DESIGN OF POWER AMPLIFIER USING HYBRID-90 AT 12 GHz

A graduate project submitted in partial fulfilment of the requirements
For the degree of Master of Science in
Electrical Engineering

By
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December 2016
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Acknowledgement

Before starting the dissertation, I take immense pleasure in expressing gratitude to the people who have provided all the support right from the initial stages of conception of the idea for the project.

My primary gratitude goes to Dr. Matthew Radmanesh, the chair of the committee, for providing all the facilities for me to be able to design and complete the project. He has paid careful attention and semi-infinite number of helpful suggestions.

I am also happy to acknowledge the assistance of Professor Benjamin Mallard and Dr. Xioujun Geng, for the advice and assistance they extended to this project.

Above all, this achievement would not have been possible, if not for the encouragement and solace provided by my parents and lord almighty. Last but not the least, many thanks to the esteemed staff of our university I owe more than what I can mention.
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ABSTRACT

DESIGN OF POWER AMPLIFIER USING HYBRID-90° AT 12 GHz

By

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Master of Science in Electrical Engineering

The aim of the project is to design an RF-power amplifier using hybrid-90° couplers. The main objective of this project is designing the layout and simulation using advanced design system and RF/MW E-book software at an operating frequency of 12 GHz. A low noise transistor MESFET MGF1941AL was chosen to attain the power gain greater than or equal 10 dB and a 28 dBm of output power. To design the input and output matching networks lossless lumped elements were used since these components exhibit lower sensitivity as frequency changes. In addition, two hybrid-90° couplers were used to divide and combine the applied power. Hybrid-90° couplers were designed using microstrip line method on a RO4003C substrate (from Rogers Corporation). Every single design step was verified by the Advanced Design System (ADS) and RF/MW E-book.
Chapter 1. INTRODUCTION

An electronic tuned amplifier is a device, which amplifies high frequency signals used in Rf communications, is known as a Radio Frequency amplifier, or RF amplifier. This amplifier can be tuned over several input frequencies. Due to the rapid growth of wireless communication, we have several types of RF amplifiers in the modern market based on user requirements such as frequency range, supply current, gain, and output power etc., using RF power amplifier there are multiple applications including TV transmissions, Wireless Communication, Radar, and RF heating. Output power of a RF power amplifier varies from a few mW to MW, depending on application. Most crucial parameters that characterize an RF power amplifier are:

1. Output power
2. Linearity
3. Gain
4. DC supply voltage
5. Stability

The performance of a power amplifier can be determined by selecting an appropriate bias point. When a power amplifier is able to produce a signal that is delivered to the load of a system, whose power is converted from the DC power of the input supply voltage, this ability of the power amplifier is known as Efficiency.

Efficiency of a power amplifier as follows:

\[ \text{Efficiency (}\eta\text{)} = \frac{\text{AC output power}}{\text{DC input power}} \times 100\% \]  

(1.1)

The Design of an amplifier involves several important steps. Primarily selecting a proper device plays a crucial role. Before selecting appropriate transistor device, characteristics should be performed using s-parameters from manufacturer’s datasheet. In my project, I selected MGF 1941 AL Power GaAs FET as my transistor, and device characteristics are checked using s-parameters from the datasheet. The Second step involves DC-bias design; this biasing of device can be done using the parameters such as threshold voltage, drain to source voltage and drain current. Once we complete the device biasing, we can further move to the stability check of the device. The Stability check of transistor can have calculated from K-\(\Delta\) test. Since we are performing power
amplifier design, our transistor should be in unconditionally stable. After completion of the stability check device, input and output-matching networks should be drawn using smith charts thereby we can design an amplifier circuit. In this project to achieve the required output power, we need to design a two-stage power amplifier. Instead of cascading two amplifiers, I am using the hybrid-90 degree couplers in this project.

Figure 1.1-amplifier block diagram

1.1 Goal:

The objective of this project is to design a power amplifier using hybrid-90 degree couplers to achieve the following specifications.

<p>| | |</p>
<table>
<thead>
<tr>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Frequency</strong></td>
<td>12 GHz</td>
</tr>
<tr>
<td><strong>Output power</strong></td>
<td>28 dBm</td>
</tr>
<tr>
<td><strong>Gain</strong></td>
<td>≥10 dB</td>
</tr>
</tbody>
</table>

Table 1 SPECIFICATIONS
CHAPTER 2. DESIGN THEORY OF POWER AMPLIFIER

In this chapter, a power amplifier using 90-degree hybrid coupler will be discussed. A power amplifier is an electronic device in which power of the output signal that is sent to the load is greater than the power of the input signal that is consumed by the driver. A power amplifier is also known as a device, which is used to convert energy as a DC to DC converter or a radio frequency (RF) oscillator. This oscillator converts power which is DC in nature, into power that is of consistent wave nature. An ideal power amplifier transforms the power that is in DC form, to power of the signal which is used as the output which is controlled continuously by the input which is a RF signal. In wireless communication, power amplifier can basically be described as a DC to modulated RF converter.

\[ P_{RF-IN} + P_{DC} = P_{LOSS} + P_{RF-OUT} \]  \hspace{1cm} (2.1)

Figure 2.1 Power flow diagram in a power amplifier.
There are a few important design considerations for the design of a power amplifier such as stability of the transistor, gain, etc. Overview of these parameters can be done using the user manual provided by transistor manufacturer. In this project, we are using a 90-degree hybrid coupler, so the operation and working of coupler explained later and study of microstrip line, which will be explained clearly at the end.

![Block diagram of power amplifier](image)

**Figure 2.2 Block diagram of power amplifier.**

### 2.1 TYPES OF POWER AMPLIFIER

Depending on the mode of operation, power amplifiers are divided into five classes shown in Figure 2.3. When the transistor operates in a Trans conductance region, the power amplifier converts the input radio frequency signal into current. This circuit is known as a linear power amplifier. When the transistor acts like a switch, the power amplifier circuit is called a switching PA.

![Power amplifier family tree](image)

**Figure 2.3 Power amplifier family tree.**
2.2 STABILITY OF POWER AMPLIFIER

In the process of designing a power amplifier, the first essential step is checking the stability condition of the given transistor. Therefore, the designing of a power amplifier requires “unconditional stable” condition, which implies entire Smith Chart is in stable to perform the input matching and output matching network. The condition “not unconditionally stable”, tells us there are some regions on the Smith Chart where the given transistor may tend to oscillate. This is not a required condition to design an amplifier. If we come across “conditionally stable” condition, in this case stability circles should be drawn. There are certain tests that determine the stability condition of a given transistor. Among these methods in this project, K-Δ test is used where K-is known as the stability factor and Δ-is defined by the determinant of the S-matrix. Considering the required operating frequency and the S-parameters of MGF1941AL, we get:

\[
\Delta = S_{11}S_{22} - S_{12}S_{21} \quad (2.2)
\]

\[
K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} \quad (2.3)
\]

Unconditionally stable condition: i) \( K > 1 \) and \( S_{11} < 1 \) and \( S_{22} < 1 \).

2.3 GAIN OF POWER AMPLIFIER

The gain is second parameter, which plays a vital role in amplifier design process. Calculating gain has two main assumptions, one being unilateral design and other is bilateral design.

- When \( S_{12} = 0 \), then we should use unilateral design formulas.
- When \( S_{12} \neq 0 \), calculate the U (unilateral figure of merit) and the error range. If the error range is very small, then uni-lateral condition is assumed. Or else, a bilateral design assumption should be taken.

Then the gain of the transducer which is maximum \( (G_{tu, \text{ max}}) \) is given by:

\[
G_{tu, \text{ max}} = [ G_{\text{smax}} \cdot G_0 \cdot G_{\text{L, max}} ]
\]

\[
G_S = \frac{1 - |\Gamma_{s}|^2}{|1 - \Gamma n \cdot \Gamma_{s}|^2} \quad (2.4)
\]


\[ G_0 = |s_{21}|^2 \]  

\[ G_L = \frac{1 - |\Gamma_L|^2}{|1 - S_{22}\Gamma_L|^2} \]  

### 2.4 ERROR RANGE

Error range plays an important role in determining the design of an amplifier whether it is a unilateral amplifier or bilateral amplifier. In our amplifier design, we found that we are going to build a uni-lateral power amplifier using hand calculations. Here we can define the uni-lateral figure of merit,

\[ U = \frac{|S_{12}|\|S_{21}\|S_{22}|S_{11}|}{(1-S_{11}^2)(1-S_{22}^2)} \]  

\[ \frac{1}{(1+U)^2} < \frac{G_t}{G_{T_{umax}}} < \frac{1}{(1-U)^2} \]  

If the transistor is unilateral:

\[ \Gamma_{IN} = S_{11} \]  

\[ \Gamma_{OUT} = S_{22} \]  

### 2.5 DC-BIASING

The Designing process of a power amplifier effectively relies on biasing of the transistor. So, the transistor has to be biased at its Q-point. Based on the biased region, we come to know that the transistor operates in a linear region or in an active region, this depends on multiple properties.

![MOS FET characteristics curves](image)

Figure 2.4 MOS FET characteristics curves.
To avoid the leakage DC-biasing circuit has to be separated there by microwave signals travels.

1. Radio frequency-choke (inductor) has to be placed between the MW circuit and the DC source.
2. To create high impedance path to a radio frequency signal, a $\lambda/4$ transformer should be placed between radio frequency circuit and DC-source.
3. To create an open circuit a high impedance capacitor should be placed as a load to the transformer.

![Figure 2.5 Circuit diagram of DC-biasing.](image)

This DC-biasing is done to control the circuit dc voltage is deliberately applied between two points. In our power amplifier the transistor is MOS FET so, we applied the dc-voltage between the gate and source.
2.6 Designing steps

1. Device selection:

Transistor has to be selected such that \( S_{21}/S_{12} \) is greater than the desired gain.

2. Biasing:

DC- biasing has to be done to meet the required parameters of MOSFET.

3. Characterisation:

Device characterisation has to be done using the s- parameters of transistor from the transistor data sheet.

4. Stability:

With the given operating frequency stability test has to be done to check whether the transistor is stable or not at that frequency of operation. Here we are going to calculate \( K \) and \( \Delta \) values; based on those values we can determine the stability condition.

5. Matching circuit design:

Using the smith chart, we are going to build input matching network and output matching networks of amplifier. In this step we are also going to determine the amplifier is whether unilateral or bilateral using the conditions. By performing matching, we are able to figure out the gain of amplifier.

![Figure 2.6 I/P and O/P matching networks.](image)

6. Designing amplifier circuit:

Based on the above steps finally we build amplifier circuit as per the design rules. In designing the amplifier circuit, every parameter should be keenly observed to avoid the unwanted noises.
2.7 Flow chart of design steps

Figure 2.7 Flow chart for amplifier design.
2.8 Hybrid 90-degrees coupler

The Hybrid 90-degree couplers are four-port devices in which incident power is split into 2 output ports. At the output ports, the signals are attenuated by 3dB and they are going to have a phase difference of 90 degrees with each other. 3dB attenuation tells that half of the input power is dissipated. Moreover reflections, which were created to mismatches are propagated to the isolation port to prevent the power from reflecting back to the input port. The hybrid 90-degrees coupler can also be used to combine power signals with an isolation between the ports.

![Branch line hybrid coupler](image)

Figure 2.8 Branch line hybrid coupler.

3 dB indicates half power, therefore 3 decibel coupler splits the power equally (within a certain tolerance) among the coupled output ports and output. The 90-degree phase shift among the outputs makes hybrid couplers useful in the design of modulators, electronically variable attenuators, microwave mixers, and multiple microwave components.
Chapter 3. Hybrid-90° coupler design steps

In this project, the hybrid-90° coupler plays a vital role. Where I am using this coupler instead of cascading two single stage amplifiers here, I am attaching one amplifier to the port-2 of divider to port-1 of combiner and from the port-3 of divider to port-4 of combiner another single stage amplifier. Therefore, the schematic looks as,

![Figure 3.1 block diagram of 2-stage power amplifier using hybrid-90° couplers.](image)

3.1 Designing steps

Step 1.

As microstrip line method is one of the effective way to integrate the circuit with both passive and active microwave devices, we are considering the value of di-electric constant (Ɛ r) as 3.55; height of substrate h is 0.508 mm by referring the Roger’s corporation datasheet.

Step 2.

To design the microstrip layout of hybrid-90° coupler I am referring the required formulas from professor. Matthew Radmanesh textbook, they are as follows,
\[ \varepsilon_r = \varepsilon_s \varepsilon_o, \]
\[ \varepsilon_o = 8.854 \times 10^{-12} \text{ F/m} \]

- Effective dielectric constant is

For \( W/h \leq 1 \):
\[ \varepsilon_{\text{ff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[ \left( 1 + 12 \frac{h}{W} \right)^{-1/2} + 0.04 \left( 1 - \frac{W}{h} \right)^2 \right], \]

For \( W/h \geq 1 \):
\[ \varepsilon_{\text{ff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left( 1 + 12 \frac{h}{W} \right)^{-1/2} \]

\[(3.1)\]

- Characteristic impedance is

For \( W/h \leq 1 \):
\[ Z_0 = \frac{60}{\sqrt{\varepsilon_{\text{ff}}}} \ln \left( \frac{8h}{W} + \frac{W}{4h} \right) \]

For \( W/h \geq 1 \):
\[ Z_0 = \frac{120\pi}{\sqrt{\varepsilon_{\text{ff}}}} \left[ \frac{W}{h} + 1.393 + 0.667 \ln(W/h + 1.444) \right] \]

\[(3.2)\]

- Ratio between width and heights are

For \( W/h \leq 2 \):
\[ \frac{W}{h} = \frac{8e^A}{e^{2A} - 2} \]

For \( W/h \geq 2 \):
\[
\frac{W}{h} = \frac{2}{\pi} \left[ B - 1 - \ln(2B - 1) + \frac{\varepsilon_r - 1}{2\varepsilon_r} \left\{ \ln(B - 1) + 0.39 - \frac{0.61}{\varepsilon_r} \right\} \right]
\]

(3.3)

Where

\[
A = \frac{Z_0}{60} \sqrt{\frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{\varepsilon_r + 1} \left( 0.23 + \frac{0.11}{\varepsilon_r} \right)}
\]

and

\[
B = \frac{377\pi}{2Z_0 \sqrt{\varepsilon_r}}
\]

(3.4)

- Wave length ($\lambda$) is given as

For $W/h < 0.6$:

\[
\lambda = \frac{\lambda_0}{\sqrt{\varepsilon_r}} \left[ \frac{\varepsilon_r}{1 + 0.6(\varepsilon_r - 1)(W/h)^{0.0297}} \right]^{1/2}
\]

For $W/h \geq 0.6$:

\[
\lambda = \frac{\lambda_0}{\sqrt{\varepsilon_r}} \left[ \frac{\varepsilon_r}{1 + 0.63(\varepsilon_r - 1)(W/h)^{0.1255}} \right]^{1/2}
\]

(3.5)

Step 3.

By using above formulae, calculated the values of microstrip line with

$Z_0 = 50$ Ω are as follows,

$W/h = 2.235$ mm,

$W = 1.135$ mm.
For $Z_0 = 35.4 \, \Omega$

$W/h = 3.73 \, \text{mm},$

$W = 1.89 \, \text{mm}.$

Length of transmission line ($l$) = 3.63x$10^{-3}$.

### 3.2 Variation of $Z_0, W/h$ from the graph

![Graph showing plots of $Z_0$ vs $W/h$](image)

Figure 3.2 Plots of $Z_0$ vs $W/h$ with ($\varepsilon_r$).
3.3 Design of hybrid-90° coupler using Advanced Design System

As we discussed earlier in this project we are using hybrid-90° combiner and divider, these two couplers are designed using microstripline method.

3.3.1 Hybrid-90° combiner:

Figure 3.3 Hybrid-90° combiner.
3.3.2 Hybrid-90° divider:

Figure 3.4 Hybrid-90° divider.
3.4 Layout diagram of Hybrid-90° coupler

Figure 3.5 Layout of Hybrid-90° coupler.
Chapter 4. RF-AMPLIFIER DESIGN STEPS

The Design of a radio frequency power amplifier is done in two stages where these two stages are coupled through a 90-degree hybrid coupler.

By referring the transistor MGF1941AL data sheet at the frequency of 12 GHz following S-parameters are taken into consideration,

\[
S_{11} = 0.548 \angle 106.4^0
\]
\[
S_{12} = 0.107 \angle -3.6^0
\]
\[
S_{21} = 3.16 \angle -25.2^0
\]
\[
S_{22} = 0.177 \angle -162.1^0
\]

Table 2 MGF1941AL S-parameters.
4.1 Calculation of new $S_{21}$

To design a class A power amplifier, we need to calculate the new $s_{21}$ parameter. For the large signal analysis, all the $s$-parameters are remains same where only $s_{21}$ is going to change.

The value of new $S_{21}$ is going to decreased as the power and frequency increases, this is due to power saturation at output.

$$10 \log |S_{21}|^2 = 9 \text{ dBm}$$

$$|S_{21}| = 2.81.$$

4.2 Stability test

Here we are doing stability test for MGF1941AL transistor at a frequency of 12 GHz. We have two methods for checking stability of a transistor 1. $K$-$\Delta$ test, 2. Using smith chart stability circles.

However, in this project I am using $K$-$\Delta$ test as follows,

a) $$\Delta = S_{11}S_{22} - S_{12}S_{21}$$

$$= (0.548 \angle 106.4^0) x (0.177 \angle -162.1^0) - (0.107 \angle -3.6^0) x (2.81 \angle -25.2^0)$$

$$|\Delta| = 0.166$$

Therefore $|\Delta| < 1$.

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}|S_{21}}$$
Therefor \( K > 1 \).

Since \(|\Delta| < 1\), \( K > 1 \) given transistor is unconditionally stable.

### b) Checking for unilateral assumption

\[
\frac{1}{(1 + U)^2} < \frac{G_T}{G_{TU_{max}}} < \frac{1}{(1 - U)^2}
\]

Where,

\[
U = \frac{|S_{12}| |S_{21}| |S_{11}| |S_{22}|}{(1 - S_{11}^2)(1 - S_{22}^2)} = \frac{|0.107||2.81||0.548||0.177|}{(1-(0.548)^2)(1-(0.177)^2)} = 0.04
\]

Therefor \(-0.34\) dB < \( \frac{G_T}{G_{TU_{max}}} \) < 0.35.

The maximum error tolerance is between + or - 0.5 dB. Therefore we can assume unilateral amplifier.

### c) Calculating \( G_{TU_{max}} \)

\[
G_{TU_{max}} = (S_{21}/S_{12}) \cdot (K-(K^2-1)^{0.5})
\]

\[
= 11.49
\]

\( G_{TU_{max}} = 10.6 \) dB

\( G_{1 \ dB} = 9.6 \) dB
4.3 Matching networks

\[ S_{\text{in}} = 0.548 \angle 106.4^\circ \]
\[ S_{\text{in}}^* = 0.548 \angle -106.4^\circ \]

Figure 4.1 Smith chart design for matching network of the input
Figure 4.2 Output matching network on smith chart.

\[ S_{22} = 0.177 \angle 162.1^\circ \]
\[ S_{22}^* = 0.177 \angle 162.1^\circ \]
4.3.1

Calculating capacitor and inductor values of input matching network:

For series capacitor,

\[
\frac{1}{j\omega C} = -j \times 1.15 \times 50
\]

\[C_{\text{series}} = 0.23 \text{ pF.}\]

For shunt inductor,

\[
\frac{1}{j\omega L} = -j \frac{1.1}{50}
\]

\[L_{\text{shunt}} = 0.6 \text{ nH.}\]

Figure 4.3 Input matching network.
4.3.2

Calculating capacitor and inductor values of output matching network:

For shunt capacitor,
\[ j\omega c = \frac{0.52}{50} \]
\[ C_{\text{shunt}} = 0.13 \text{ pF}. \]

For series inductor,
\[ j\omega L = j \cdot 0.68 \times 50 \]
\[ L_{\text{series}} = 0.45 \text{ nH}. \]

Figure 4.4 Output matching network.
4.4 DC-biasing of transistor

Biasing of FET is done analytically and verified graphically as follows:

4.4.1 Analytical approach:

\[ V_{DS} = 3 \text{V}, \]
\[ I_{DS} = 30 \text{ mA}, \]
\[ V_t = -1.7 \text{ V}, \]
\[ I_{DSS} = 70\text{mA}. \]

\[ I_{DS} = K \left[ V_{GS} - V_t \right] \]

\[ K = \frac{70 \times 10^{-3}}{(-1.7)^2} \]

\[ K = 0.028 \]

So,

\[ I_{DS} = K \left[ V_{GS} - V_t \right]^2 \]

\[ V_{GS1} = -0.67 \text{ or } V_{GS2} = -2.73 \]

Since \( V_{GS} > V_t \).

\( V_{GS2} \) is invalid

Thus

\[ V_G = -3.67 \text{V}. \]

\[ V_{GG} = 3[V_G] \]

\[ V_{GG} = -11 \text{ V}. \]

Use

\[ \frac{R2}{R1+R2} = \frac{1}{3} \]

So, assuming \( R1 = 2 \text{ MΩ} \),

\[ R2 = 1 \text{ MΩ}. \]
\[ R_S = \frac{V}{I} = \frac{V_{DS} - V_G}{I_D} = 267 \, \Omega. \]

For inductance \( X_{RFC} \gg 50 \, \Omega \)

\[ jX_{RFC} = j\omega L_{RFC} \]

\[ L_{RFC} = \frac{500}{2 \times 12 \times 3.14 \times 10^9} \]

\[ L_{RFC} = 6.63 \, \text{nH}. \]

For capacitor \( X_{RFC} \ll 50 \, \Omega \)

\[ X_{DC-BLK} = \frac{1}{2 \times 3.14 \times 12 \times 10^9 \times 5} \]

\[ C_{DC-BLK} = 2.65 \, \text{pF}. \]

Figure 4.5 DC-biasing of MGF1941AL.
4.5 DC-biasing graphical representation.

From the above graph, we can conclude that analytical values are:

\[ V_{DS} = 3 \text{V}, \]
\[ I_{DS} = 30 \text{mA}, \]
\[ V_t = -1.7 \text{V}, \]
\[ I_{DSS} = 70 \text{mA}. \]
4.6 Design of power amplifier in ADS

As I selected my transistor as MGF1941AL, I designed that transistor in ADS by assigning the S-parameters from the designer’s data sheet.

![Figure 4.7 MGF1941AL schematic in ADS.](image)

After designing the schematic of MGF1941AL, we designed the amplifier with single stage.

4.7 Single stage power amplifier

As I already designed the single stage, amplifier analytically using the smith chart that consists of both input matching network and output matching network. The values of respective inductor and capacitor were calculated and verified using MATLAB. After that, I implemented the single stage amplifier in ADS.

![Figure 4.8 single stage power amplifier.](image)
4.8 Two stage power amplifier

Since we are using hybrid-90\textdegree couplers, here one amplifier attaching to the port-2 of divider to port-3 of combiner and second amplifier to the port-3 of divider to the port-4 of combiner, thereby we are constructing two stage power amplifier using hybrid-90\textdegree couplers.

Figure 4.9 Two stage power amplifier.
From the above figure, we can test our input matching and output-matching network. Therefore we can obtain the graphical value of power gain as well as vswr.

Figure 4.10 Power gain graph.
Figure 4.11 Frequency vs VSWR1
4.9 Design analysis

Figure 4.12 block diagram of power amplifier.

Linear gain, $G_A = 10.60$ dB,

$G_{1\text{dB}} = 9.60$ dB,

Insertion loss $= 0.4$ dB.

At point 1:

$P_1 = \text{output power} + \text{insertion loss} - 3 \text{ dB}$

$= 28 \text{ dBm} + 0.4 - 3 \text{ dB}$

$P_1 = 25.4 \text{ dBm}.$

At point 2:

$P_2 = \text{output power at point 1} - G_{1\text{dB}}$

$= 25.4 - 9.60$

$P_2 = 15.8 \text{ dBm}.$

At point 3:

$P_3 = \text{output power at point 2} + \text{insertion loss} + 3 \text{ dB}$

$= 15.8 + 0.4 + 3$

$P_3 = 19.2 \text{ dBm}.$

At point 4:

$P_4 = \text{output power at point 3} - \text{linear gain}$
\[ P_4 = 8.6 \text{ dBm.} \]

### 4.10 Gap-Capacitance Calculation

\[ C_e (\text{pf}) = \frac{1}{2} (h Q_1 \exp (-1.86 x S / h) \cdot \{1 + 4.19 \{1 - \exp (-0.785 (h / w_2)^{0.5} \cdot (w_1 / w_2))\}\}) \]

\[ Q_1 = 0.04598 \cdot [0.03 + (w_2 / h)^{Q_5}] \cdot (0.272 + 0.07 \varepsilon_i) \]

\[ Q_5 = 1.23 \]

Therefore,

\[ Q_1 = 0.064 \]

\[ 2.65 \times 10^{-12} = 1.6 \times 10^{-5} \cdot \exp (-1.86 \times S / 0.5 \times 10^{-3}) \cdot (2.7) \Rightarrow S = 0.4 \text{ mm} \]
4.11 The layout diagram of a power-amplifier

Figure 4.13 Layout diagram of power amplifier.
Chapter 5. Conclusion

A power amplifier using hybrid-90° couplers designed, verified and simulated using ADS, RF/MW E-book to obtain a gain of 10.6 dB and 28dBm output power. Analytical and simulation results that we obtained were satisfactory.

A low noise high electron mobility transistor MGF1941AL from the Mitsubishi manufacturer was selected such that the low noise MES FET is un-conditionally stable at an operating frequency of 12 GHz using K-Δ test. To design the input and output matching networks lumped elements were used to avoid the high sensitivity from change in frequency.

Hybrid-90° couplers were designed using microstrip line equations on a RO4003C substrate that has a dielectric constant value of 3.55 (from the Rogers Corporation). The couplers were used to divide the input signal and recombine at the output port.

Analytical calculations provided a gain of 10.6dB and an output power of 28dBm, which were verified by matlab and ADS software. The following table shows that the project goals were achieved in a satisfactory manner.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Design Goal</th>
<th>Analytical results</th>
<th>MATLAB results</th>
<th>ADS results</th>
<th>RF/MW E-book results.</th>
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<td>28dBm</td>
<td>28dBm</td>
<td>N/A</td>
<td>N/A</td>
</tr>
</tbody>
</table>

Table 3. Comparison of Results.
REFERENCES

Design and Analysis from DC to Microwaves”, Bloomington, IN 47403,
AuthorHouse 2007.

Wiley & Sons Inc., 2012

ultimate guide to superior design”, Bloomington, IN 47403, AuthorHouse 2009.

APPENDIX-A

DATA SHEET OF MGF1941AL

<Power GaAs FET>

MGF1941AL
Micro-X type plastic package

DESCRIPTION
The MGF1941AL power MES FET is designed for use in S to Ku band power amplifiers.

FEATURES
- High gain and High P1dB
  - P1dB=15dBm, Gip=10 dB (Typ.) @ f=12GHz

APPLICATION
- S to Ku band low noise amplifiers

QUALITY GRADE
- GG

RECOMMENDED BIAS CONDITIONS
- VDG=3V, ID=30mA

ORDERING INFORMATION
- Tape & reel: 4,000pieces/reel

RoHS COMPLIANT
MGF1941AL is a RoHS compliant product. RoHS compliance is indicated by the letter "G" after the Lot Marking.

ABSOLUTE MAXIMUM RATINGS

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<th>Unit</th>
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<td>-5</td>
<td>V</td>
</tr>
<tr>
<td>VGS</td>
<td>Gate to source voltage</td>
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<tr>
<td>ID</td>
<td>Drain current</td>
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<td>mA</td>
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<td>PT</td>
<td>Total power dissipation</td>
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<td>Tch</td>
<td>Channel temperature</td>
<td>175</td>
<td>°C</td>
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<tr>
<td>Tstg</td>
<td>Storage temperature</td>
<td>-65 to +150</td>
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ELECTRICAL CHARACTERISTICS

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<th>Test conditions</th>
<th>Limits</th>
<th>Unit</th>
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</thead>
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<td>ID=350μA</td>
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<td>Iss</td>
<td>Saturated drain current</td>
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<td>35</td>
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<td>Vgss(h)</td>
<td>Gate to source off voltage</td>
<td>VDS=3V, ID=300μA</td>
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<td>-1.4</td>
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<tr>
<td>P1dB</td>
<td>Output power at 1dB gain compression</td>
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<td>15</td>
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<tr>
<td>Gip</td>
<td>Linear power gain</td>
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<td>Gs</td>
<td>Associated gain</td>
<td>VDS=3V, ID=10mA, f=12GHz</td>
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<td>9</td>
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<td>NFmin</td>
<td>Minimum noise figure</td>
<td>f=12GHz</td>
<td>-1.2</td>
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Note: P1dB and Gip are tested with sampling inspection. Gs/NFmin are not tested.

Publication Date: Mar., 2012
CSTG-14954
MITSUBISHI ELECTRIC CORPORATION
Fig. 1

Top

Bottom

Side

Unit: mm

1. Gate
2. Source
3. Drain

(GD-32)
### S PARAMETERS

(Conditions: VDS=3V, ID=30mA, Ta=25degC)

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<tr>
<th>f (GHz)</th>
<th>S11 Mag</th>
<th>Angle</th>
<th>S21 Mag</th>
<th>Angle</th>
<th>S12 Mag</th>
<th>Angle</th>
<th>S22 Mag</th>
<th>Angle</th>
<th>K</th>
<th>Mag/MSG (dB)</th>
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<tbody>
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<td>5.081</td>
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Measurement plane (2.0mm)

Recommended foot print:

RO4003C/Rogers (εr=3.38, t=0.508mm)

---

Publication Date: Mar., 2012

MITSUBISHI ELECTRIC CORPORATION
# S PARAMETERS

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Measurement plane (2.6mm)
Recommended footprint:
RO4003C/Rogers (s=3.38, t=0.506mm)

Publication Date: Mar., 2012

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APPENDIX-B

Matlab code to calculate K, \( \Delta \) and U.

**MATLAB CODE**

```matlab
mS11 = 0.548;
mS12 = 0.107;
mS21 = 2.569;
mS22 = 0.177;
aS11 = 106.4 * pi/180;
aS12 = -3.6 * pi/180;
aS21 = -25.2 * pi/180;
aS22 = -162.1 * pi/180;
[rS11,iS11] = pol2cart(aS11,mS11); [rS12,iS12] = pol2cart(aS12,mS12);
[rS21,iS21] = pol2cart(aS21,mS21); [rS22,iS22] = pol2cart(aS22,mS22);
S11 = rS11 + j*iS11; S12 = rS12 + j*iS12;
S21 = rS21 + j*iS21; S22 = rS22 + j*iS22;
Delta = S11*S22 - S12*S21;
[aDelta,mDelta]=cart2pol(real(Delta),imag(Delta));
aDelta=(180*aDelta)/pi;
K = ((1 - (mS11^2) - (mS22^2) + (mDelta^2)) / (2 * mS12 * mS21));
Gtumax = ((mS21)/(mS12))*(K-((K^2)-1)^0.5);
U = (mS12*mS21*mS11*mS22)/((1-(mS11^2))*(1-(mS22^2)));
if (Delta<1 && K>1)
    disp('MES FET is unconditionally stable');
end
```

**RESULTS**

\[ K = 1.283783814642469 \]

\[ \Delta = 0.193426194298320 \]

\[ G_{\text{tumax}} = 11.494154587898464 \]

\[ U = 0.039338338911265 \]

MES FET is unconditionally stable
MATLAB CODE TO CALCULATE LUMPED ELEMENT VALUES

clear all;
Z0 = input('Enter the value of Zo \n');
f = input('Enter the of f in GHz \n');

omega = 2*pi*f;

disp('A for series L');
disp('B for shunt L');
disp('C for series C');
disp('D for shunt C');
I = input('Enter the component from the above list \n');
if (I==1)
    Xs = input('Enter the value Xs = ');
    L = Xs*(Z/(w));
    X1 = ['L = ' num2str(L) ' nH '];
    disp(X1);
elseif (I==2)
    Bp = input('Enter the value Bp = '); 
    L = Z/(w*Bp);
    X1 = ['L = ' num2str(L) ' nH '];
    disp(X1);
elseif (I==3)
    Xs = input('Enter the value of Xs = '); 
    C = 1000/(w*Z*Xs);
    X1 = ['C = ' num2str(C) ' pF '];
    disp(X1);
elseif (I==4)
    Bp = input('Enter the value of Bp = '); 
    C = 1000*Bp/(w*Z);
    X1 = ['C = ' num2str(C) ' pF '];
    disp(X1);
end
MATLAB CODE TO CALCULATE LENGTH OF TRANSMISSION LINE AND WIDTH/HEIGHT

clear all;
Z= input('Enter the value of Zo \n');
f= input('Enter the f \n');

Er = input ('Enter the value of dielectric constant\n');
height= input('Enter the value of height in mm \n');

A = (2/60)*sqrt((Er+1)/2)*((Er-1)/(Er+1))*0.23*{0.11/Er};
display(A);
W = (8*exp(A))/((exp(2*A))-2)*height;
display(W);
s=W/height;
display(s);
if s>2
  % W/h is greater than 2
  B=377*pi/(2*2*sqrt(Er));
  display(B);
  W=(2/pi)*(B-1-log(11.38)+((Er-1)/(2*Er))*(log(B-1)+0.39-(0.61/Er)))*height;
  display(W);
  s=W/height;
  display(s);
end
c = 3*10^8;
lambda0 = (c/f);
lambda = {lambda0/sqrt(Er)}*sqrt(Er/(1+0.63*(Er-1))*{W/height}^0.2155);
display(lambda);

% The length of Transmission line
l = {lambda/4};
display(l);

RESULTS

W/H = 2.235
L = 3.63*10^-3
APPENDIX-C

RF/MW E-BOOK

Large Signal Maximum Power Amplifier

1. Frequency Specification
2. Device S-Parameter
3. Gain Calculations
4. Input Matching Network
5. Output Matching Network

Return to Design Center
<table>
<thead>
<tr>
<th>Parameter</th>
<th>Mag.</th>
<th>Angle</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>S11</td>
<td>0.549</td>
<td>100°</td>
<td></td>
</tr>
<tr>
<td>S21</td>
<td>2.569</td>
<td>-35°</td>
<td></td>
</tr>
<tr>
<td>S12</td>
<td>0.002</td>
<td>-4°</td>
<td></td>
</tr>
<tr>
<td>S22</td>
<td>0.177</td>
<td>-162°</td>
<td></td>
</tr>
<tr>
<td>P14</td>
<td>15</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**Solution:**

- K = 1.38
- \( S = 0.10 \) 
- Output

Device is unconditionally stable.

Go to Step 3: Gain Design

---

Welcome to Microsoft Excel!

I'm the Office Assistant, and my job is to help you with this application.

- See key information for upgraders and new users
- Find out about the Office Assistant (That's me)
- Start using Microsoft Excel
APPENDIX-D

RO4000® Series
High Frequency Circuit Materials

RO4000® hydrocarbon ceramic laminates are designed to offer superior high frequency performance and low cost circuit fabrication. The result is a low loss material which can be fabricated using standard epoxy/glass (FR-4) processes offered at competitive prices.

The selection of laminates typically available to designers is significantly reduced once operational frequencies increase to 500 MHz and above. RO4000 material possesses the properties needed by designers of RF/microwave circuits and matching networks and controlled impedance transmission lines. Low dielectric loss allows RO4000 series material to be used in many applications where higher operating frequencies limit the use of conventional circuit board laminates. The temperature coefficient of dielectric constant is among the lowest of any circuit board material (Chart 5), and the dielectric constant is stable over a broad frequency range (Chart 2). For reduced insertion loss, LoPro™ foil is available (Chart 3). This makes it an ideal substrate for broadband applications.

RO4000 material’s thermal coefficient of expansion (CTE) provides several key benefits to the circuit designer. The expansion coefficient of RO4000 material is similar to that of copper which allows the material to exhibit excellent dimensional stability, a property needed for mixed dielectric multi-layer board constructions. The low 2-axis CTE of RO4000 laminates provides reliable plated through-hole quality, even in severe thermal shock applications. RO4000 series material has a CTE of 280°C (556°F) so its expansion characteristics remain stable over the entire range of circuit processing temperatures.

RO4000 series laminates can easily be fabricated into printed circuit boards using standard FR-4 circuit board processing techniques. Unlike PTFE based high performance materials, RO4000 series laminates do not require specialized via preparation processes such as sodium etch. This material is a rigid, thermoset laminate that is capable of being processed by automated handling systems and scrubbing equipment used for copper surface preparation.

RO4000C™ laminates are currently offered in various configurations utilizing both 1080 and 1574 glass fabric styles, with all configurations meeting the same laminate electrical performance specification. Specifically designed as a drop-in replacement for the RO4000C material, RO4550B™ laminates utilize RO4 compliant flame-retardant technology for applications requiring UL 94V-0 certification. These materials conform to the requirements of IPC-4103, slash sheet /10 for RO4003C and /11 for RO4350B materials.
<table>
<thead>
<tr>
<th>Property</th>
<th>Typical Value</th>
<th>Direction</th>
<th>Units</th>
<th>Condition</th>
<th>Test Method</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Dilectric Constant, εr</strong></td>
<td>3.38 ± 0.05</td>
<td>Z</td>
<td>--</td>
<td>10 GHz/23°C</td>
<td>IPC-TM-650 Clamped Stripline</td>
</tr>
<tr>
<td><strong>Dielectric Constant, εr</strong></td>
<td>3.48 ± 0.05</td>
<td>Z</td>
<td>--</td>
<td>8 to 40 GHz</td>
<td>Differential Phase Length Method</td>
</tr>
<tr>
<td><strong>Dissipation Factor tan δ</strong></td>
<td>0.00027</td>
<td>Z</td>
<td>--</td>
<td>10 GHz/23°C</td>
<td>IPC-TM-650 2.5.5.5</td>
</tr>
<tr>
<td><strong>Dissipation Factor tan δ</strong></td>
<td>0.00021</td>
<td>Z</td>
<td>--</td>
<td>2.5 GHz/23°C</td>
<td>IPC-TM-650 2.5.5.5</td>
</tr>
<tr>
<td><strong>Thermal Coefficient εr</strong></td>
<td>40</td>
<td>Z</td>
<td>ppm/K</td>
<td>-50°C to 150°C</td>
<td>IPC-TM-650 2.5.3.5</td>
</tr>
<tr>
<td><strong>Volume Resistivity</strong></td>
<td>1.7 X 10^6</td>
<td>Z</td>
<td>MOcm</td>
<td>COND A</td>
<td>IPC-TM-650 2.5.1.7.1</td>
</tr>
<tr>
<td><strong>Surface Resistivity</strong></td>
<td>4.2 X 10^6</td>
<td>Z</td>
<td>MO</td>
<td>COND A</td>
<td>IPC-TM-650 2.5.1.7.1</td>
</tr>
<tr>
<td><strong>Electrical Strength</strong></td>
<td>51.2</td>
<td>Z</td>
<td>KV/mm</td>
<td>0.51m 50K</td>
<td>IPC-TM-650 2.5.6.2</td>
</tr>
<tr>
<td><strong>Tensile Modulus, G</strong></td>
<td>10,050 (2.030)</td>
<td>X</td>
<td>MPa</td>
<td>RT</td>
<td>ASTM D638</td>
</tr>
<tr>
<td><strong>Tensile Modulus, G</strong></td>
<td>15,450 (2.831)</td>
<td>Y</td>
<td>MPa</td>
<td>RT</td>
<td>ASTM D638</td>
</tr>
<tr>
<td><strong>Flexural Strength</strong></td>
<td>276</td>
<td>X</td>
<td>MPa</td>
<td>RT</td>
<td>ASTM D638</td>
</tr>
<tr>
<td><strong>Dimensional Stability</strong></td>
<td>&lt; 0.35</td>
<td>X</td>
<td>1000µm/mil</td>
<td>after 23/150°C</td>
<td>IPC-TM-650 2.4.11</td>
</tr>
<tr>
<td><strong>Coefficient of Thermal Expansion</strong></td>
<td>&lt; 0.35</td>
<td>X</td>
<td>1000µm/mil</td>
<td>after 23/150°C</td>
<td>IPC-TM-650 2.4.1</td>
</tr>
<tr>
<td><strong>Tg</strong></td>
<td>280</td>
<td>X</td>
<td>°C</td>
<td>RT</td>
<td>ASTM D5850</td>
</tr>
<tr>
<td><strong>Td</strong></td>
<td>425</td>
<td>X</td>
<td>°C</td>
<td>RT</td>
<td>ASTM D5850</td>
</tr>
<tr>
<td><strong>Thermal Conductivity</strong></td>
<td>0.71</td>
<td>X</td>
<td>W/mK</td>
<td>80°C</td>
<td>ASTM C1518</td>
</tr>
<tr>
<td><strong>Moisture Absorption</strong></td>
<td>0.06</td>
<td>X</td>
<td>%</td>
<td>40 hrs immersion at 80°C</td>
<td>ASTM D570</td>
</tr>
<tr>
<td><strong>Density</strong></td>
<td>1.79</td>
<td>X</td>
<td>g/cm³</td>
<td>23°C</td>
<td>ASTM D792</td>
</tr>
<tr>
<td><strong>Copper Peel Strength</strong></td>
<td>1.05</td>
<td>X</td>
<td>N/m²</td>
<td>RT</td>
<td>IPC-TM-650 2.4.8</td>
</tr>
<tr>
<td><strong>Flammability</strong></td>
<td>N/A</td>
<td>X</td>
<td>UL</td>
<td>UL 94</td>
<td></td>
</tr>
<tr>
<td><strong>Lead-Free Process</strong></td>
<td>Yes</td>
<td>X</td>
<td>Yes</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**Notes:**
(1) The design Dk is an average number from several different coated lots of material and is the most common thickness. For detailed information, please contact Rogers Corporation or refer to Rogers' technical papers in the Rogers Technology Support Hub available at https://www.rogerscorp.com.
(2) RO4350B RoP®® laminates do not share the same UL designation as standard RO4350B laminates. A separate UL qualification may be necessary.

Typical values are a representation of an average value for the population of the property. For specific values contact Rogers Corporation.

RO4500 RoO® laminate uses a modified version of the RO4500 resin system to bond reverse treated foil. Values shown above are RO4500 laminates without the addition of the RoP® resin. For double-sided boards, the RoP® foil results in a thickness increase of approximately 0.00007" (0.18µm) and the Dk is approximately 2.4. The Dk decreases by about 0.1 as the core thickness decreases from 0.030" to 0.014.

Prolonged exposure to an oxidative environment may cause changes to the dielectric properties of polyimide based materials. The rate of change increases at higher temperatures and is highly dependent on the circuit design. Although Rogers' high frequency materials have been used successfully in innumerable applications and reports of oxidation resulting in performance problems are extremely rare, Rogers recommends that the customer evaluate each material and design combination to determine fitness for use over the entire life of the end product.
<table>
<thead>
<tr>
<th>Standard Thickness</th>
<th>Standard Panel Size</th>
<th>Standard Copper Cladding</th>
</tr>
</thead>
<tbody>
<tr>
<td>RO4003C</td>
<td>12&quot; X 18&quot; (305 X 457 mm)</td>
<td>1/2 oz. (17µm) electrodeposited copper foil (SED/5ED)</td>
</tr>
<tr>
<td>0.008&quot; (0.203mm)</td>
<td>24&quot; X 18&quot; (610 X 457 mm)</td>
<td>1 oz. (35µm) electrodeposited copper foil (LED/1ED)</td>
</tr>
<tr>
<td>0.012&quot; (0.305mm)</td>
<td>24&quot; X 36&quot; (610 X 915 mm)</td>
<td>2 oz. (70µm) electrodeposited copper foil (2ED/2ED)</td>
</tr>
<tr>
<td>0.016&quot; (0.406mm)</td>
<td>48&quot; X 36&quot; (1,219 X 915 mm)</td>
<td>PIM Sensitive Applications</td>
</tr>
<tr>
<td>0.020&quot; (0.508mm)</td>
<td>&quot;0.004&quot; (0.101mm) material is not available in panel sizes larger than 24&quot;x18&quot; (610 X 457mm)</td>
<td>1 oz (35um) LoPro Reverse Treated EDC (.5TC/STC)</td>
</tr>
<tr>
<td>0.032&quot; (0.813mm)</td>
<td></td>
<td>3 oz (45um) LoPro Reverse Treated EDC (1TC/LTC)</td>
</tr>
<tr>
<td>0.060&quot; (1.524mm)</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Note: Material clad with LoPro foil and 0.0007" (0.018mm) dielectric thickness

The information in this data sheet is intended to assist you in designing with Rogers' circuit materials. It is not intended to and does not create any warranties express or implied, including any warranty of merchantability or fitness for a particular purpose or that the results shown or this data sheet will be achieved by the user for a particular purpose. The user should determine the suitability of Rogers' circuit materials for each application.

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APPENDIX-E

Microstrip gap

A symmetrical microstrip gap can be modeled by two open ends with a capacitive series coupling between the two ends. The physical layout is shown in fig. 11.6.

The equivalent $π$-network of a microstrip gap is shown in figure 11.7. The values of the components are according to [37] and [39].

\[ C_T \; [\text{pF}] = 500 \cdot h \cdot \exp \left( -1.86 \cdot \frac{z}{h} \right) \cdot Q_1 \left( 1 + 4.39 \left( 1 - \exp \left( -0.785 \cdot \sqrt{\frac{h}{W_1}} \cdot \frac{W_2}{W_1} \right) \right) \right) \]  
\[ (11.193) \]

\[ C_{P1} = C_1 \cdot \frac{Q_2 + Q_4}{Q_2 + 1} \]
\[ (11.194) \]

\[ C_{P2} = C_2 \cdot \frac{Q_2 + Q_3}{Q_2 + 1} \]
\[ (11.195) \]

with

\[ Q_1 = 0.04958 \cdot \left( 0.03 + \left( \frac{W_2}{h} \right) \right)^{Q_6} \cdot (0.272 + 0.07 \cdot \varepsilon_r) \]
\[ (11.196) \]

\[ Q_2 = 0.307 \cdot \left( \frac{W_1}{h} + 0.9 \cdot \left( 1 + 1.05 \cdot \frac{W_1}{h} \right) \right)^{Q_7} \]
\[ (11.197) \]

\[ Q_3 = \exp \left( -0.5978 \cdot \left( \frac{W_2}{W_1} \right) \right)^{Q_2} \]
\[ (11.198) \]

\[ Q_4 = \exp \left( -0.5978 \cdot \left( \frac{W_3}{W_2} \right) \right)^{Q_2} \]
\[ (11.199) \]

\[ Q_5 = \frac{1.25}{1 + 0.12 \cdot (W_2/W_1 - 1)^{Q_6}} \]
\[ (11.200) \]
with $C_1$ and $C_2$ being the open end capacitances of a microstrip line (see eq. (11.192)). The numerical error of the capacitive admittances is less than 0.1 mS for

$$
0.1 \leq W_1 / h \leq 3 \\
0.1 \leq W_2 / h \leq 3 \\
1 \leq W_2 / W_1 \leq 3 \\
6 \leq \varepsilon_r \leq 13 \\
0.2 \leq s / h \leq 100 \\
0.2 \text{GHz} \leq f \leq 18 \text{GHz}
$$

![Figure 11.7: microstrip gap and its equivalent circuit](image)

The Y-parameters for the given equivalent small signal circuit can be written as stated in eq. (11.201) and are easy to convert to scattering parameters.

$$
Y = \begin{bmatrix}
-j\omega \cdot (C_{P1} + C_S) & -j\omega \cdot C_S \\
-j\omega \cdot C_S & j\omega \cdot (C_{P2} + C_S)
\end{bmatrix}
$$  (11.201)