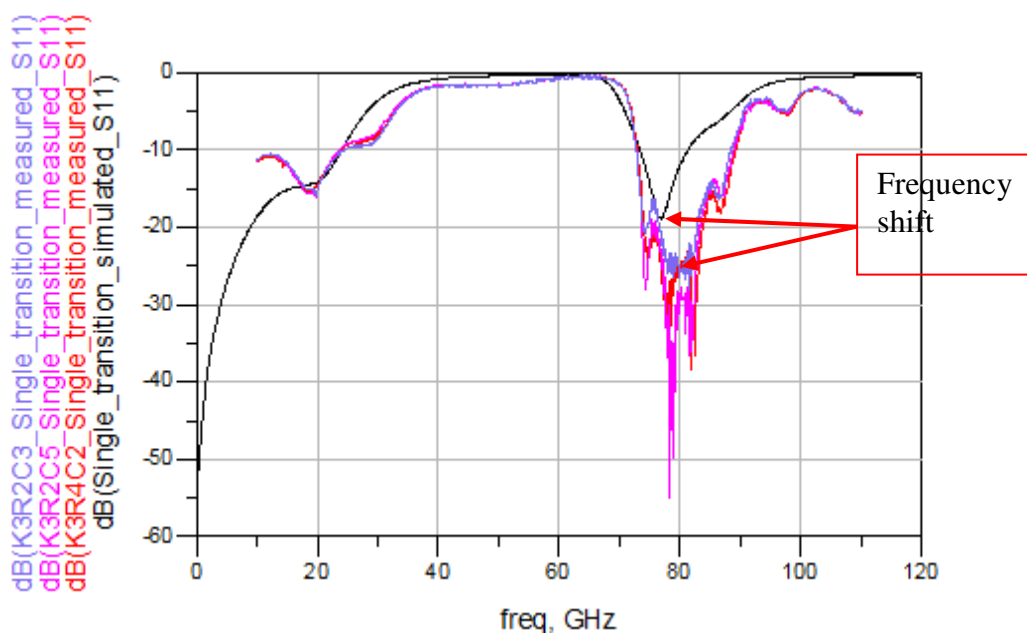


Figure 8. Single transition S11-parameter. The simulated response in black and three measured packages in red, pink and purple.

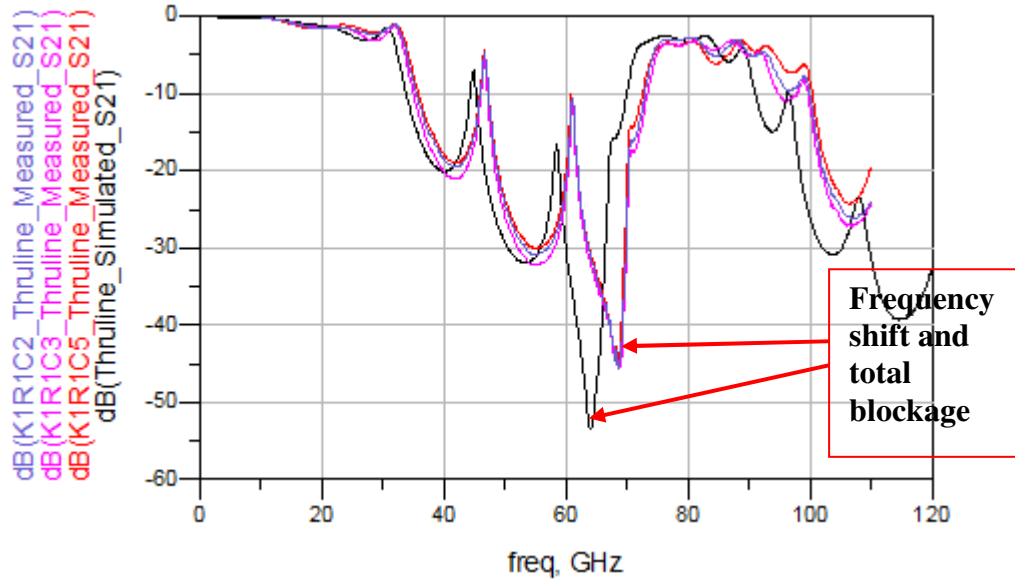


Figure 9. Simulated thru transition S21-paramter in black in red, pink and purple measured.

During the measurements some unfortunate circumstances occurred. One of the probes broke. Since now only one probe was viable up to 110 GHz a new probe had to be used. Unfortunately the new probe was only accurate up to 67 GHz. Since this frequency is well below E-band this became a problem. A calibration and testing of the new probe showed that it was “accurate enough” up to 80 GHz. Above this frequency the measurement result can only be seen as indications. Since the probe was expected to be viable up to 80 GHz the main focus in this thesis will be evaluating the lower sub band of 71-76 GHz. Thereby it is no use to start by validating the higher sub band. This damaged probe can also have influences on lower frequencies, how much can although not be established. As Silvia describes it in the report ref (Sivlia, 2009 [4]):

*“Poor tip calibration”. Since we did not have spare probes we had to improvise, and now tried with I67 A GSG-200 and adapters from 1.0 mm to 1.85 connector right before the probe, and a ceramic calkit for tailored for these probes, put on its absorber. These parts are specified only up to 67 GHz. This calibration seemed ok up to around 80 GHz, but above that we got large gain variation, of the order 1.5 dB, on the short. This problem could be due to the thick ceramic, large pitch, poor specification of the parasitic , and influence of leakage between probes (10 dB between shorts), etc. The thru did not show such pronounced problems. We believe that this calibration gives useful information, and results not far from those for other types of calibration”*

Because no measurements without poor tip calibration exists there is no way to know how large impact this type of calibration have. There is a known frequency shift from the measurements compared to the simulations. Whereas this shift originates from measurement errors or lack of accuracy in HFSS can not easily be determined without new measurements. It will however be investigated more thoroughly in this report. New matching networks will although be designed to match E-band specifications.

## **The measurements and simulation characteristics**

Silvia designed QFN 4x4 packages to meet certain specifications as return loss, insertion loss, band performance etc. However when these packages were manufactured and measured they did not match the desired specifications. Ericsson wants to know what and where it went wrong.

The design goal was to have a return loss of maximum 15 dB and insertion loss of maximum 1 dB over the pass band for a single transition. A single transition refers to a transition from the ingoing carrier to the chip, but not from the chip back to the outgoing carrier. A single transition is a very easy way to measure and visualize the performance of the pass band.

Next step is to investigate thru transition. The signal travels from the ingoing carrier up to the chip, down from the chip and to the outgoing carrier. By comparing this with single transition different kinds of unwanted frequency behavior can be detected. Since thru transition merely consists of two single transitions cascaded after each other with a micro strip line in-between they should only differ a few, i.e. 1-2dB in losses. Mode problems, standing wave problems etc could be an issue if there are more losses at certain specific frequencies.

The last test is to investigate the mode problem. A very easy and convenient way to do this is to test open and shorted stubs and compare the result with a thru transmission. An open or shorted stub easily excites new modes if they are allowed by the design. A wave that hits a shorted or open stub will be reflected and return back with our without a phase-shift of 180 degrees depending on open or shorted stub. Over the pass band open/shorted stub and thru transition should not differ by more than 1-2 dB. Any shortcomings will show up in the transmission through the package. Especially if at some distinct frequencies there are gaps between the stub performance and thru transition performance of several dB.

Depending on the outcome of this part the new matching network design process will start. If there are mode problems or otherwise unwanted performance behavior these should be more thoroughly investigated before trying to implement any new matching network. These problems, if any, are to be solved and then new matching network will be designed. First for the lower sub band of 71-76GHz, since it has the worse performance. If there is time left matching networks for the higher band 81-86 GHz will also be designed.

## Comparison of simulated and measured packages

We start with validating the old measurements and comparing them with simulations. The old reports did not contain any simulated data on designed packages. We had to resimulate them in HFSS. These simulations were then compared to the measured data. A brief presentation of the design will be shown here. Due to time and computer capacity restraints there were some limitations needed. Therefore the main focus will be the QFN4x4 package without lid and designed for the lower sub band of 71-76GHz. Mainly due to the worse performance of this band compared to the higher. At these frequencies the probing with poor tip calibration was believed to be accurate.

We consider the following designs:

- Thru transmission, Figure 10, the signal travels from the ingoing carrier up to the chip and then down from the chip to the outgoing carrier
- Single transition, Figure 11 to the left, the signal travels from incoming carrier to the chip level. This is the design for which the matching networks were designed for and the pass band of this design should determine the overall performance of the package
- An open and shorted stub design Figure 11 to the right. The stub design is used for comparison with thru transmission parameters to visualize any mode problem.

These designs will be compared in the first part of this report, with measurements to investigate if there are any obvious differences between the simulations and measurements.

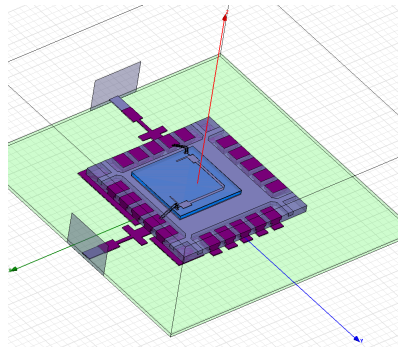


Figure 10. Thru transition for the QFN4x4-package.

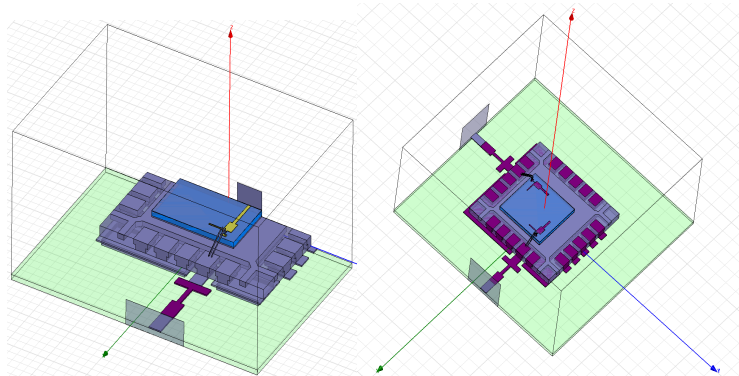
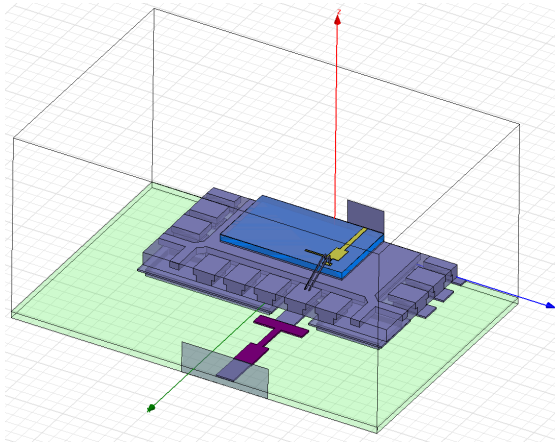


Figure 11. Single transition to the left, open and shorted stub to the right.

## Single transition



Single transition is presented here; the S-parameters are shown for three measurements and one simulation. Three measured packages are presented to exclude any misaligned or otherwise erroneous manufactured packages. Note the measured S11 parameter in Figure 12 is not a genuine single transition, due to two dips occurring on both side of the primary band. From the figures, the distance between the markers m1 and m3, a clear frequency shift is seen and is estimated to be around 2.4 GHz. This shift is seen with

all tested packages but no origin for this error has been found yet. Poor tip calibration as described earlier could be a source.

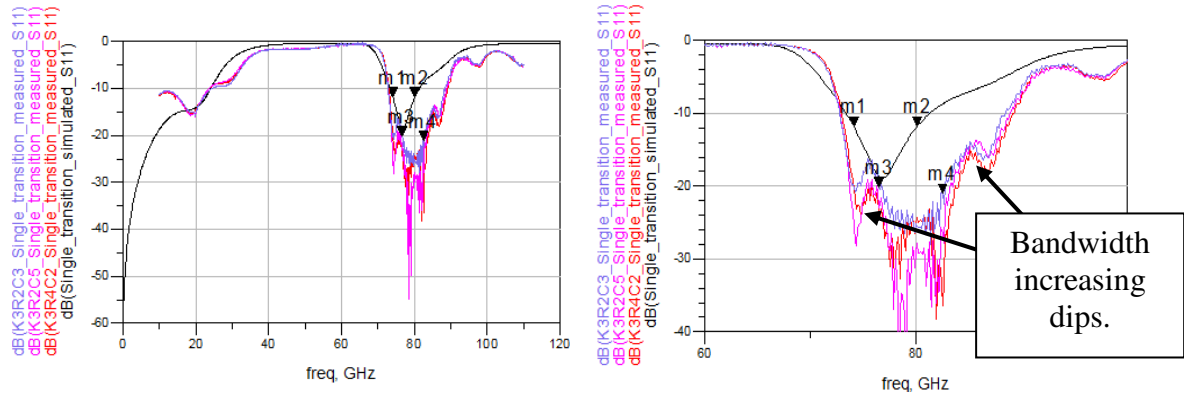


Figure 12. S11 for single transition, measured in red to purple, simulated in black. Note the frequency shift.

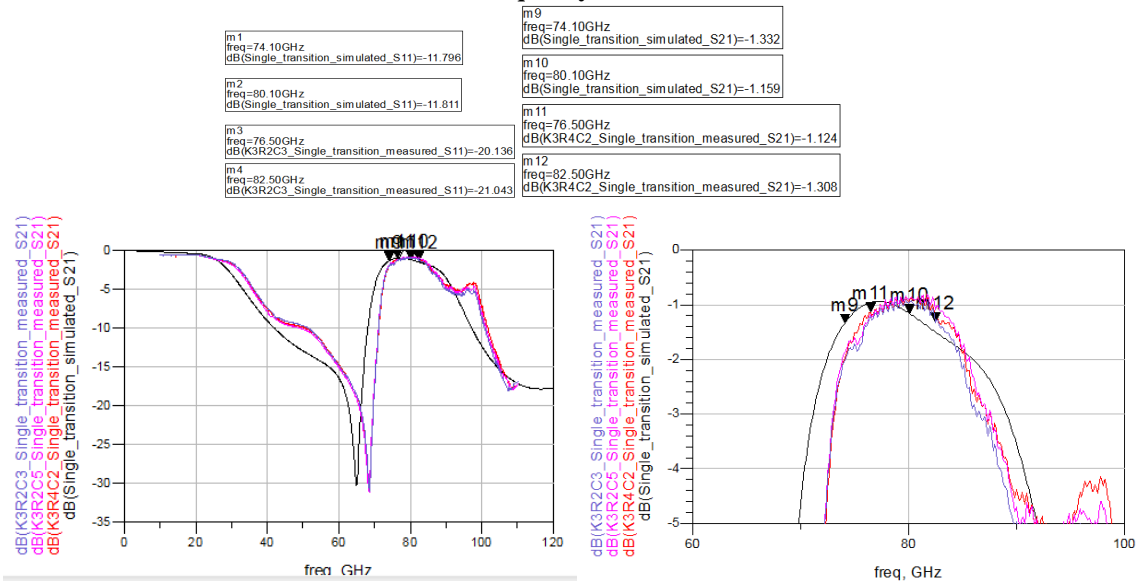
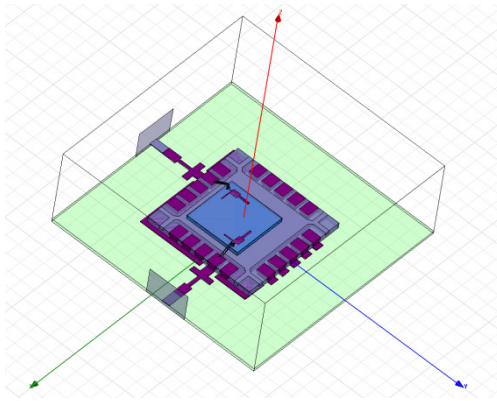


Figure 13. S12 for single transition measured in red to purple and simulated in black.

The simulated and measured packages have different frequency response and can not be compared to each other straightforward. Since the pass band is shifted upwards for the measured packages compared to the simulated, different areas have to be chosen for the

pass band in order to compare them. The simulated packages were designed, or at least intended, to have a pass band of 6 GHz (71-76 GHz sub band). A study of Figure 12 and Figure 13 show that the center frequency for the simulated package is located at 77 GHz. Therefore a new band from 74.1-80.1 GHz is chosen for comparison. For the measured packages the pass band is chosen in the same way. A 6 GHz band between 76.5 GHz to 82.5 GHz is used for comparison. The delimiting frequencies chosen to be of equal distance from the center frequency and to have similar values of return loss. Choosing the band for the measured packages so that the bandwidth increasing dips are outside the pass band, i.e. without possible moding problem. The possible moding is seen as the two extra dips just outside the frequency markers in Figure 12. The manufactured packages are much broader banded than the simulated package. The reason for that is unknown.

## Open and shorted stub



We present here the comparison between simulated package and three measured packages of open and shorted stub. The markers indicate the same frequencies as described for single transition. No major difference between the simulated and measured packages other than the obvious frequency shift of 2.4 GHz can be noted. The dips are deeper, for open stub up to 30 dB, in the measured packages compared to the simulated package. This difference is mainly noticed for the open stubs.

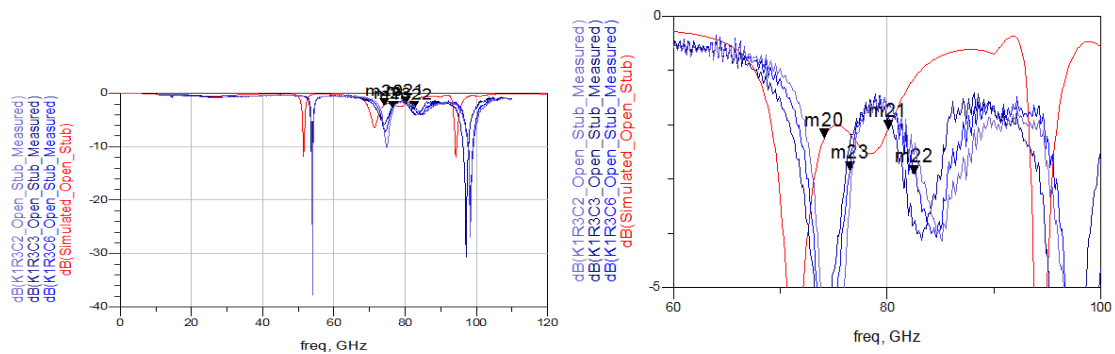


Figure 14. Open stub compared to thru transition.

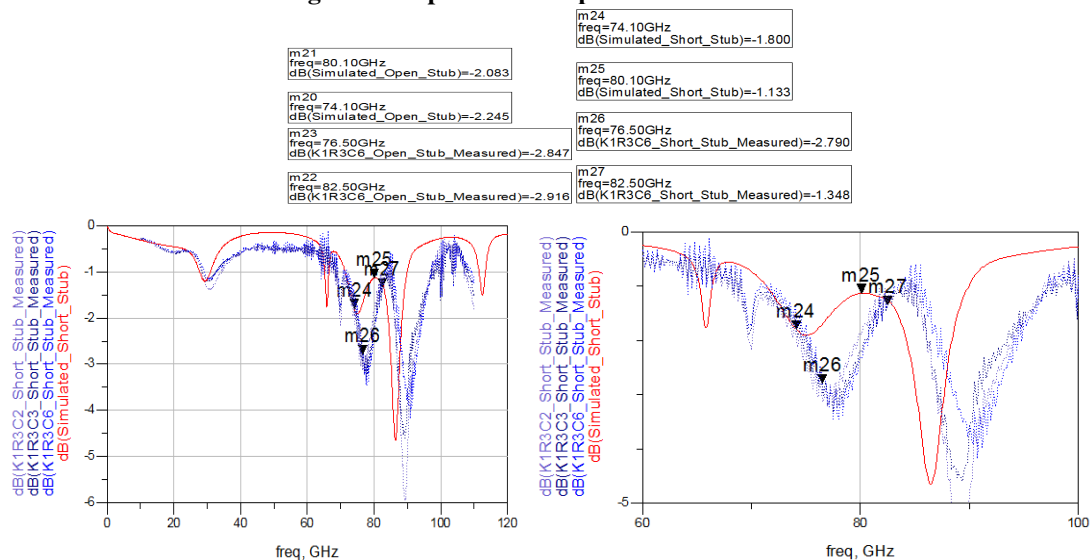
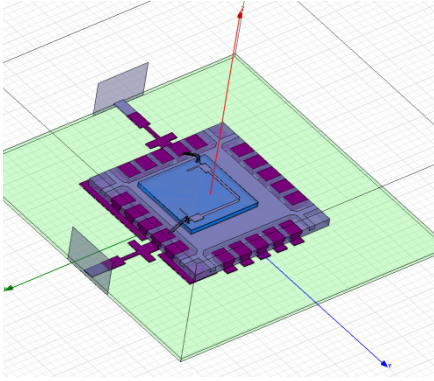


Figure 15. Shorted stub compared to thru transition.

No difference can be noted other than the frequency shift for the shorted stub. The matching between simulated and measured packages seems very robust. Here it should also be noted that open and shorted stub are not identical over the pass band. Open and shorted stub should in theory only differ by a phase shift of 180 degrees. Here as seen they do not match, why is unknown at the moment.

## Thru transmission parameters



We present here the comparison between simulated and measured packages for thru transmission. The simulated and measured packages seem to match very well. The markers indicate the same frequencies as for single transition and the upward frequency of 2.4 GHz shift is clearly seen for S21. This shift is equal to the one noted for single transition. Thereby it can be concluded that most likely it is the simulations from HFSS that are erroneous in frequency. Since the frequency shift is

identical in all the designed packages and over the entire measured frequency range most likely HFSS is not accurate enough. However this can not be certain, calibration error from the measurement set up and influences from the broken probe can not be excluded.

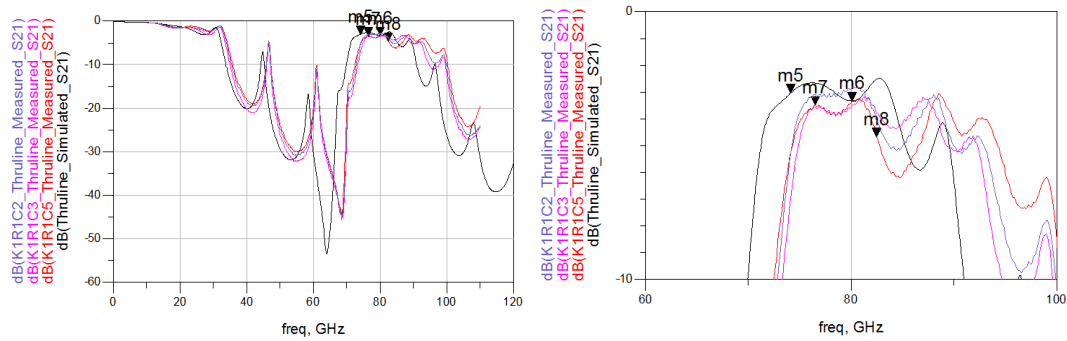


Figure 16. Comparison between S21 for simulated and measured packages.

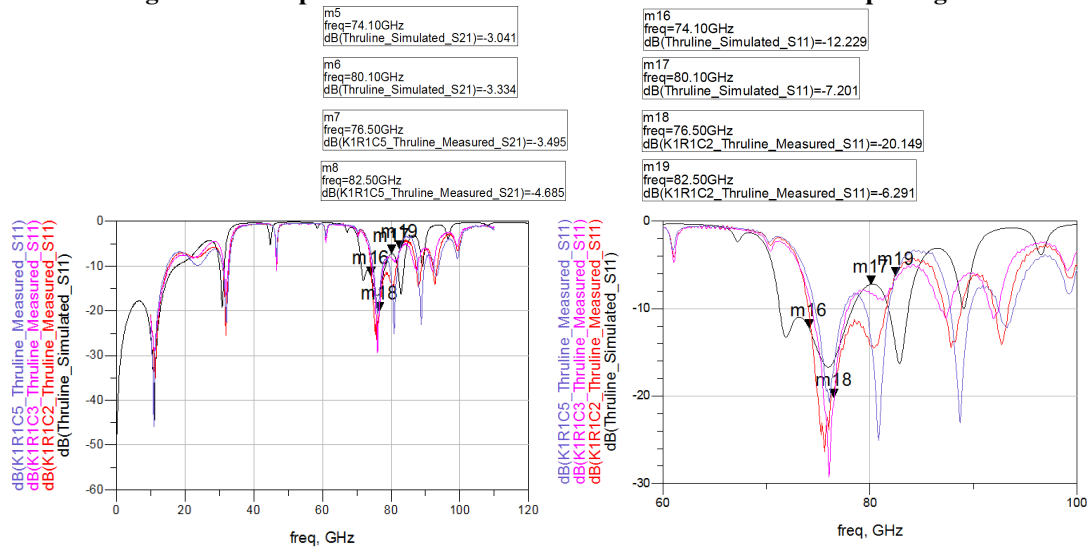
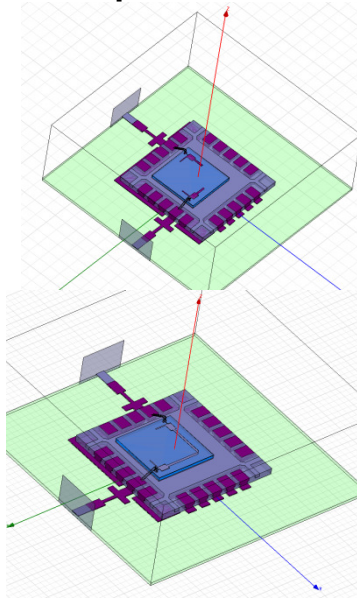


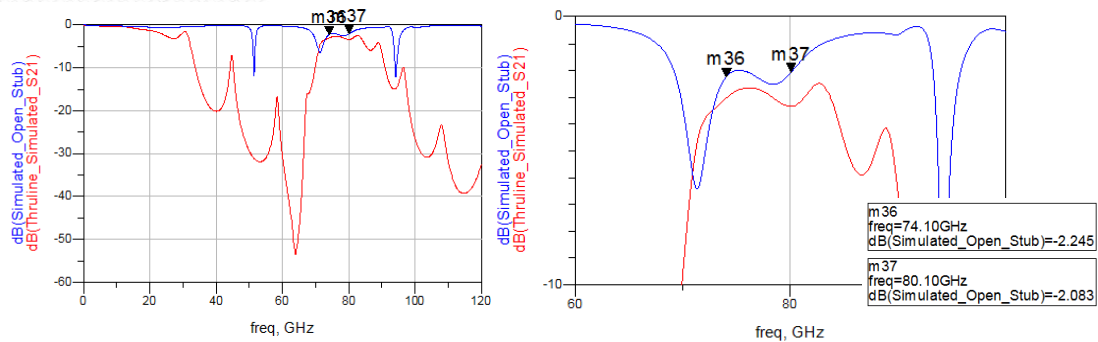
Figure 17. Comparison between S11 for simulated and measured transmission coefficients.

The ripple effect as seen in Figure 16 and Figure 17 differs between the measured and simulated packages over the pass band. They also differ between the measured packages. Otherwise a very good match is noted.

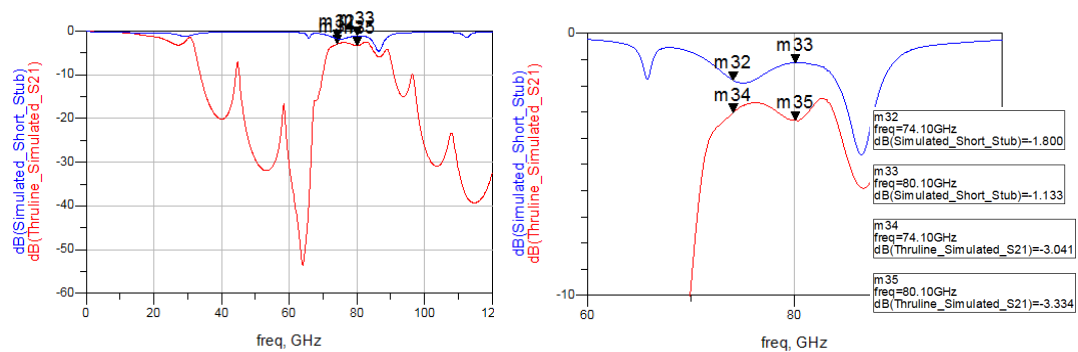
## Comparison between open shorted stub and thru transition



A convenient way to investigate the mode problem is to compare thru transmission with open and shorted stub. Over the pass band these different transitions should match well. There should only be a few dB in difference since an open or shorted stub essentially is a thru transmission. A signal travels from the carrier up to the chip out in the stub and is reflected back down from the chip to the carrier again. This is a thru transmission without some extra microstrip length on the chip and a possible phase shift if the stub is shorted. Mode problem could be present if they do not match. The match over the pass band for thru transition and shorted stub seems very similar. There is just a few dB in difference which is expected since the thru transmission also includes losses in the micro strip line on the chip.



**Figure 18. Open stub is compared to thru transition. The difference is just 1-2 dB over the pass band.**



**Figure 19. Shorted stub and thru transmission is compared. The difference over the pass band is just 1-2 dB.**

We can see that no mode problem is present effecting the pass band within the package. The match over the pass band is within 1-2 dB and must be seen as acceptable losses in the micro strip line on chip.

# Measured comparison of open, shorted and thru transmission

The measured thru, open and short stub were plotted in the same way as the simulated ones to investigate if there are any differences Except from the frequency shift there does not seem to be any obvious differences to observe.

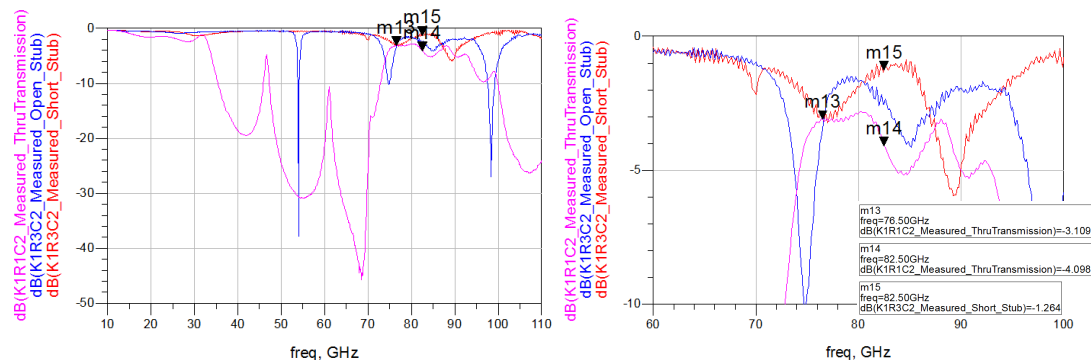
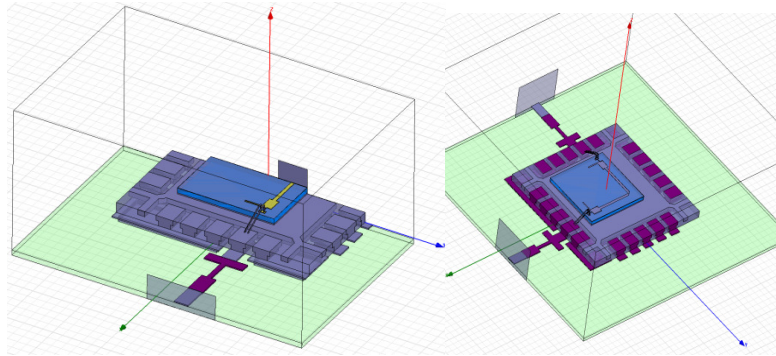


Figure 20. Open and shorted stub plotted with thru transmission.

## Comparing thru and single transition



Comparing single transition and thru transition can reveal losses within the pass band that occurs in a thru transmission but not in single transition. The ripple effect could for example make thru transmission behave much better than it in fact is. It is clear from Figure 21 that there is only a loss of a few dB within the desired pass band.

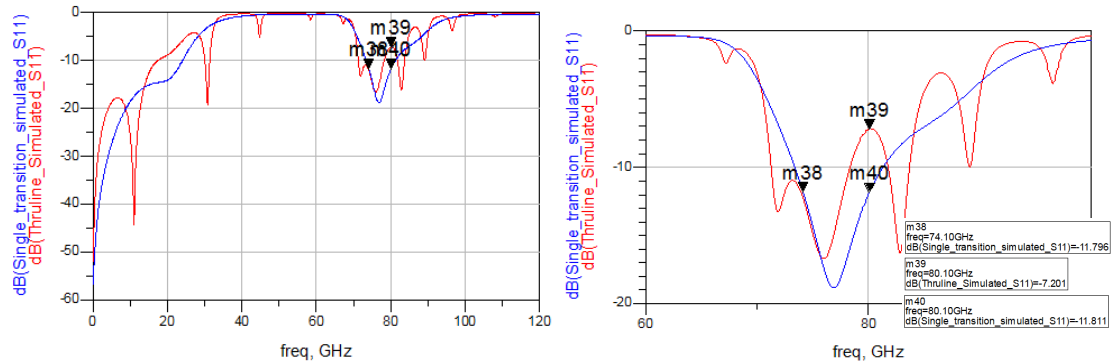


Figure 21. Single transition and thru transition compared, S11-parameter.

As seen single transition follows thru transmission well. There is a dip around 65 GHz which most likely originates from the capacitive adding, i.e. the stub, of the matching network on the chip. Within the pass band the loss in thru transmission compared to single transition is just 1-2 dB, see Figure 21 and 22. These are acceptable losses for the transition in the micro strip line to the chip. The origin of the ripples will be investigated further later on in this report see page 20.

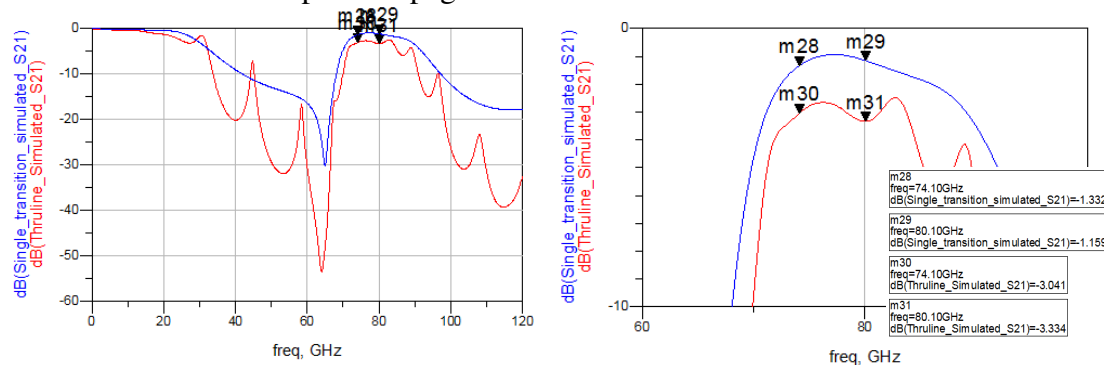


Figure 22. Single transition and thru transition compared, S21-parameter.

# Origin of the frequency dips

## Removing Y-corners

One possible cause for the frequency dips in the thru transition was believed to be a standing wave in the die's attaching pad. The Y-formed corners of it were known to have impact on the frequency response of plastic packages at lower frequencies. Therefore Ericsson wants to know if these corners have any impact on E-band frequencies. Packages without these corners were designed and simulated to test this. The main results are presented here together with packages with Y-corners.

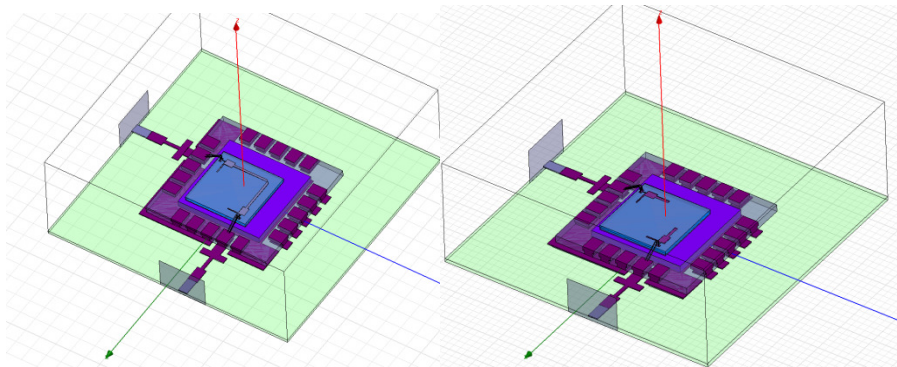


Figure 23. 71-76 GHz design without Y-corners. Thru line, open and short

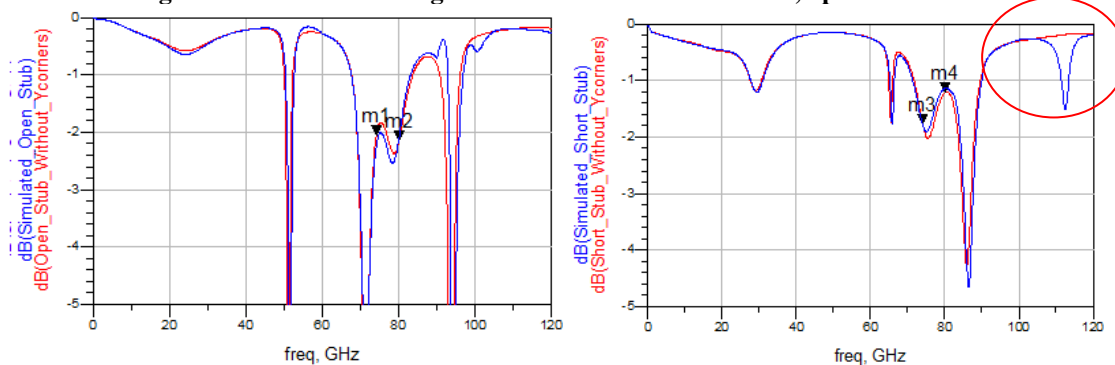


Figure 24. Comparison between original design (in blue) and design without Y-corners (in red). Open stub to the left and short to the right. Note the difference for short stub at higher frequencies. Markers indicate 74.1 GHz to 80.1GHz.

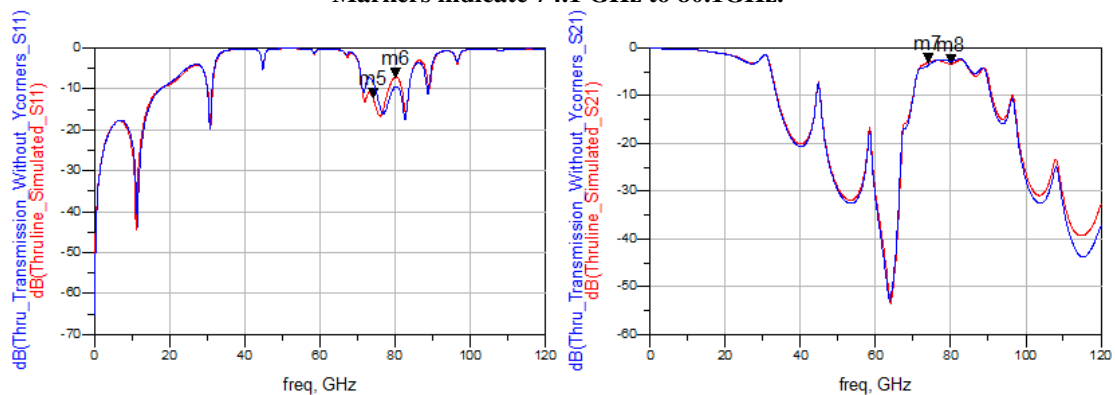
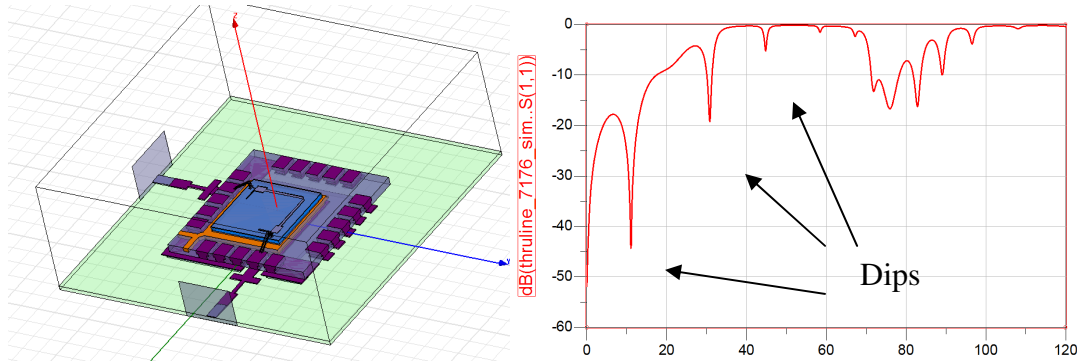


Figure 25. Comparison between original design (in blue) and design without Y-corners (in red). S-parameters for thru-transmission shown here. S11 to the left and S21 to the right. Markers indicate 74.1 GHz to 80.1GHz.

The only obvious difference is the missing dip for shorted stub in Figure 24. The conclusion must be that the dips are not caused by the Y-corners. Neither do they have

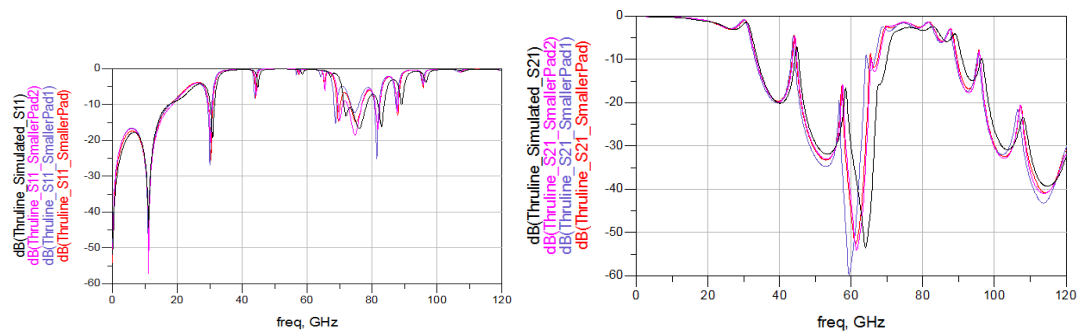
any impact on the frequency behavior in general of the plastic packages at these frequencies.

## Search for the dips within the die attach pad



**Figure 26. Modified die attach pad note that three of the four Y-corners have been removed. This is not believed to have any impact on the result of the simulation. Die attach pad in orange.**

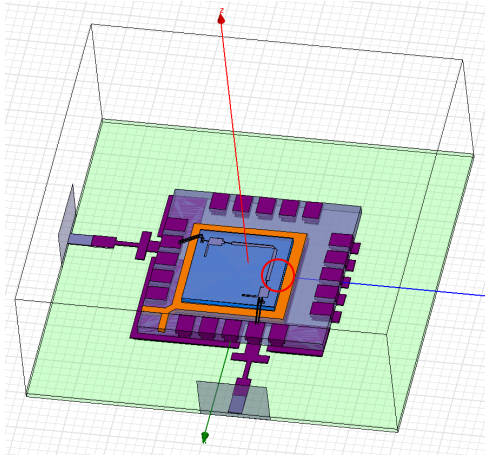
Excluding the possibility of the dips originating from the die's attach pad several packages were designed. These packages had different size on the die's attach pad, i.e. length width etc. These were simulated in HFSS. The results are presented here and show no sign that the frequency dips would originate from the die's attach pad. The small frequency shift that can be noted is not significant enough to explain the dips in thru transition.



**Figure 27. As seen the dips do not change significantly wit smaller pad. The markers indicate the region 73.4-81.3 GHz**

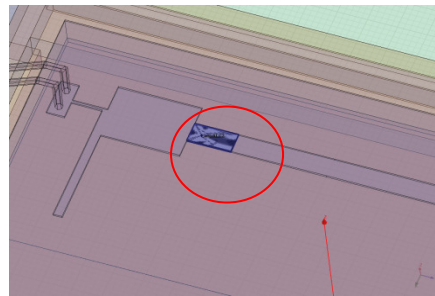
As seen in Figure 27 the dips seem more or less constant. The most likely explanation for the shift is not sufficiently accurate calculations in HFSS due to computer restraints. However this is not believed to have any major impact on the simulated frequency response. The conclusion must be that the dips does not originate from the die attach pad nor the GaAs layer. More simulations were performed but all indicate the same behavior. All of these are not shown here due to practical reasons.

## Damping the standing wave

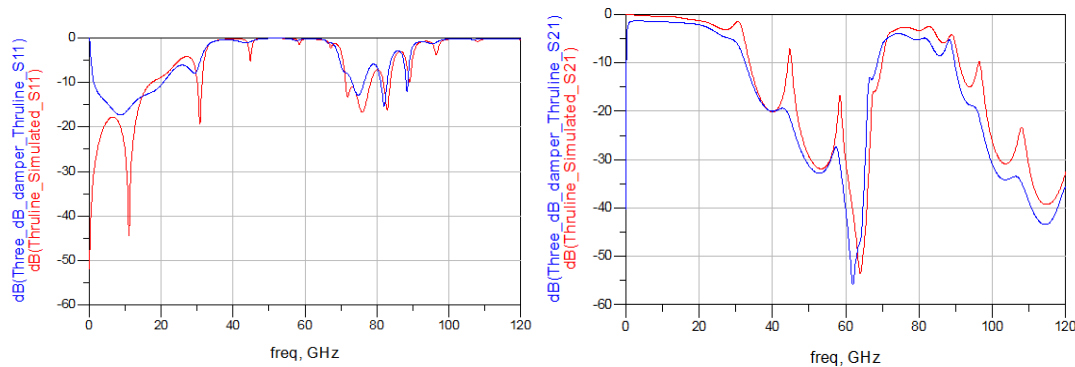


**Figure 28.** Inserting a 3dB purely resistive damper to investigate the behavior of the dips.

A 3 dB attenuator is inserted after the matching network on the chip to further examination if the frequency dips originate solely from a standing wave between the two ports. If only the amplitude of the dips changes and not the frequency where they occur, the conclusion must be that they originate from a standing wave. This is very convenient, since then nothing else within the QFN4X4- package has an impact on the S-parameters. This implies that the performance of the package at these frequencies can be improved without any unwanted effects or changes of the die attach pad.



**Figure 29.** Zooming of the resistive adding. Realized with a layer of higher resistance to accomplish a 3 dB damper.



**Figure 30.** As seen there is a clear damping when a pure resistive damper of 3 dB is inserted.

No major frequency shift is present as seen in Figure 30. There is only a difference in amplitude. Since the amplitude changes we can conclude that the dips originate from a standing wave between the two ports. These tests give that the QFN4x4 package can be used without any risk of frequency dips originating from the die attach pad. The dips solely occur from a standing wave between the ingoing and outgoing port. This implies, for QFN 4x4 plastic package, that the only crucial performance issue is the matching network.

## Attempting to match single and thru transition

It is interesting to know if two cascaded single transitions with an addition of a MLIN, microstrip line, between them will behave like a thru transmission. Since a thru transition essentially consists of two single transitions cascaded with an addition of microstrip between them it should be fairly easy to model it in HFSS from one single transition. The reason for this investigation is to match the thru simulation from ADS with the simulated thru transmission from HFSS. The cause for the ripple effect can be confirmed to be a standing wave between the two ports if this is the case. This test is a direct sequel of the examination of the die's attaching pad. Since that test indicates a standing wave between the two ports and not in the die it should be easy to make two single transitions match a thru transition. In the setup, Figure 31, the length of the MLIN was tuned to match the simulated thru transmission. The result is seen in Figure 32. A MLIN value of 2.35mm seems to provide a very good match.

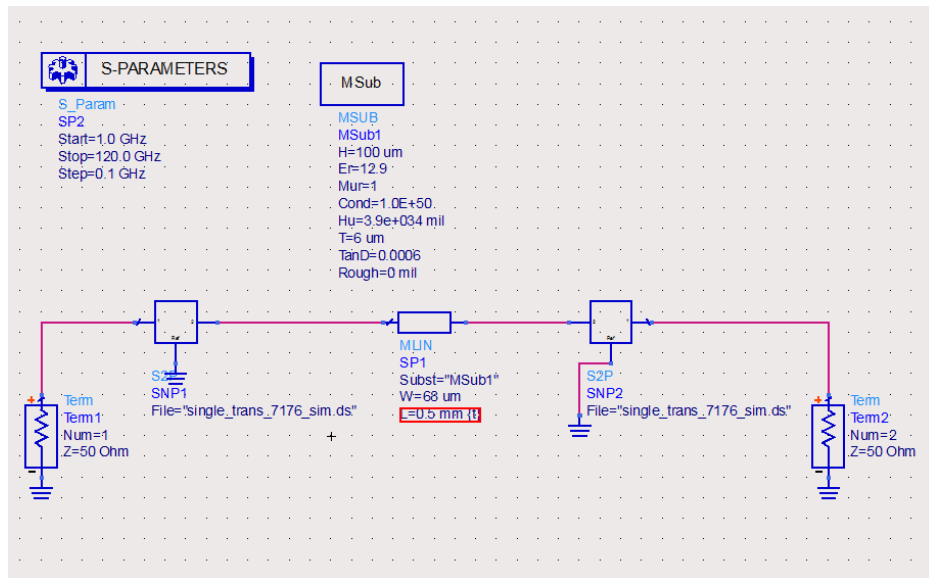


Figure 31. Tuning the MLIN length to make single and thru transition to match.

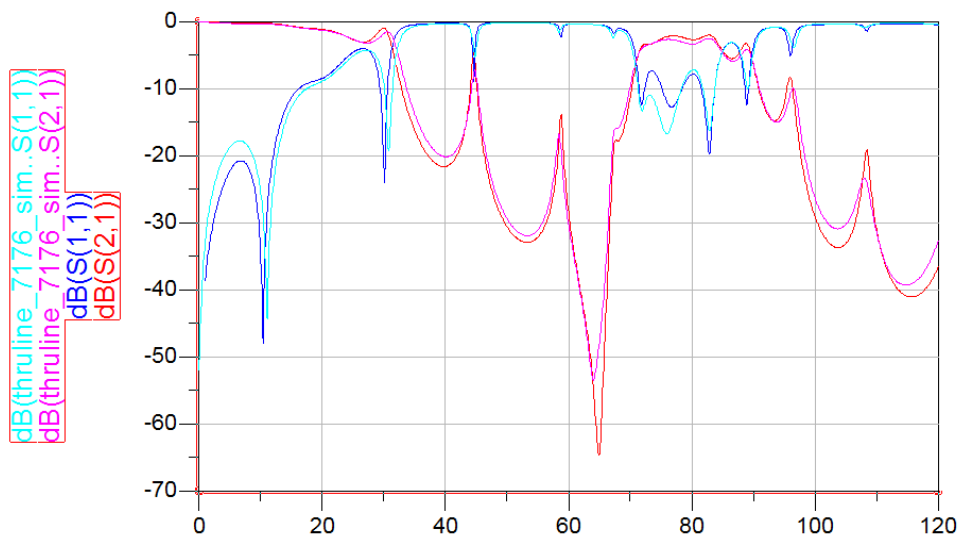


Figure 32. With a micro strip length of 2, 35 mm we get a match. Red and blue is tuned; purple and pale blue is simulated from thru line in HFSS.

## A standing wave between the two ports

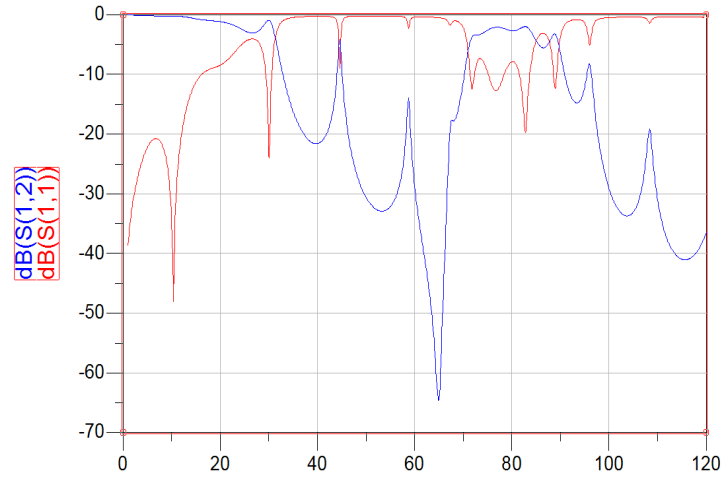


Figure 33. Here two single transitions are cascaded with a MLIN length of 2.35 mm.

There does not seem to be any indication of problems connected to the die's attaching pad. The reasonable alternative is that the dips arise from a standing wave between the two ports of the package. To confirm this some packages with longer thru lines on the chip were designed in HFSS. These were simulated and then matched in the same way as describe before. As seen in the Figure 34 the location and number of dips change significantly with micro strip length. The only stationary dip is the large one caused by the matching network on chip, a matching pole, marked by an arrow.

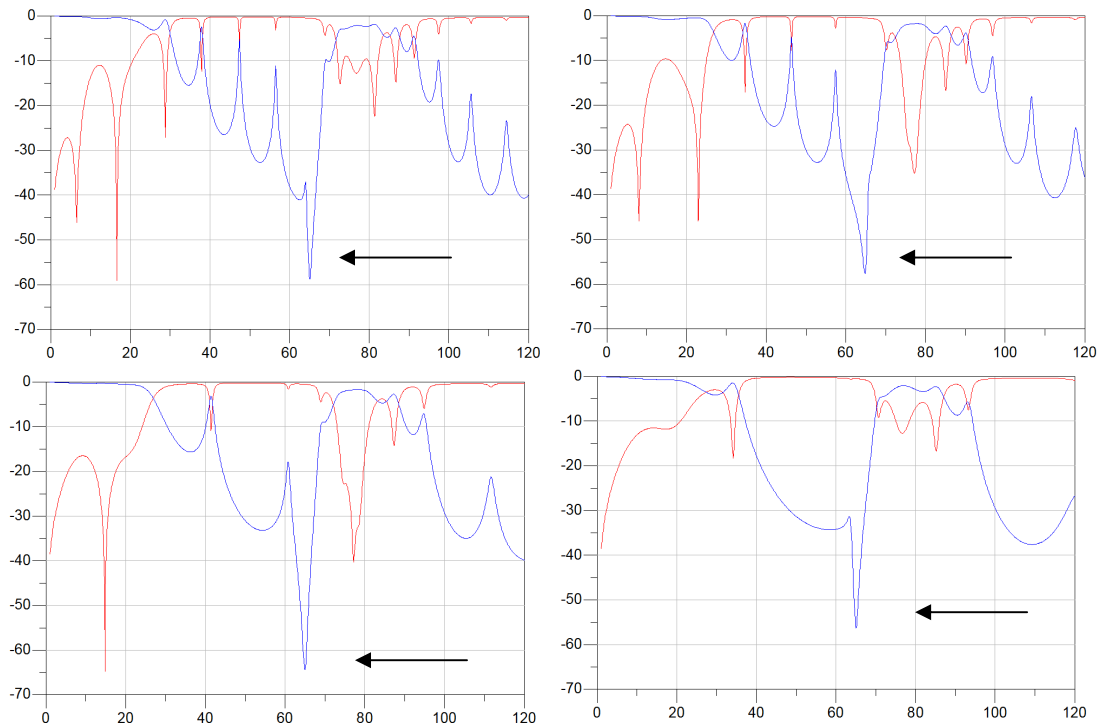


Figure 34. Here is two single transition with a MLIN length of 4.35; 3.35; 1.35; 0.35mm is cascaded. As seen the dips are changing significantly.

## Dips in open and short stubs

Open and shorted stub were recreated in ADS to examine where the dips originate from. Since the dips in thru transmission have been explained by a standing wave the most likely reason for the dips in open and shorted stubs ought to be the stub length. First a single transition was imported into ADS. Thereafter subtracting the micro strip length on the chip, this length could also have been subtracted directly in HFSS by using the same tool. Here the tool was not used to examine if the length subtracted in ADS an HFSS match. After the subtracted micro strip line an open or shorted stub was added, see Figure 35. The length of the stubs was then tuned to match the simulated stubs from HFSS. An almost perfect match was achieved see Figure 37. The conclusion must be that the dips are only dependent on the stub length. To verify this, a 3dB attenuator was added in ADS directly after the transition from the carrier. This attenuator did not show any signs of having any impact on the dips. The conclusion must be that the dips are caused by the length of the stub and nothing else. Also note that the negative length of the subtracted micro strip line matches the length that should have been deemed in HFSS. The length of the stubs also matches the real length of the stubs in HFSS.

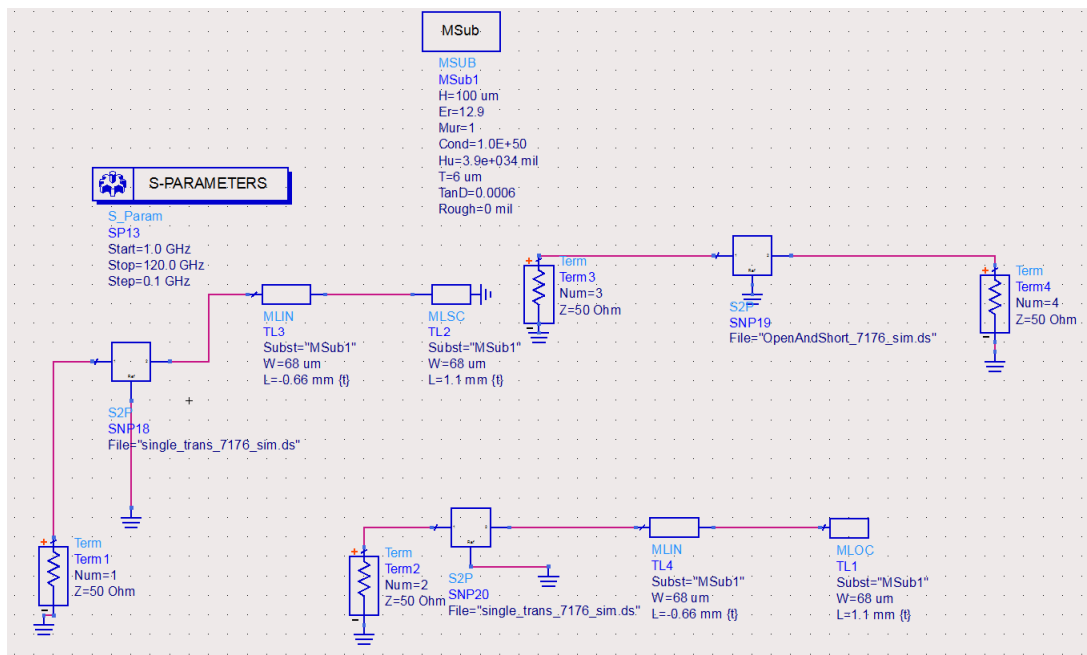
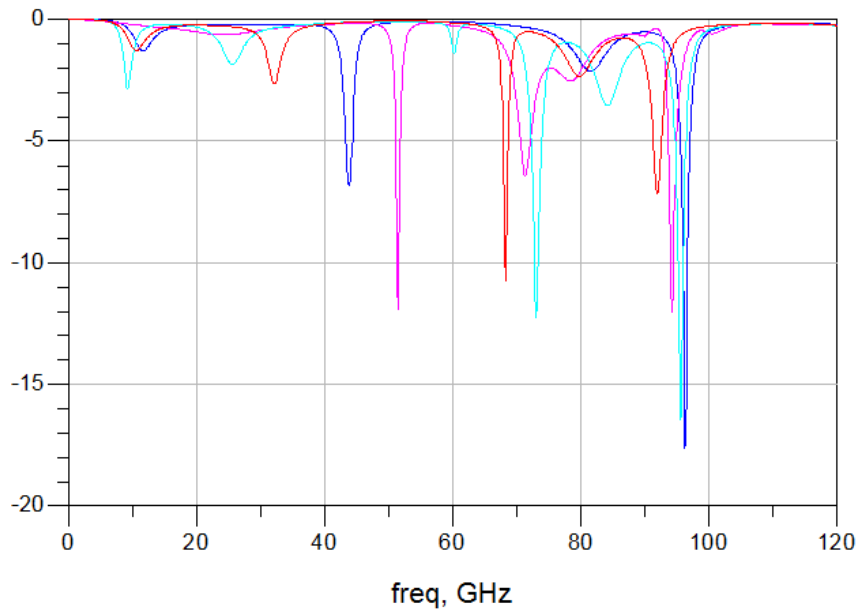
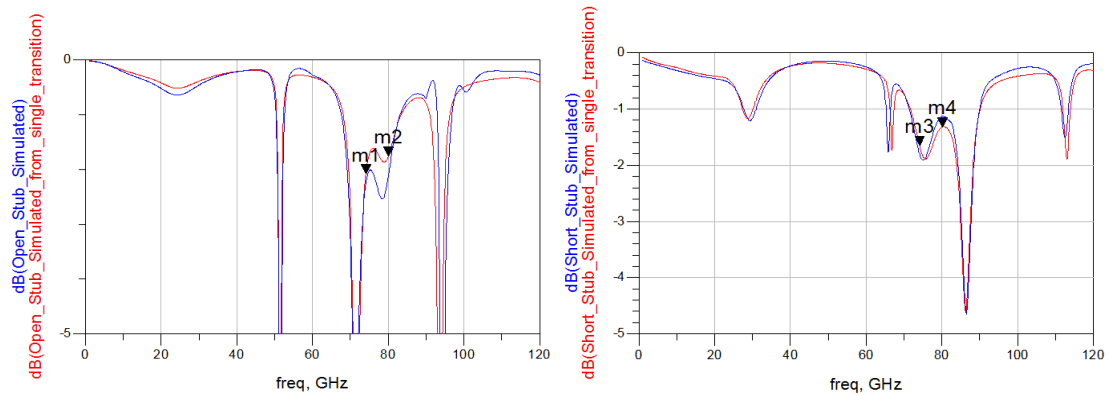


Figure 35. The schematics in ADS. Note the negative MLIN length.

Four different stub lengths were simulated to know how much the dips changes in frequency. In each simulation the stub length is lengthen by 0.05mm. The location of the stub changes significantly as seen in Figure 36.



**Figure 36.** Figure of 4 open stubs with different stub length. As seen the location of the stubs change significantly. The figure is just representative to show a major change in the location of the dips at different lengths on the stripline.

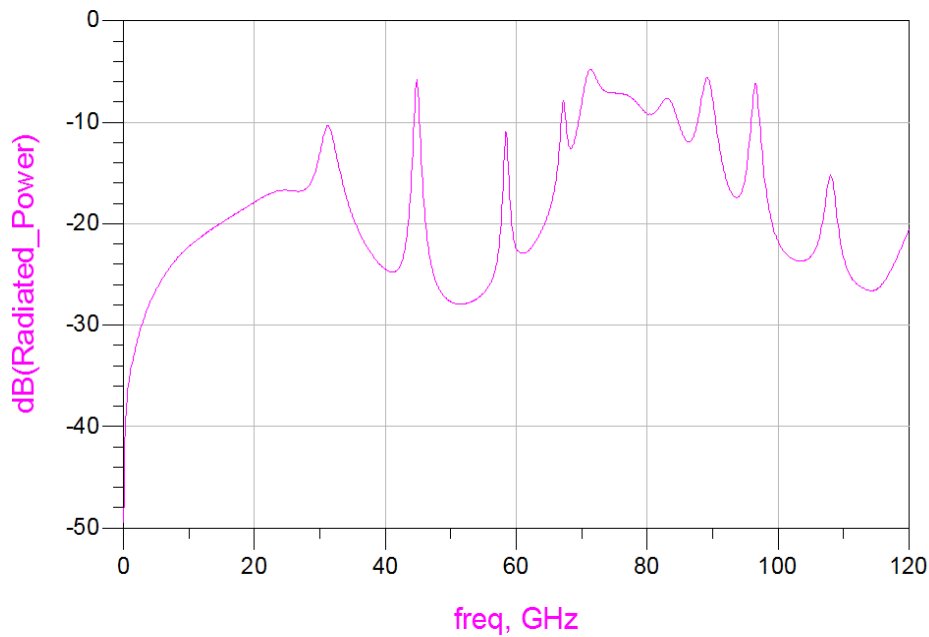


**Figure 37.** Open and shorted stub configuration in ADS, which has been tuned to match the real stubs simulations from HFSS.

m1 freq=74.10GHz dB(Open_Stub_Simulated_from_single_transition)=-1.651	m3 freq=74.10GHz dB(Short_Stub_Simulated_from_single_transition)=-2.071
m2 freq=80.10GHz dB(Open_Stub_Simulated_from_single_transition)=-1.316	m4 freq=80.10GHz dB(Short_Stub_Simulated_from_single_transition)=-1.748

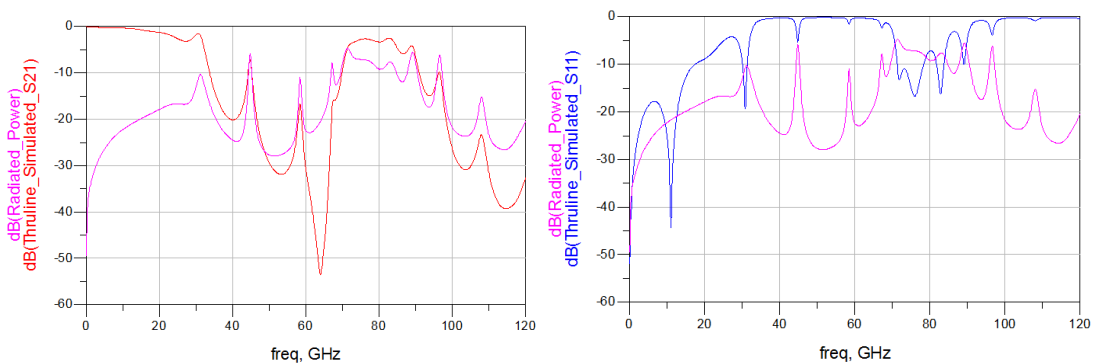
## Losses by reflections and radiated power

The assessment of the radiation losses was done in ADS. Since the reflected power and transmitted power should sum to unity, assuming lossless circuit and normalized input power, it is quite easy to calculate losses. The sum of the absolute value of S11 squared added to the absolute value of S21 squared subtracted from 1 gives the losses. Plotting this value in dB shows where the losses occur.



**Figure 38. Losses from reflection defined by  $m = 1 - |S_{11}|^2 - |S_{21}|^2$  in dB**

Since it is much more convenient to compare the radiated power with how much power that is reflected back and transmitted through the circuit the curves are plotted in the same window.

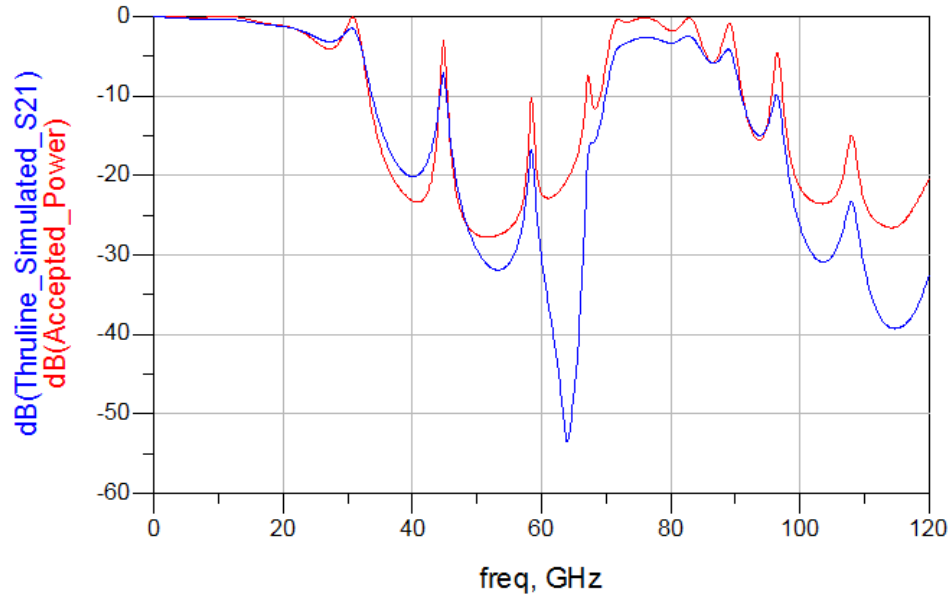


**Figure 39. The losses compared to S11 and S21 of the thru transmission. This to investigate if the losses occur at some distinct matching with the thru transmission.**

As seen in Figure 39 the losses occur mainly at frequencies corresponding to the dips in thru transmission. This is an indication that the micro strip on chip at certain frequencies radiates mainly at frequencies where the standing wave occurs.

The accepted power is the power that theoretically should enter the circuit. The accepted power is the absolute value of S11 squared subtracted from 1. In a pure lossless circuit

this should be equal to the power transferred through the circuit, the absolute value of S21 squared. The accepted power compared to S21 is a convenient tool to visualize losses. In Figure 40 it is seen that no major differences between accepted power and thru S21 is visible



**Figure 40.** The accepted power defined by  $\text{acc} = 1 - \text{abs}(S_{11})^2$  in dB.

## 71-76 GHz matching network design

The design process started with the matching network on carrier side due to two reasons. This matching network has larger dimensions and at the moment it is too angular and therefore very lossy which is a major flaw. Second the design process for this substrate is easier to implement than the one on the chip. Therefore it is better to construct a viable matching network on carrier first and then tune the overall performance of the transition with the matching network on chip. The manufacture process on chip side is a much more controlled process and is easier to tune with smaller lengths and widths, i.e. MMIC- design on chip.

The start was Silvia's design of a 71-76 GHz matching network on carrier side. Some slight modifications had to be done to adjust the matching network to the new source impedance. Since the matching network was too angular it was smoothened by attaching soft bends and using half circles to end stubs etc. Unfortunately there was no way to sweep radius of circles and curves simultaneously in ADS. Therefore the problem was solved with the trial and error method. (This problem could have been solved by programming in ADS but there was not enough time for this). Starting by smoothening the design and simulate it in ADS momentum. The result was thereafter analyzed and the design adjusted and resimulated. I.e. smaller curves, wider micro strip lines etc. This method was very time consuming since ADS momentum had to work for a long time, around 3 hours or more. Several momentum sessions later a "good enough" solution was reached. Here "good enough" implies a matching network that meets the pass band specifications but most likely is not the best theoretically achievable.

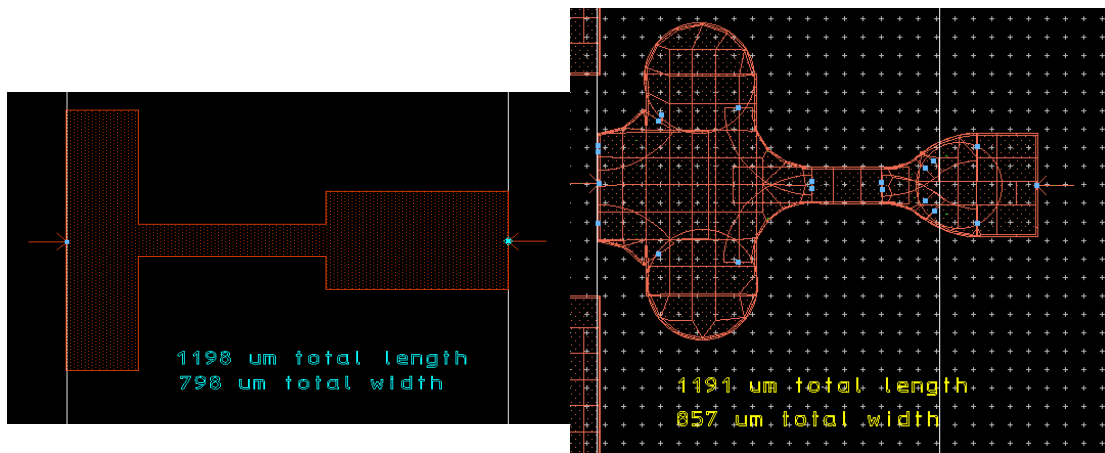
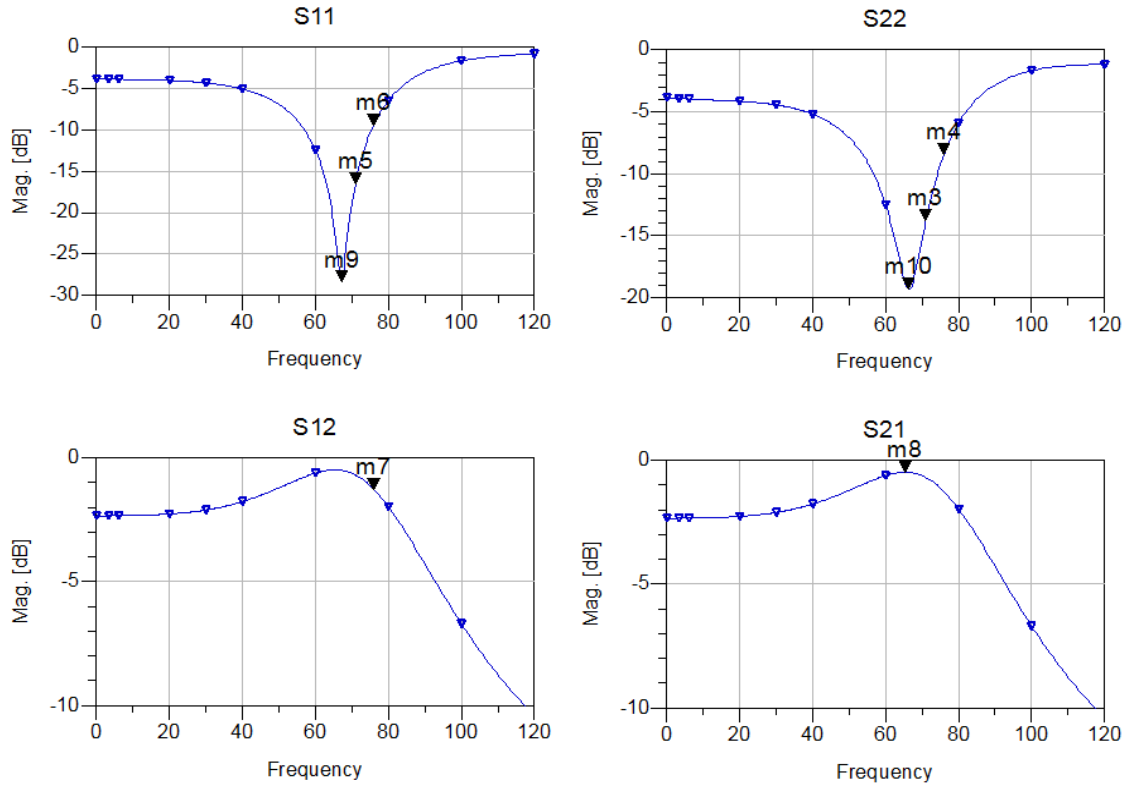


Figure 41. New matching network design on carrier side. Silvia's design to the left and the new smoothened design to the right.

One interesting thing was noted when the designed matching network was inserted into HFSS for simulation with the entire package. The return loss and insertion loss was met over the pass band of 6 GHz but the center frequency was shifted upwards 5 GHz. Thus the entire process had to be remade to achieve correct center frequency. This frequency shift has not been explained, although there are some theories discussed later on. A more direct method was applied to solve this problem. The new matching network was designed with a center frequency of around 67 GHz instead of 73 GHz. Simulations in HFSS showed this matching network met the specifications over the desired pass band of 71-76 GHz. See Figure 47. The correspondent S-parameters are presented in Figure 42.



**Figure 42. S-parameter response for matching network on carrier side.**

m5 freq=70.98GHz dB(MNcarrier_7176_GHz_a..S(1,1))=-16.463	m3 freq=70.98GHz dB(MNcarrier_7176_GHz_a..S(2,2))=-13.797
m6 freq=75.97GHz dB(MNcarrier_7176_GHz_a..S(1,1))=-9.400	m4 freq=75.97GHz dB(MNcarrier_7176_GHz_a..S(2,2))=-8.472
m9 freq=67.23GHz dB(MNcarrier_7176_GHz_a..S(1,1))=-28.407 Valley	m10 freq=66.29GHz dB(MNcarrier_7176_GHz_a..S(2,2))=-19.278 Valley
m7 freq=75.97GHz dB(MNcarrier_7176_GHz_a..S(1,2))=-1.257	m8 freq=65.36GHz dB(MNcarrier_7176_GHz_a..S(2,1))=-0.484 Peak

The design of matching network on chip side followed another approach since the main issue here was size. Starting by creating a simple transmission line model in ADS and then optimize the performance over the pass band. The starting design consisted of 4 MLIN lengths that could be tuned between 10  $\mu\text{m}$  and 50  $\mu\text{m}$  in length and between 10 and 350  $\mu\text{m}$  in width. Also an open stub that could be optimized in length and width was attached to the model. The stub was chosen to be of open type since Silvia had had problems with simulation of shorted stubs. This model was optimized with the goal to have an insertion loss less than 1 dB and a return loss less than 20 dB over the entire pass band of 71-76GHz.

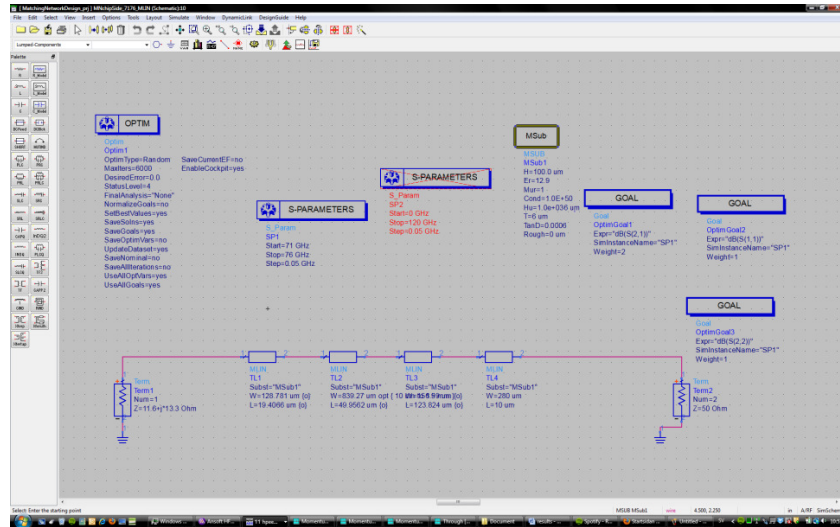


Figure 43. ADS transmission line model used to design new matching network on chip side, here shown without stub attached.

Achieving these specifications appeared not to be easy with strict limitations on size and pass band performance. Several simulations later it was realized that one or more of the constraints had to be eased since no viable pass band could be designed. Thus the length was able to be as large as 100  $\mu\text{m}$  for each transmission line. This resulted in a matching network with a very narrow and long transmission line of 9  $\mu\text{m}$  which ended with a transmission line with width of 300  $\mu\text{m}$ . This matching network was far too large to be accepted as it was. But the narrow line could easily be looped to efficiently reduce the size of the matching network. Now this matching network had to be smoothed as the one on carrier side. Realizing this smoothing matching network was very time consuming and had to be re made several times until reaching a “good enough solution”.



Figure 44. New matching network on chip side. To the left the size difference from Silvia’s design. The old matching network is in white lines.

A complete simulation with the physical transition in HFSS showed that the required specifications were not met. There was an insertion loss of 0.8-1.3 dB over the pass band and a return loss of more than 15 dB over almost the entire pass band. The pass band was however shifted upwards and was centered around 80-81 GHz instead of 73 GHz. This shift is yet unexplained but there are some theories. The transmission line model does not conform to the physical model as well as thought, thus the source and load impedances are wrong. Designing the transition model several effects have been missed or neglected. For example only the transition part is taken into account and not how the matching network itself interact with the transition or the rest of the plastic package.

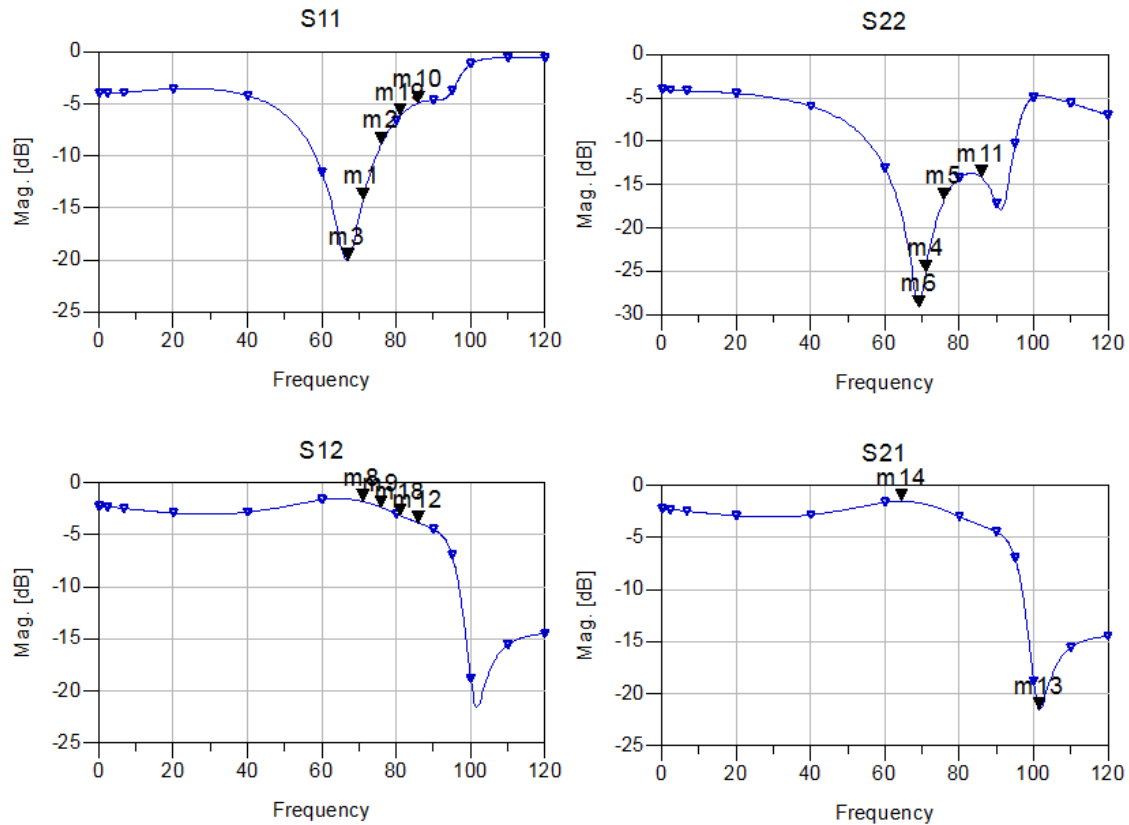


Figure 45. S-parameter response, from ADS, for the 71-76 GHz matching network chip design.

m 1 freq=70.98GHz dB(MNchip_a..S(1,1))=-14.228	m 4 freq=70.98GHz dB(MNchip_a..S(2,2))=-25.059	m 13 freq=101.6GHz dB(MNchip_a..S(2,1))=-21.538 Valley
m 2 freq=75.97GHz dB(MNchip_a..S(1,1))=-8.872	m 5 freq=75.97GHz dB(MNchip_a..S(2,2))=-16.708	m 14 freq=64.42GHz dB(MNchip_a..S(2,1))=-1.516 Peak
m 3 freq=66.92GHz dB(MNchip_a..S(1,1))=-20.002 Valley	m 6 freq=69.10GHz dB(MNchip_a..S(2,2))=-29.144 Valley	m 8 freq=70.98GHz dB(MNchip_a..S(1,2))=-1.790
m 19 freq=80.97GHz dB(MNchip_a..S(1,1))=-6.177	m 11 freq=85.97GHz dB(MNchip_a..S(2,2))=-14.150	m 9 freq=75.97GHz dB(MNchip_a..S(1,2))=-2.365
m 10 freq=85.97GHz dB(MNchip_a..S(1,1))=-4.921	m 18 freq=80.97GHz dB(MNchip_a..S(1,2))=-3.118	m 12 freq=85.97GHz dB(MNchip_a..S(1,2))=-3.860

This problem could not be handled in any standard way since the cause of the frequency shift was unknown. Therefore the new matching network was designed to have a pass band with center frequency of 67 GHz instead of 73 GHz. An identical up-shift was seen for the matching network on carrier side as well. This solution does not explain why there is a frequency shift of 6 GHz. The new designed matching network resulted in a pass band which met the specifications very well. It was centered around 73 GHz

and had a return loss less than 14 dB over pass band and an insertion loss of 0.8-1.3 dB within the pass band. See Figure 47.

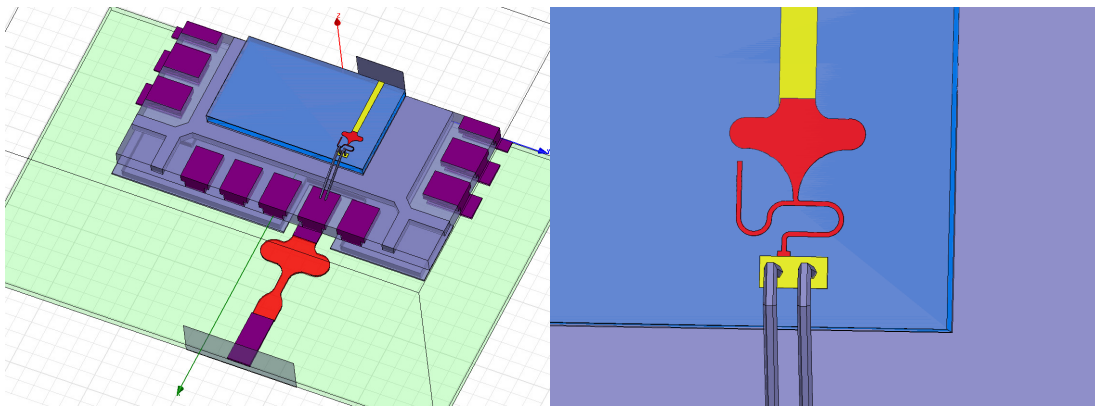


Figure 46. The resulting design in HFSS. New matching networks in red.

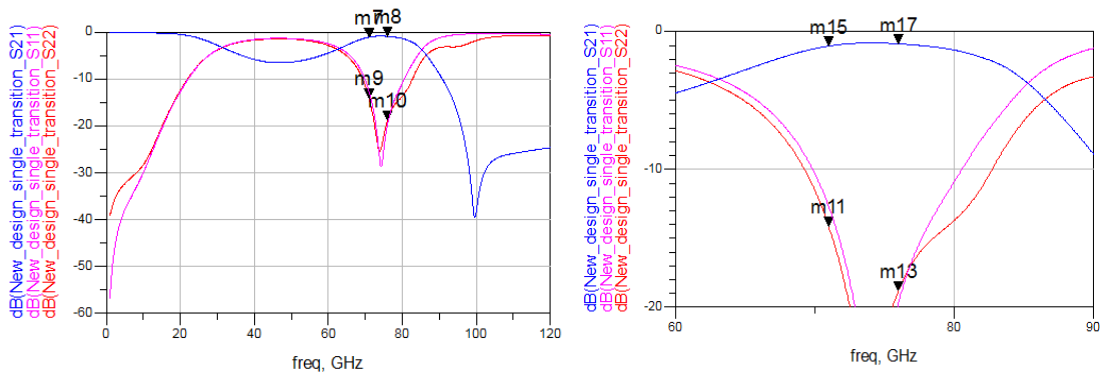
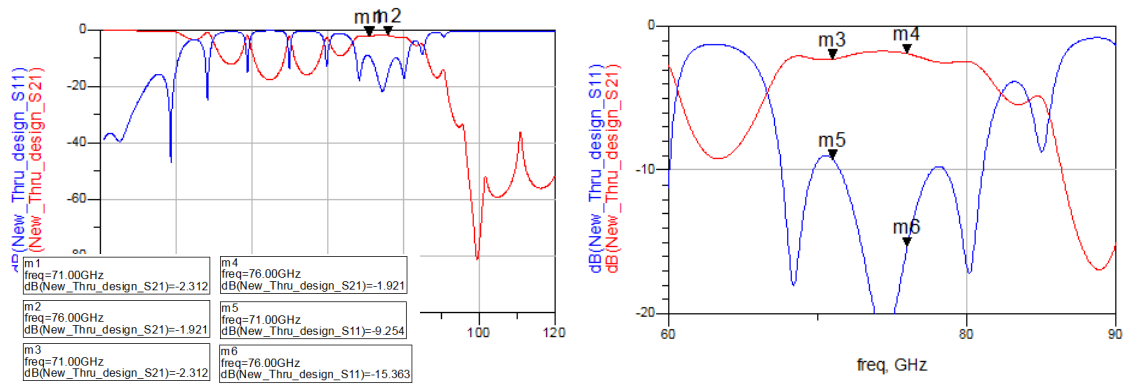


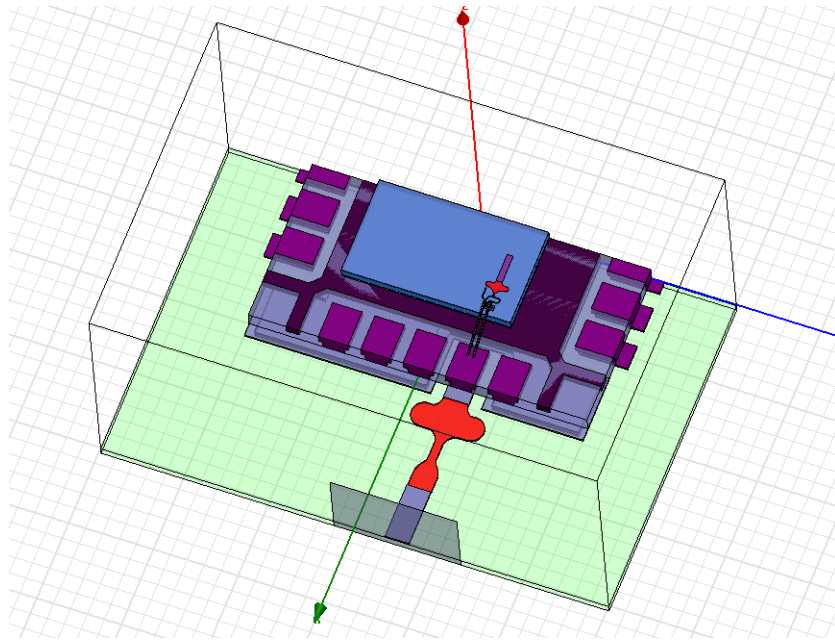
Figure 47. Single transition frequency response for 71-76 GHz design.

m7 freq=71.00GHz dB(New_design_single_transition_S21)=-1.087	m15 freq=71.00GHz dB(New_design_single_transition_S21)=-1.087
m8 freq=76.00GHz dB(New_design_single_transition_S21)=-0.927	m17 freq=76.00GHz dB(New_design_single_transition_S21)=-0.927
m9 freq=71.00GHz dB(New_design_single_transition_S22)=-14.130	m11 freq=71.00GHz dB(New_design_single_transition_S22)=-14.130
m10 freq=76.00GHz dB(New_design_single_transition_S22)=-18.805	m13 freq=76.00GHz dB(New_design_single_transition_S22)=-18.805

A thru transmission simulation in ADS indicates the insertion loss is around 3 dB and that the return loss is at its worse 9 dB, see Figure 48. This thru transmission simulation is designed solely in ADS from one single transition simulation from HFSS. The procedure is the same as describe in “matching single and thru transmission”.



**Figure 48.** Thru transition when two single transition cascaded in ADS, as described earlier.



**Figure 49.** Open and shorted stub test design.

An open and shorted stub simulation should be compared to thru transmission parameters to analyze moding problems and other unwanted frequency behavior. This is examined in the same way as described in the earliest investigations of the QFF 4x4 package. Here presenting the result of these simulations and as seen no problems arises. The difference between open, shorted stub and thru transmission are around 1 dB over the pass band. This difference can be explained by losses in the microstrip line on the chip. These simulations are also valuable for further measurements on manufactured packages. Thus also open and short stub response is presented in own window.

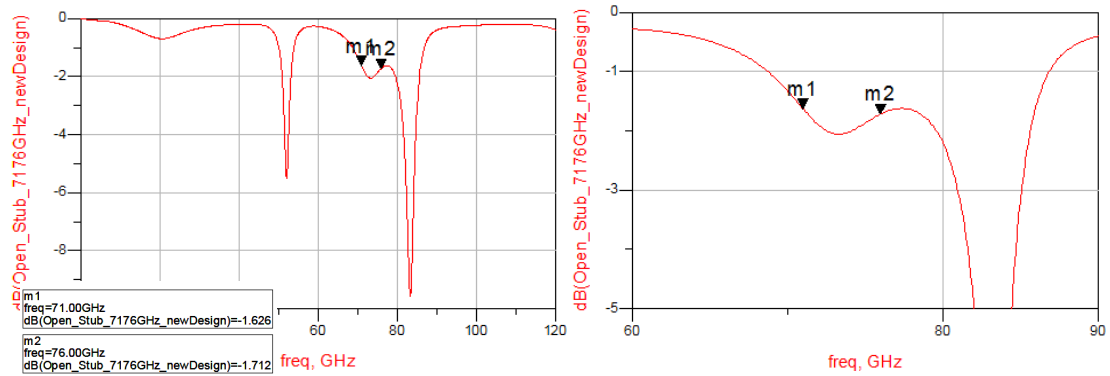


Figure 50. Open stub S-parameter for the new design.

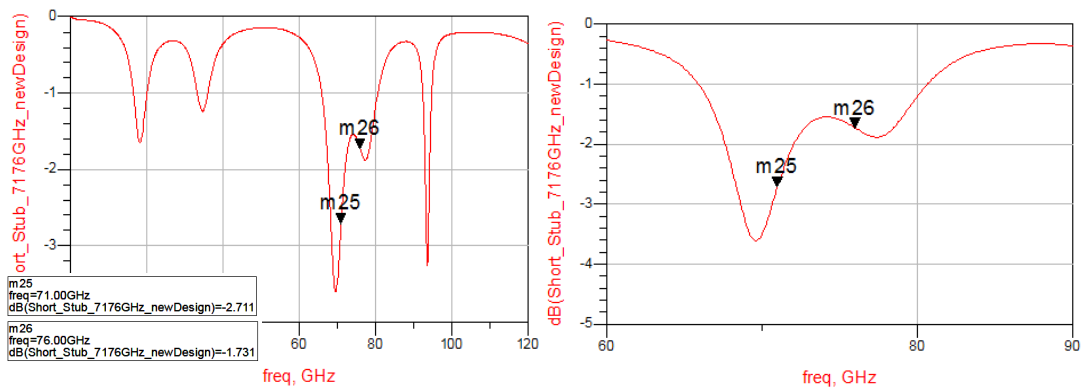


Figure 51. Short stub S-parameter for the new design.

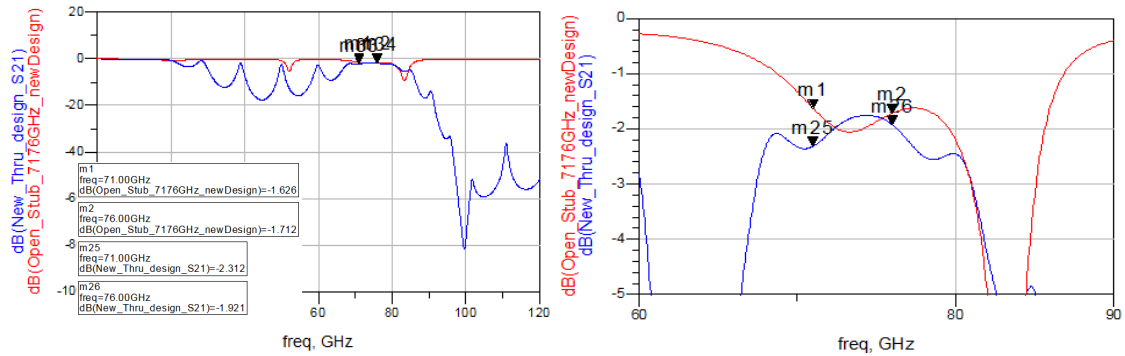


Figure 52. Comparison of open stub with thru transition, the matching over the pass band is within

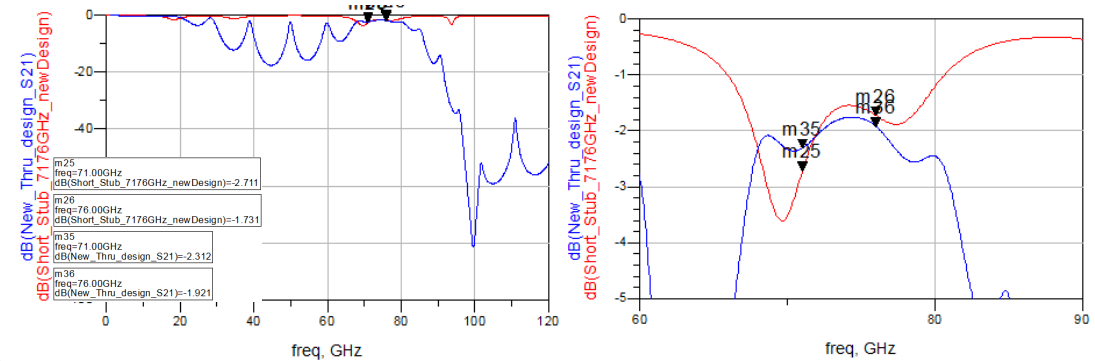
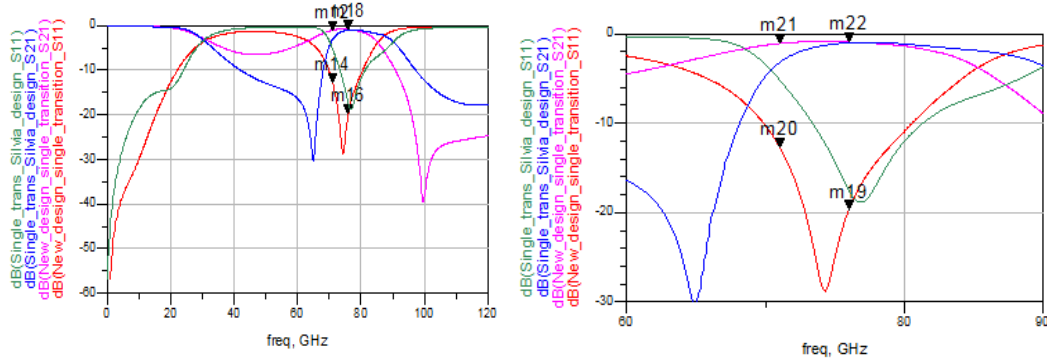


Figure 53. Comparison of shorted stub and thru transition, the matching over the pass band is within 1 dB.

It is of interest to know the differences between our design and the existing one by Silvia. To visualize the results the simulations of s parameters are plotted in the same

window. The most notable difference is the main frequency dip in S21 that is located around 100 GHz for the new design compared to 65 GHz for Silvia's design. Since there is an unexplained frequency up-shift of 2.4 GHz the dip in Silvia's design is pushed into or nearby the pass band. The new matching networks were designed to create a pass band around 73 GHz with the frequency dip pushed away from the pass band up to 100GHz to avoid this deterioration. The main differences can be seen in Figure 54. The new design has a better return loss than Silvia's design, below 14dB in 71-76 GHz. The center frequency is located at 73.5 GHz for the new design instead of 77 GHz for Silvia's design.



**Figure 54. The main frequency response differences between the new design and Silvia's design.  
The new design in pink and red, Silvia's design in blue and green.**

m12 freq=71.00GHz dB(New_design_single_transition_S21)=-1.087	m21 freq=71.00GHz dB(New_design_single_transition_S21)=-1.087
m14 freq=71.00GHz dB(New_design_single_transition_S11)=-12.695	m20 freq=71.00GHz dB(New_design_single_transition_S11)=-12.695
m16 freq=76.00GHz dB(New_design_single_transition_S11)=-19.696	m19 freq=76.00GHz dB(New_design_single_transition_S11)=-19.696
m18 freq=76.00GHz dB(New_design_single_transition_S21)=-0.927	m22 freq=76.00GHz dB(New_design_single_transition_S21)=-0.927

## 81-86 GHz matching network design

The attempts to improve the matching network for the sub band 81-86 did not show to be as easy as for the lower sub band. However some modifications were done. We start with the same specification as for 71-76 GHz band,  $RL < 15$  dB and  $IL < 1$  dB. The first design attempt was using Silvia's matching network on carrier design for 81-86GHz. This was smoothened up and attempted to get a viable pass band at these frequencies. However this method failed, no sufficient pass band could be achieved. Possibly due to the large difference of the new source impedance as compared to Silvia's source impedance.

Next step was to create a model of the matching network in ADS. This model consisted of 4 MLINs that were optimized, both width and length, to achieve a pass band at 81-86 GHz. This method is more thoroughly described for the design of matching network on chip side for 71-76 GHz. However when this model was realized in momentum the designed pass band was shifted up 30 GHz in frequency. Again, the reason for the frequency shift is unknown. No obvious error could be found with the optimization set up. Since this method worked for the lower sub band a more thorough investigation of this should be done. After some corrections and modifications a pass band for 81-86 GHz was achieved in momentum. I.e. optimizing a matching network with center frequency of 113 GHz. This model was then realized in ADS momentum and smoothened up to create a more lossless matching network. Due to time constraints this matching network has not been optimized to correct center frequency and pass band performance as thoroughly as the matching network for the lower sub band.

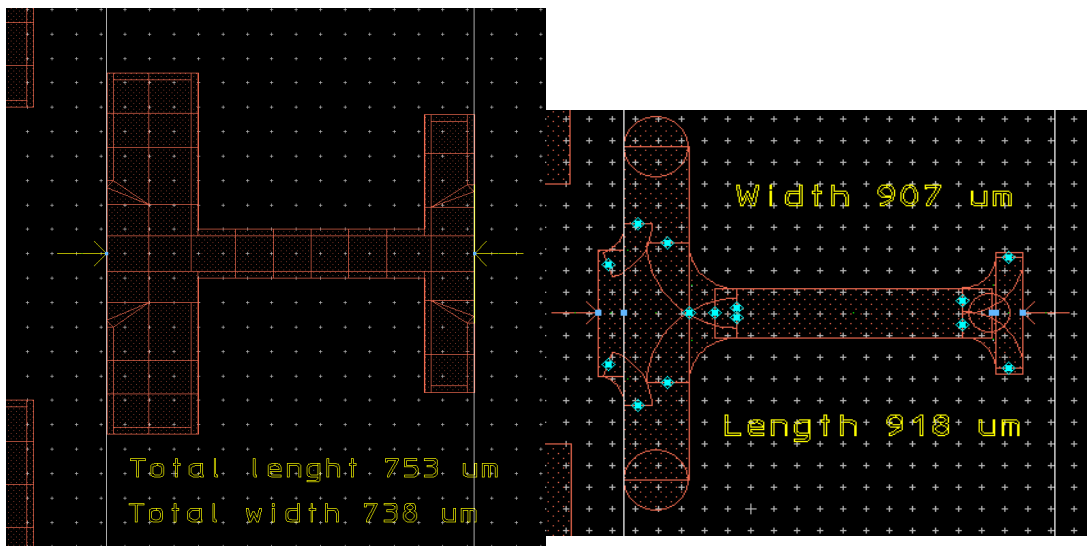
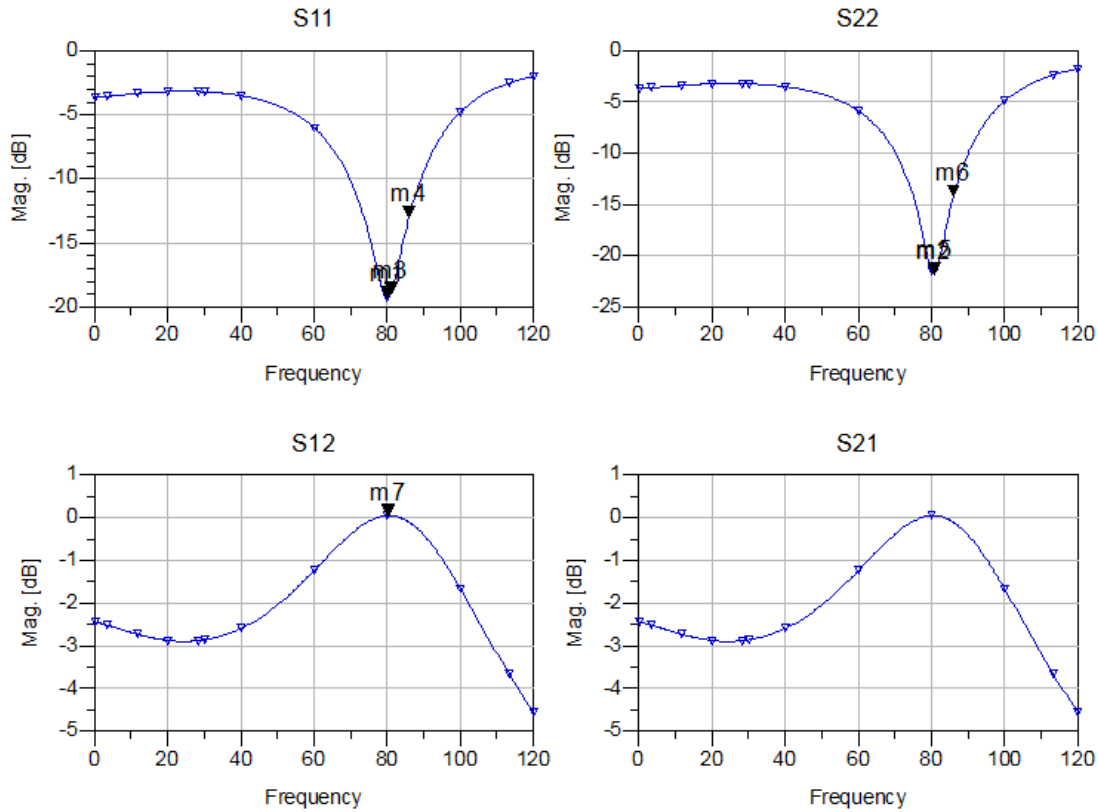


Figure 55. To the left Silvia's design and to the right the new design for carrier side 81-86 GHz.



**Figure 56. The S-parameter response for the matching network on carrier side. Simulation from ADS with source impedance.**

m1 freq=80.03GHz dB(MNcarrier8186GHz_a.S(1,1))=-19.277 Valley	m2 freq=80.66GHz dB(MNcarrier8186GHz_a.S(2,2))=-21.832 Valley	m7 freq=80.35GHz dB(MNcarrier8186GHz_a.S(1,2))=0.054 Peak
m3 freq=80.97GHz dB(MNcarrier8186GHz_a.S(1,1))=-18.939	m5 freq=80.97GHz dB(MNcarrier8186GHz_a.S(2,2))=-21.769	m4 freq=85.97GHz dB(MNcarrier8186GHz_a.S(1,1))=-13.022
	m6 freq=85.97GHz dB(MNcarrier8186GHz_a.S(2,2))=-14.254	

The design of a matching network on chip side showed to be much more difficult than expected. We use the same method as with the matching network on chip side for 71-76 GHz, a transmission line model with open stubs which could be tuned both in length in width. The model was optimized for the 81-86 GHz band. We have achieved  $RL < 20$  dB and  $IL < 1$  dB. The resulting matching network was smoothened in ADS momentum. However the designed matching network did not perform well inserted in HFSS for simulation with the plastic package. The reason for the large difference, large pass band shift and a return loss of 5 dB, is unknown. The design of this matching network had to use some more unconventional methods. This matching network was more or less designed directly in HFSS with trial and error method. Starting with the design from ADS that met the desired pass band partly open stubs were added and “tuned” to realize a pass band good enough. Partly implies here that it was investigated if this matching network could be tuned to perform well with open stubs. This resulted in a matching network which performs well in HFSS, but it is most likely not the best possible. As mentioned for the 81-86GHz matching network on carrier side time restrictions has made this matching network less thoroughly investigated and optimized compared to the one at lower frequency.

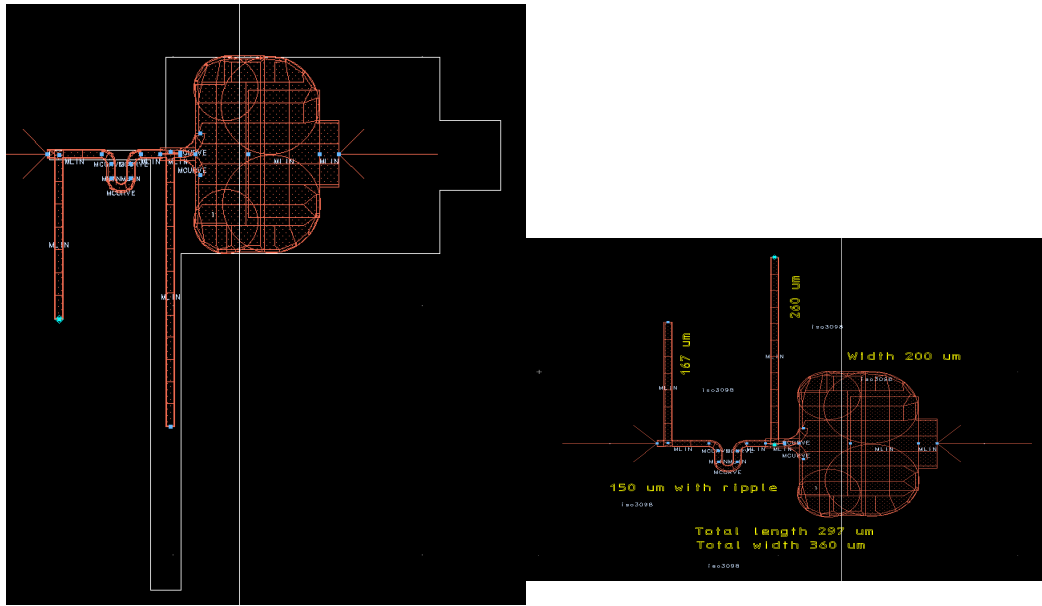


Figure 57. To the left the new design compared with Silvia's design, to the right the new design with stubs on other side to check performance.

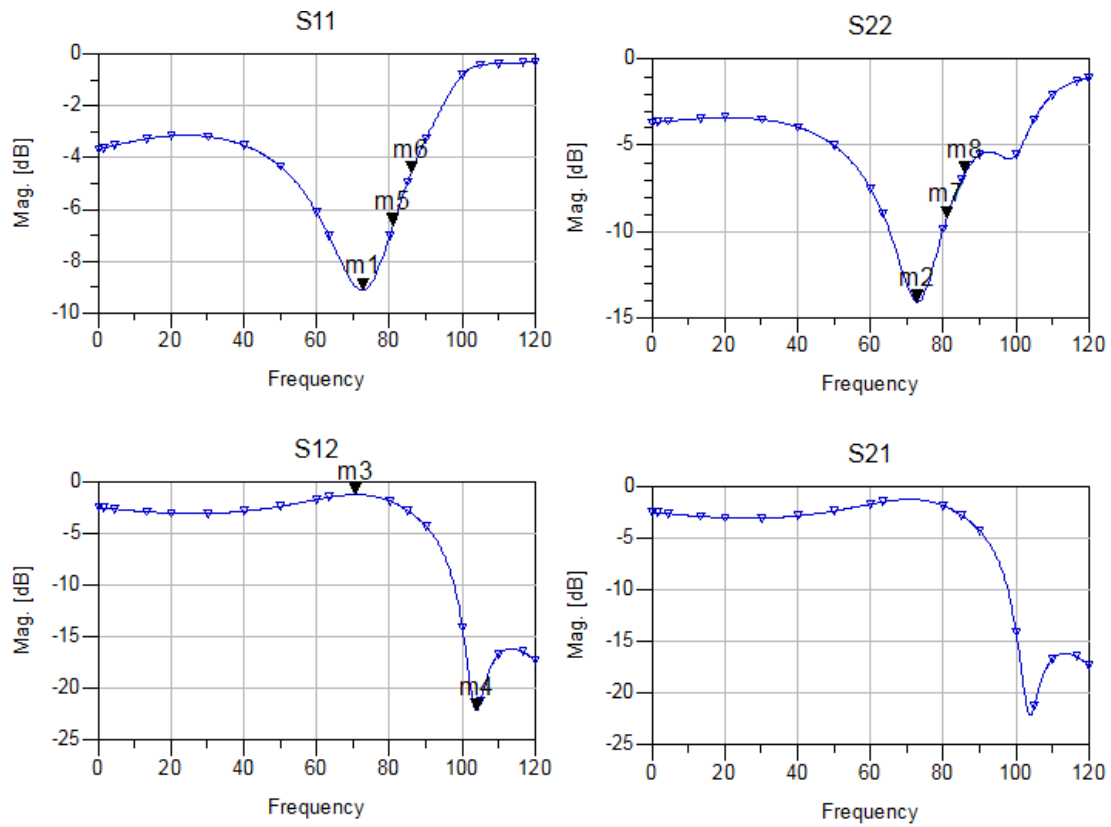


Figure 58. S-parameter response, with load impedance, for the matching network on chip side.

m1 freq=72.75GHz dB(MNchip8186_a..S(1,1))=-9.115 Valley	m2 freq=72.75GHz dB(MNchip8186_a..S(2,2))=-14.082 Valley	m3 freq=70.67GHz dB(MNchip8186_a..S(1,2))=-1.203 Max
m5 freq=80.97GHz dB(MNchip8186_a..S(1,1))=-6.622	m7 freq=80.97GHz dB(MNchip8186_a..S(2,2))=-9.216	m4 freq=103.9GHz dB(MNchip8186_a..S(1,2))=-22.121 Valley
m6 freq=85.97GHz dB(MNchip8186_a..S(1,1))=-4.587	m8 freq=85.97GHz dB(MNchip8186_a..S(2,2))=-6.619	

A design in HFSS with these matching networks is shown in Figure 59. We simulate first a single transition to assure that the pass band criteria are met. From Figure 60 it is seen that they are almost met. We have a return loss less than 15 dB and an insertion loss less than 1.2 dB over the entire band of 81-86 GHz.

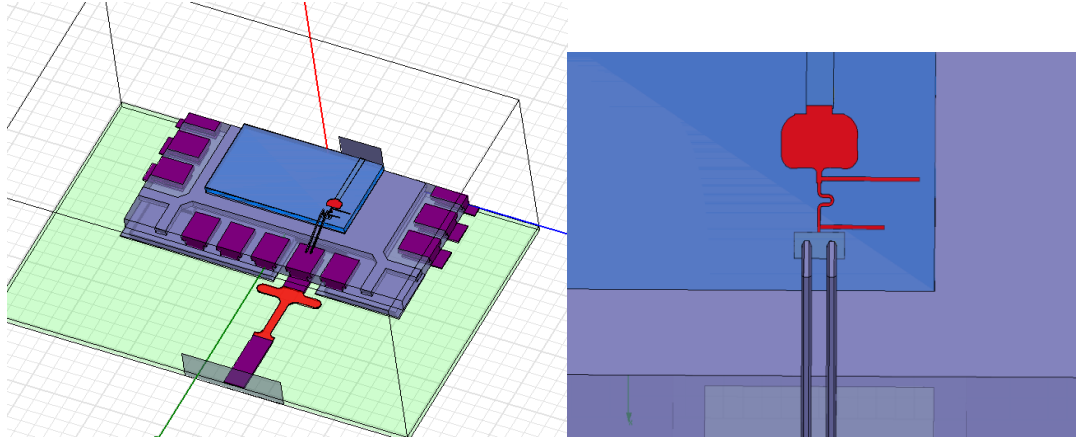


Figure 59. New matching network in red. To the left on carrier and to the right on chip.

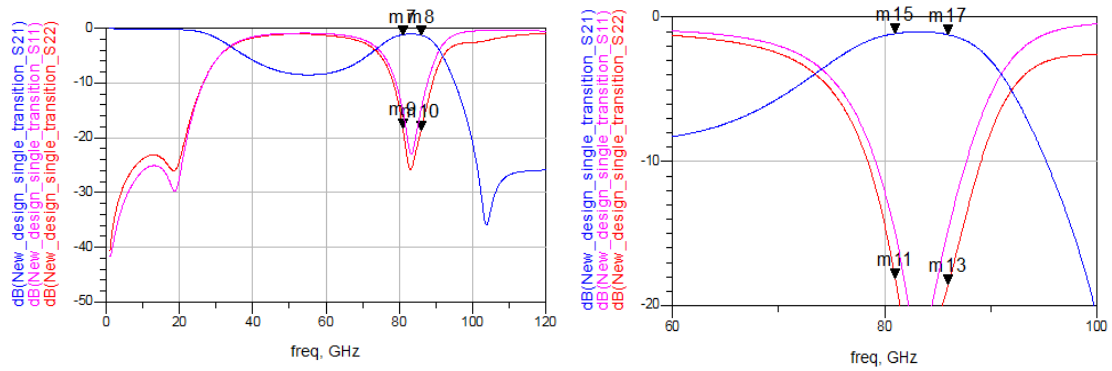
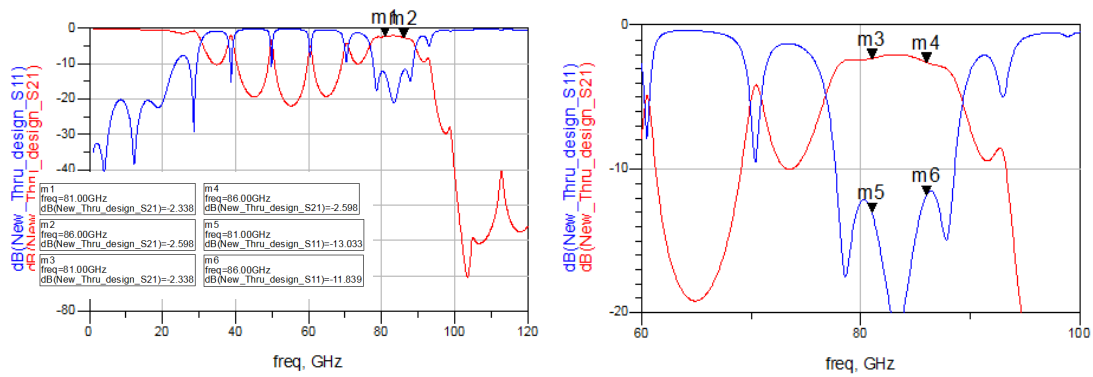


Figure 60. S-parameter response for single transition and new matching networks.

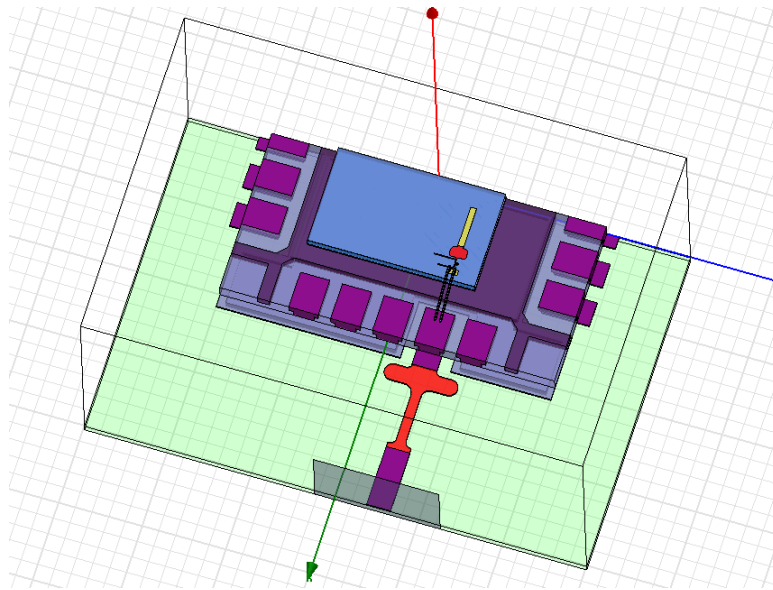
m7 freq=81.00GHz dB(New_design_single_transition_S21)=-1.159	m15 freq=81.00GHz dB(New_design_single_transition_S21)=-1.159
m8 freq=86.00GHz dB(New_design_single_transition_S21)=-1.237	m17 freq=86.00GHz dB(New_design_single_transition_S21)=-1.237
m9 freq=81.00GHz dB(New_design_single_transition_S22)=-18.121	m11 freq=81.00GHz dB(New_design_single_transition_S22)=-18.121
m10 freq=86.00GHz dB(New_design_single_transition_S22)=-18.533	m13 freq=86.00GHz dB(New_design_single_transition_S22)=-18.533

A simulation of the entire thru transmission, consisting of two cascaded single transition, in ADS, reveals that S21 has losses less than 3dB and that the S11 parameter is less than 10dB holding for any MLIN length. This simulation is testing the overall ripple effect of the thru transition and the impact on the band performance. As described earlier in this report the MLIN length is varied to examine how the ripples changes over the pass band. A worst case scenario indicates that the return loss can be as low as 8-9 dB at distinct frequencies in the pass band. This is not shown in the report. The insertion loss of 2-3dB is almost independent of the MLIN length.

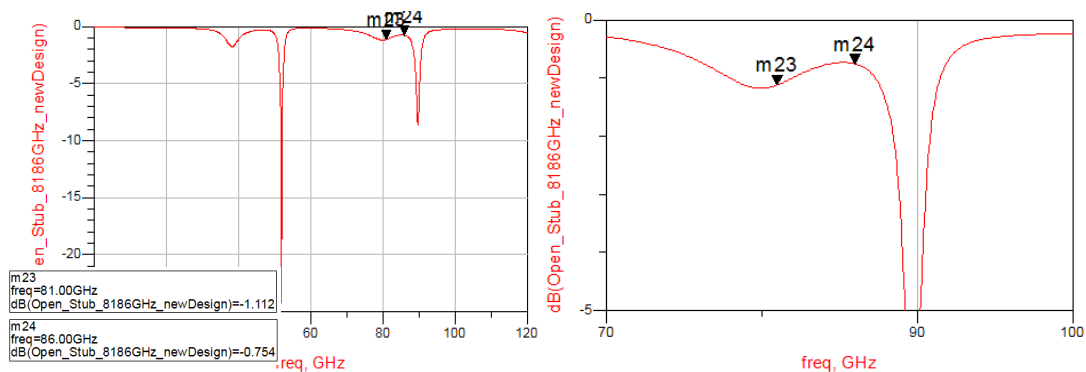


**Figure 61. Thru design simulated in HFSS from one single transition.**

An open and shorted stub simulation should be compared to thru transmission parameters to ensure that moding problems do not exist. The difference over the pass band is within 1-2dB which is acceptable and explained by losses in the microstripline on the chip, se Figure 65 and Figure 66. These simulations are very convenient to have during measurements as well. Any difference between simulations and measurements can readily be found.



**Figure 62. Open or shorted stub design for 81-86 GHz design.**



**Figure 63. Open stub S-parameter for the new design.**

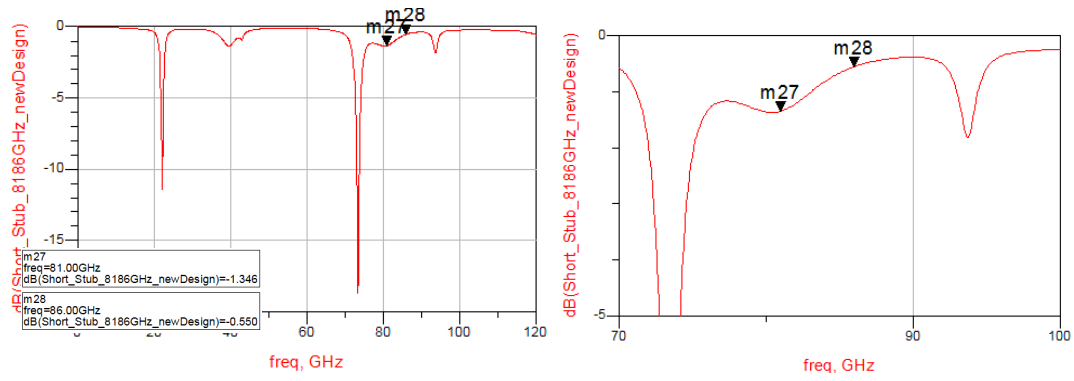


Figure 64. Short stub S-parameter for the new design.

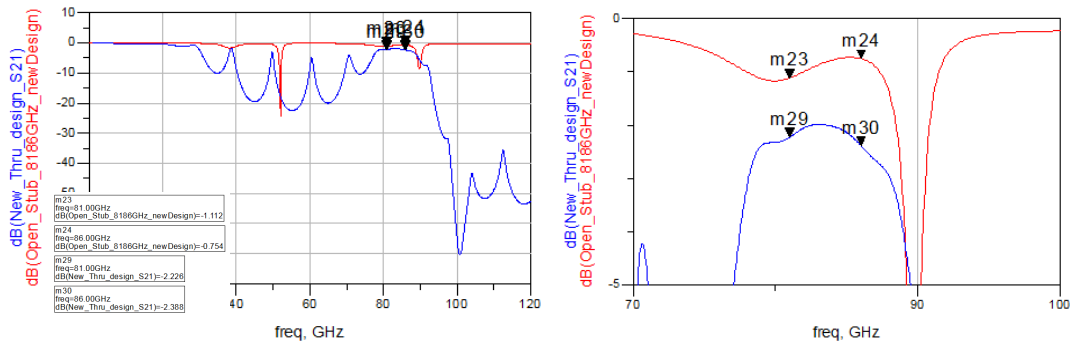


Figure 65. Open and thru design comparison. The difference over the pass band is within 2 dB.

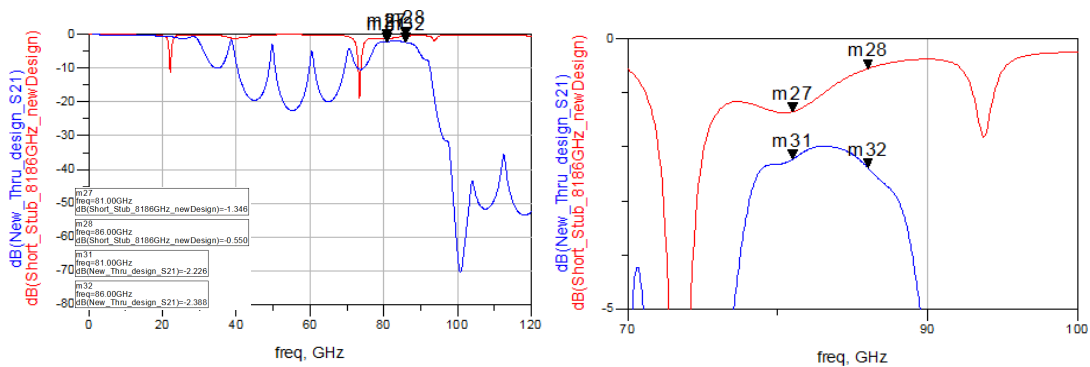


Figure 66. Short stub and thru design, the difference over the pass band is within 2 dB.

Comparing Silvia's design with the new one in Figure 67 shows that the frequency response does not diverge much between them. The new design is somewhat narrower band but it does meet the desired specifications, return loss less than 15dB and insertion loss less than 1dB. Over the pass band the difference in bandwidth is hardly noticed.

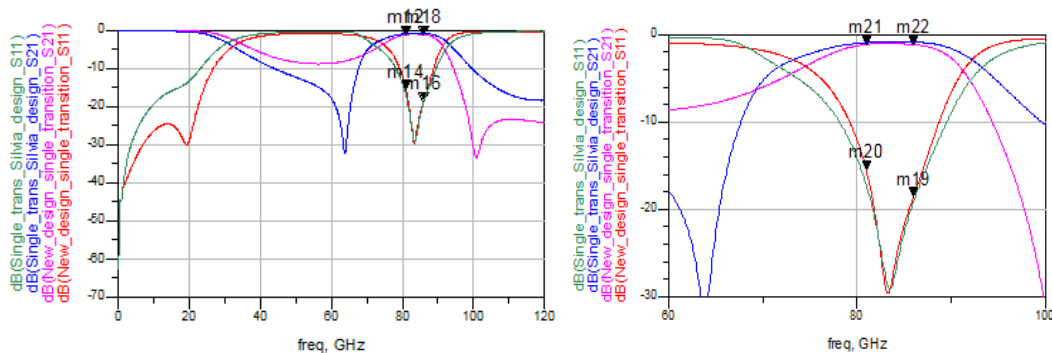


Figure 67. Comparing Silvia's design with the new design.

m12 freq=81.00GHz dB(New_design_single_transition_S21)=-1.127	m21 freq=81.00GHz dB(New_design_single_transition_S21)=-1.127
m14 freq=81.00GHz dB(New_design_single_transition_S11)=-15.509	m20 freq=81.00GHz dB(New_design_single_transition_S11)=-15.509
m16 freq=86.00GHz dB(New_design_single_transition_S11)=-18.558	m19 freq=86.00GHz dB(New_design_single_transition_S11)=-18.558
m18 freq=86.00GHz dB(New_design_single_transition_S21)=-1.160	m22 freq=86.00GHz dB(New_design_single_transition_S21)=-1.160

## Last validation test

A last test on a computer with higher capacity to assure that the earlier simulations and designed matching network are validated. The meshing in HFSS has large problems with rounded corners. It has to use large amounts of memory. There is a large difference in meshing between the normal simulations compared to the validation test, see Figure 68. This machine had enough memory to manage meshing down to a delta S of 0.02. The results are presented for single transition and thru transition 71-76 GHz and 81-86GHz.

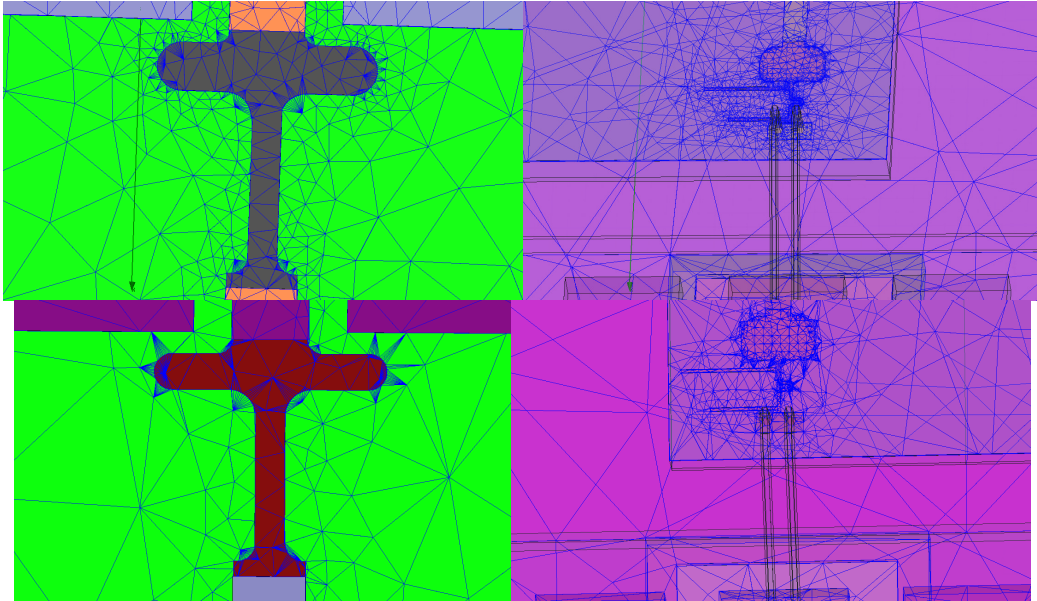


Figure 68. The difference in meshing, above with many adaptive passes and below with few.

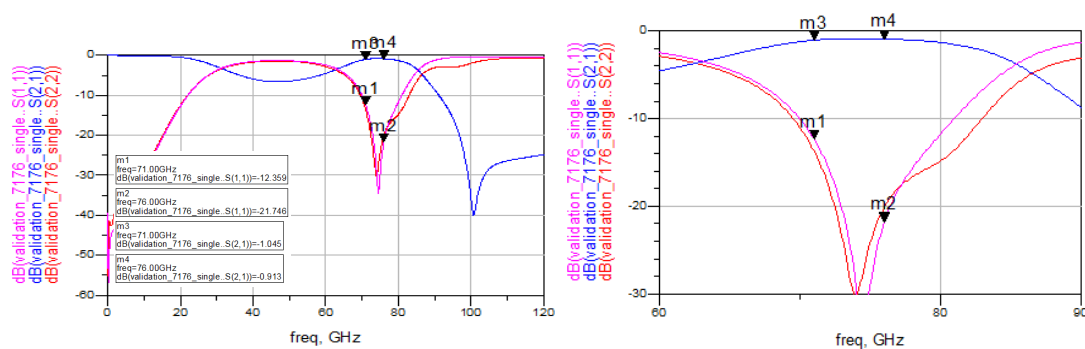


Figure 69. The new 71-76 GHz design. The validation test is very close to the other test.

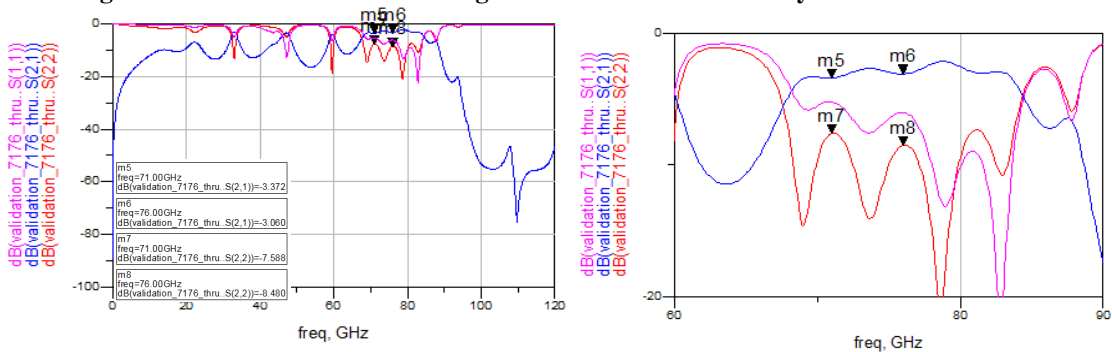
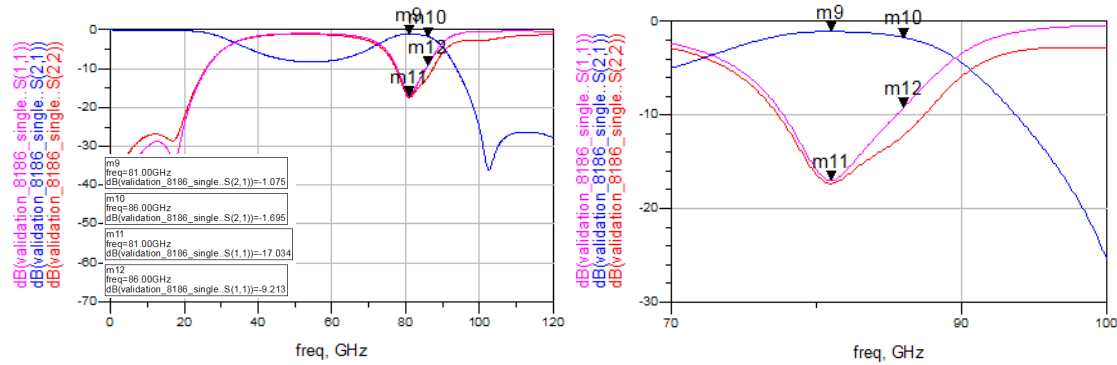
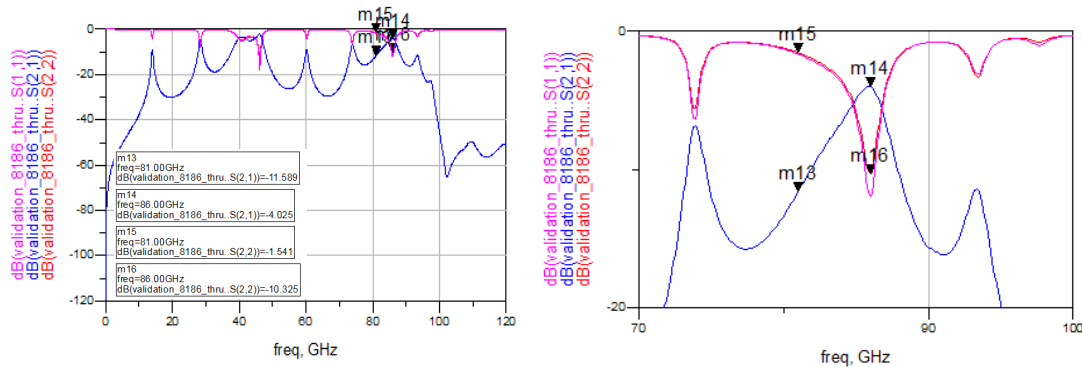


Figure 70. The validation test for the 71-76 GHz thru design differs from the designed response.

The results for the sub band 71-76 GHz are well in line with earlier simulation. Although the pass band for thru transition seem shifted upwards. The S11 and S22 parameters for thru transition seem to have worse performance than the earlier simulation from ADS. This difference could perhaps originate from the length of the micro strip line on the chip. Here it could be an unlucky length that deteriorates the performance. This must however be investigated further. Due to time restraints this cannot be done here.



**Figure 71. The 81-86 GHz single transition validation test differs very much from the designed transition.**



**Figure 72. The 81-86 GHz thru design does not have a pass band at all.**

The last test of the 81-86 GHz design is not promising. The pass band is not similar in any way to the designed one. There is a frequency shift and deteriorated return loss for single transition that cannot be explained. It should be investigated further if this difference depends on earlier lack of computer accuracy or if the matching network is erroneous in any way. The band performance for thru transmission is even worse. There is no pass band at all. This must be examined to understand where this disastrous performance originates from.



## Conclusion

The QFN4x4 package does not show any indications of having any mode problem or otherwise unwanted behavior. The early examination of the package reveals the origin of the dips in thru transmission to be a standing wave between the two ports of the package. The dips are not originating from a standing wave within the die's attach pad nor do they depend on the size of the pad. During the measurement Silvia discovered that the location of the dip were dependent on the height of the lid. This dependence is explained by the microstrip on chip acting as an antenna at certain frequencies when a standing wave occurs. Thus the distance from the microstrip to the lid determines the standing wave frequencies. Since these dips solely depend on the microstrip length on the chip nothing has to be changed with this package to achieve a viable pass band.

The cause of the frequency shift from measurement compared to simulations has not been fully investigated and understood. It is most likely caused by insufficiently accurate equations in HFSS and ADS. ADS is only valid up to 20 GHz and above these frequencies the result could diverge from reality. However the damaged probe and "poor tip calibration" from measurement set up cannot be excluded. The new matching networks were therefore designed to have pass band specifications at desired E-band frequencies. If this shift exists after new measurements it has to be investigated more thoroughly.

Matching networks on both chip and carrier side were designed for the 71-76 GHz. These matching networks meet the pass band criteria well. A return loss less than 15 dB and an insertion loss less than 1 dB over the pass band. The size of the matching network on chip has been reduced significantly. Both matching networks on carrier and chip have been smoothened to reduce losses. The last validation test confirms that these matching networks create a pass band meeting the insertion loss and return loss specifications for a single transition. However a thru transition simulated in HFSS does not perform as well as the one simulated in ADS.

The matching networks at 81-86 GHz do not have as good performance as the lower sub band. The pass band specifications are met for single transition when designing on the less powerful computer. During the last validation simulations it was shown that the pass band specifications for single transition were not met. The pass band was shifted downwards and the return loss and insertion loss were deteriorated. Thru transition did not perform well at all. These differences have not been investigated but it is clear that these matching networks do not meet specifications for 81-86 GHz.

## Future outlook

Next step should be to get a viable pass band for the 81-86 GHz region. When this is achieved these matching networks should be manufactured with the QFN 4x4 package for measurements. A set of test circuits should be built, i.e. open and shorted stub, single transition, thru transition etc. The new measurements should be performed with two probes viable up to 110 GHz. Depending on the outcome from these measurements next step should be implemented. This step will be framed from the result. If there is a frequency shift of 2.4 GHz with these designs compared to simulations this shift should be more thoroughly investigated. Otherwise other simulations and test should be examined. Oscillation problems, stability regions and noise performance should also be investigated.

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