MASTER

Microwave noise figure measurements on double barrier resonant tunneling diodes

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Award date:
1990

Link to publication
MICROWAVE NOISE FIGURE MEASUREMENTS ON DOUBLE BARRIER RESONANT TUNNELING DIODES
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Report no : EEA 410

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Master thesis August 1990

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ACKNOWLEDGEMENTS

I wish to thank all those whose kind advice, patience and help have supported me in my work. I have benefited from discussions with Dr. Ir. Th. G. van de Roer, Ir. H. C. Heyker, Ing. J. J. M. Kwaspen and Ir. M. F. C. Schemmann, who also were of great help in preparing and doing the measurements.
Summary

Double barrier resonant tunneling (DBRT) devices are being investigated at the Eindhoven University of Technology at the Electronic devices group. These DBRT devices have a nonlinear current voltage relationship with negative differential resistance (NDR) regions. These NDR regions can be used for amplification purposes.

In order to develop a noise parameter measurement method the DBRT one-port device, has to be incorporated into a two-port. Such a two-port is the reflection amplifier. However the incoming signals at the DBRT must be separated from the outgoing signals. This is accomplished by the use of a circulator.

The noise figure of the DBRT must be extracted from the noise figure of the reflection amplifier. To be able to do so the formula for the noise figure of the reflection amplifier is derived. This is done with two measurement different methods. The results from these two methods are compared and they coincide very well.

A closer look at the noise parameters of the DBRT shows a distinct dependence on the bias voltage, and almost no dependence on the frequency in the frequency range of 1.0 GHz - 1.5 GHz.

The lowest value of $M$ in the NDR region is 5.45. The lowest value of $F$ in the NDR region is 3.21 dB. (see table 10.22)
SYMBOL LIST

- Begin or end
- Input or output
- Decision
- Calculation
- Two unconnected lines crossing each other
- Two connected lines
- Connection identifier
- Noise generator
- Lattice constant of substance $x$
- Total energy of a particle in a quantum state
- Width of the quantum well
- Quantum state specifier
h: Planck constant
m_e: electron mass
F: noise figure
G_a: available gain
L: attenuation
Γ_i: reflection coefficient of port i
φ: determinant from Mason's gain rule
S_{ij}: 3x3 S-matrix parameter
L_{i}: loop i from Mason's gain rule
Δ: determinant from the 2x2 S matrix
S_{ij}^g: measured S-parameter
S_{ij}^n: 2x2 S-parameter
P_G: generator available power
b_G: generator wave amplitude
Z_{o1}: characteristic impedance i
P_G^\sim: available power of the new generator
b_G^\sim: new generator wave amplitude
a_i: incident wave at port i
b_i: emergent wave at port i
P_{ng}: available input noise power equal to the noise power coming from a resistor at a temperature of 290 K
P_{pg}^\sim: available output noise power
the subscript \( n \) states that the variable is a noise variable

the subscript \( G \) states that the variable is a generator variable in general

the superscript \( \sim \) states that the variable is a new generator variable

the superscript \( * \) states that the variable is the complex conjugated

\( N_D \)
noise power at the detector

\( N_{1} \)
input noise power coming from a resistor at a temperature of 290 K

\( N_{a_{1}} \)
The noise power added by element \( i \)

\( F_{45} \)
noise figure of the forward circuit part

\( F_{89} \)
noise figure of the feedback circuit part

\( M \)
noise measure

\( B \)
bandwidth

\( T_0 \)
reference temperature

\( T_{i} \)
temperature of circuit element \( i \)

\( T_s \)
noise source temperature

\( k \)
Bolzmann constant
## CONTENTS

1. Introduction  
2. Double Barrier Resonant Tunneling devices  
   2.1 Introduction  
   2.2 The quantum well  
   2.3 The I-V curve of a DBRT device  
3. The reflection amplifier  
   3.1 The reflection amplifier circuit  
   3.1.1 The circulator  
   3.1.2 The bias T  
   3.1.3 The DBRT  
   3.2 The S-matrix  
   3.3 The available gain  
4. The noisy two-port  
   4.1 The equivalent generator principle  
   4.2 Noise figure of a two-port  
   4.3 Comparison with literature  
   4.4 the adapted two-port  
   4.4.1 Comparison between the noise figures of the general two-port and the adapted two-port  
5. The amplifier circuit  
   5.1 The circuit diagram  
   5.2 The noise figure  
   5.3 Measuring the noise figure with the two attenuator method  
   5.4 Measuring the noise figure with the HP 8970B noise figure meter  
   5.4.1 Description of the noise source and the noise figure meter  
   5.4.2 The measurement method of the gain and of F  
   5.4.3 The calibration of F  

7
Measurements

6.1 Noise figure measurement using the HP NF meter
6.2 Noise figure measurement using the network analyzer and the NF meter
6.3 The comparison between both measurements
6.4 The noise measure of the DBRT
6.5 The noise figure for two frequencies

Conclusions

Recommendations for further investigations

Appendix A

9.1 Definitions
9.2 The 2x2 S-matrix
  9.2.1 The principle
  9.2.2 The calculation
  9.2.3 The calculation of \( \Gamma_D \)
9.3 The available gain
  9.3.1 The principle
  9.3.2 The calculation
9.4 The relation between \( P_G \) and \( P_G^- \)
9.5 The relation between \( P_{ng} \) and \( P_{ng}^- \)
  9.5.1 The principle
  9.5.2 The calculation
  9.5.3 The noise figure
9.6 The noise figure of the adapted two-port
9.7 The total noise figure
  9.7.1 The relation between \( |b_G|^2 \) and \( |b_G^-|^2 \)
  9.7.2 The relation between \( |b_{ng}|^2 \) and \( |b_{ng}^-|^2 \)
  9.7.3 The total noise figure
9.8 The flow-diagram of "nfcalc"
1 INTRODUCTION

The devices employing a single or multiple quantum well(s) are increasing in number. The double barrier resonant tunneling (DBRT) device is one of them. The investigation of GaAs-AlGaAs DBRT devices is part of the research of the Electronic Devices Group (EEA) of the faculty of Electrical Engineering of Eindhoven University at Eindhoven.

The current voltage relationship of a DBRT is nonlinear, and has negative differential resistance (NDR) regions. These NDR regions can be observed up to very high frequencies.

A reflection amplifier is an amplifier that makes use of a one-port with a NDR characteristic. The DBRT is such a one-port with a NDR region. The reflection amplifier is used in the frequency range \(10^8 - 10^{11}\) Hz.

or this application it is important to know the noise properties of the diode.

This report describes the development of a noise figure measurement technique and the measurements required to characterize the DBRTs noise behaviour. Chapter 2 presents some topics from the DBRT Quantum theory. Chapter 3 deals with the important features of the reflection amplifier. Chapter 4 describes the theory of the noise figure of a two-port. Some adaptations are introduced on the general two-port, in order to fit it better to the features of the reflection amplifier. Chapter 5 deals with the noise figure measurement methods. Chapter 6 describes the measurements performed on two DBRTs, and gives the measured characteristics. Chapter 7 presents the conclusions. Chapter 8 presents the recommendations for further investigations. Chapter 9 presents the derivations of the equations of the other chapters. Chapter 10 shows the results of the measurements.
2 DOUBLE BARRIER RESONANT TUNNELING DEVICES

2.1 Introduction

The Double Barrier Resonant Tunneling device (DBRT) is a so called semiconductor heterostructure. The heterostructure consists of GaAs and AlGaAs layers. The lattice constant (a) of AlGaAs obeys formula 2.1

\[ a_{\text{GaAs}} < a_{\text{AlGaAs}} < a_{\text{AlAs}} \]

\[ 5.65 \, \varepsilon < a_{\text{AlGaAs}} < 5.66 \, \varepsilon \]

This means that the interface between GaAs and AlGaAs is almost perfect and there will be very few defects. So it is possible to make an almost abrupt interface with a low defect density.

In figure 2.1 and in table 2.1 a detailed description of the layer structure of a DBRT device is given.

![Diagram of a DBRT device](image)

*Fig 2.1 The layer structure of a mesa-shaped DBRT device*
Table 2.1 The layer structure of a mesa-shaped DBRT device [1]

<table>
<thead>
<tr>
<th>Layer</th>
<th>Material</th>
<th>Doping (at/cm$^3$)</th>
<th>Thickness (nm)</th>
<th>Function</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>GaAs</td>
<td>$2.1 \times 10^{18}$</td>
<td>500</td>
<td>facilitate contact</td>
</tr>
<tr>
<td>2</td>
<td>GaAs</td>
<td>$2.1 \times 10^{16}$</td>
<td>50</td>
<td>buffer</td>
</tr>
<tr>
<td>3</td>
<td>GaAs</td>
<td>undoped</td>
<td>2.5</td>
<td>spacer</td>
</tr>
<tr>
<td>4</td>
<td>AlGaAs</td>
<td>undoped</td>
<td>5.6</td>
<td>barrier</td>
</tr>
<tr>
<td>5</td>
<td>GaAs</td>
<td>undoped</td>
<td>5.0</td>
<td>well</td>
</tr>
<tr>
<td>6</td>
<td>AlGaAs</td>
<td>undoped</td>
<td>5.6</td>
<td>barrier</td>
</tr>
<tr>
<td>7</td>
<td>GaAs</td>
<td>undoped</td>
<td>2.5</td>
<td>spacer</td>
</tr>
<tr>
<td>8</td>
<td>GaAs</td>
<td>$2.1 \times 10^{16}$</td>
<td>50</td>
<td>buffer</td>
</tr>
<tr>
<td>9</td>
<td>GaAs</td>
<td>$2.1 \times 10^{18}$</td>
<td>500</td>
<td>grown on substrate</td>
</tr>
<tr>
<td>10</td>
<td>GaAs</td>
<td>$2.1 \times 10^{18}$</td>
<td>$&gt;10^5$</td>
<td>substrate</td>
</tr>
</tbody>
</table>

* the real formula for this layer is $Al_{0.3}Ga_{0.7}As$

2.2 The quantum well

The energy gap of GaAs is smaller than the energy gap of AlGaAs. In figure 2.2 the energy-band diagram of a DBRT device is shown.

![Energy-band diagram](image)

Fig 2.2 The energy-band diagram of a DBRT device

As can be seen in figure 2.2 it is possible to construct a quantum well with succeeding layers of GaAs and AlGaAs. This is exactly what is done in a DBRT device. From quantum physics it is known that a particle in a quantum well can only have a limited number of allowed energy levels. These allowed energy levels are given by formula 2.2

$$E_n = \frac{n^2 \pi^2 \hbar^2}{m_e w^2} \quad n = 1, 2, 3, \ldots$$

where $E_n$ is the total energy of the particle in a quantum state, $n$ specifies the quantum state, $w$ is the width of the quantum well, $\hbar$ is the Planck constant and $m_e$ is the electron mass. In figure 2.3 an energy-band diagram is given with one allowed quantum state in the quantum well.
Fig 2.3 The energy-band diagram of a DBRT device with an allowed-state in the well.

2.3 The I-V curve of a DBRT device

Figure 2.4a shows a typical I-V curve of a DBRT device.

Fig 2.4 a, b  I-V curves of a DBRT device
a  + indicates a positive polarity on the mesa top
b  the four segments of the I-V curve

In figure 2.4b the I-V curve is divided into four parts. Each of these parts describes a quantum physical resonant tunneling phenomenon. In figure 2.5 the conduction band is sketched with these tunnel phenomena.
Figs 2.5 a, b, c, d The conduction band diagram of a DBRT device with several applied voltages.

a $qV_{\text{device}} = 0$

b $0 < qV_{\text{device}} < E_1$

c $qV_{\text{device}} \approx E_1$

d $E_1 < qV_{\text{device}} < E_{\text{barrier}}$
Part I of figure 2.4b corresponds with figure 2.5b. The electrons have to tunnel through the entire barrier-well-barrier structure in order to travel from the cathode to the anode. If the device bias voltage is increased the electrons which tunnel through the first barrier, enter the well with an energy that is approximately equal to the energy of the allowed state in the well. This leads to a more than proportional increase in the current with increasing device bias voltage. This situation corresponds with part II of figure 2.4b and figure 2.5c. If the device bias voltage is further increased the electrons have to tunnel through the whole barrier-well-barrier structure again. This means that the current decreases with increasing device bias voltage. This situation corresponds with part III of figure 2.4b and figure 2.5d. When the device bias voltage is increased even further a new phenomenon occurs, namely thermionic emission currents over the barrier-well-barrier structure. This leads again to an increase in current with increasing device bias voltage. This corresponds with part IV of figure 2.4b. For further information see [2, 3, 4].
3 The reflection amplifier

It is possible to use the negative resistance effects in certain one-port devices, for signal amplification. In order to achieve this amplification it is necessary to separate the incoming wave and the outgoing wave. This separation of waves is done by the circulator. Figure 3.1 shows the simplest configuration of a reflection amplifier with a DBRT.

\[ \text{Fig 3.1 The simplest configuration of a reflection amplifier.} \]

3.1 The reflection amplifier circuit [5,6,7,8]

In the reflection amplifier circuit three important elements can be distinguished:
- a circulator
- a bias T plus attenuator
- a DBRT in its package

These three elements will be dealt with in the following sub-paragraphs.

3.1.1 The circulator

The circulator type that is used in this reflection amplifier is a three-port circulator. Which is also being called Y-circulator. A simplified construction sketch is given in figure 3.1.
In the middle of the circulator a cylindrical ferrite disk is situated, between two permanent magnetic disks, forcing the signals to rotate in a certain direction around it.

The features of a circulator are:
- The DC signals are being transferred from the input port to all the other ports.
- The AC behaviour of a circulator can be divided into two regions. These regions are:
  - The frequency band it is designed for (pass band)
  - The other frequencies (stop band)

In its pass band a circulator comes closest to the ideal circulator. The further away the frequency lies from this band the worse the behaviour gets. In the pass band the attenuation for the signals traveling against the rotating direction, forced by the core, is about a hundred times greater than when the signals travel in the rotating direction. In a first order approximation the following can be said:

If the rotating direction is port 1, port 2, port 3. then the signal coming in at port 1 will travel to port 2, the signal coming in at port 2 will travel to port 3.

With this circulator behaviour, a one-port can be transformed into a two-port, because the circulator will separate the incoming and the outgoing signals (in the pass band of the circulator).

The circulator is a passive three-port. For a passive element the noise-figure can be derived according to formula 3.1.
Where \( F \) stands for the noise-figure and \( L \) stands for the attenuation. The attenuation can be derived from formula 3.2

\[
L = \frac{1}{G_a}
\]

Where \( G_a \) stands for the available gain.

### 3.1.2 The bias T

The function of the bias T in the reflection amplifier is to apply DC bias from the bias source to the DBRT diode, to allow AC signals to pass through the main line and to stop AC signals from flowing into the bias source. Figure 3.3 shows a reflection amplifier circuit with a bias T.

![Reflection Amplifier Circuit](image)

**Fig 3.3 The reflection amplifier with bias circuit.**

### 3.1.3 The DBRT

Some topics from the DBRT theory are given in chapter 2. This paragraph will only give the equivalent circuit of the DBRT plus package as given in [1].
3.2 The S-matrix

The circuit without the DBRT is a three-port, as shown in figure 3.5.

The numbering of the ports of the circulator in combination with the rotation direction, is not according to the convention. Internally the three-port element is built up out of three separate two-ports. At each port, a reflection coefficient will be found. This leads to figure 3.6
Fig 3.6 The three-port built up out of three two-ports. (see paragraph 9.1) The arrow depicts the circulator forward direction. (The numbering of the two-ports is in accordance with figure 3.7)

It is necessary to calculate the 2x2 S-matrix for all three two-ports. The variables which are necessary for the calculation of the two-port 2x2 S-matrix are:

- The three-port 3x3 S-matrix
- The three reflection coefficients

The only variable that is not obtainable by direct measurement is $\Gamma_D$ (the reflection coefficient of the DBRT). However it is possible to obtain $\Gamma_D$ by calculation. In order to be able to perform this calculation it is necessary to perform an additional measurement. The calculations of the 2x2 S-matrices and $\Gamma_D$ are performed in paragraph 9.2. The explanation of $L_1$ through $L_6$ and $\Phi$ can also be found in paragraph 9.2.
The 2x2 S-matrix of element 1 is:

\[
\begin{pmatrix}
S_{11}^# & S_{13}^# \\
S_{31}^# & S_{33}^#
\end{pmatrix}
\]

(# stands for the 2x2 S-matrix)

\[
S_{11}^# = S_{11} (1-L_2 – L_1 + L_1 – L_2) + S_{13} \Gamma_{D} S_{31} (1-L_2) + S_{12} \Gamma_{L} S_{21} (1-L_3)
\]

\[
S_{12}^# \Gamma_{L} S_{23} \Gamma_{D} S_{31} + S_{13} \Gamma_{D} S_{32} \Gamma_{L} S_{21}
\]

3.3

\[
S_{13}^# = S_{13} (1-L_2) + S_{12} \Gamma_{L} S_{23}
\]

3.4

\[
S_{31}^# = S_{31} (1-L_2) + S_{32} \Gamma_{L} S_{21}
\]

3.5

\[
S_{33}^# = S_{33} (1-L_1 - L_2 + L_1 - L_2) + S_{31} \Gamma_{C} S_{13} (1-L_2) + S_{32} \Gamma_{L} S_{23} (1-L_3)
\]

\[
S_{31} \Gamma_{C} S_{12} \Gamma_{L} S_{23} + S_{32} \Gamma_{L} S_{21} \Gamma_{C} S_{13}
\]

3.6
The 2x2 S-matrix of element 3 is:

\[
\begin{pmatrix}
S_{33}^* & S_{32}^* \\
S_{23}^* & S_{22}^*
\end{pmatrix}
\]

\[
S_{33}^* = S_{33} \Gamma S_{12} + S_{31} \Gamma S_{13} + S_{32} \Gamma S_{23} (1-L) + \Gamma S_{22}
\]

3.6

\[
S_{32}^* = S_{32} (1-L) + S_{31} \Gamma S_{12}
\]

3.7

\[
S_{23}^* = S_{23} (1-L) + S_{21} \Gamma S_{13}
\]

3.8

\[
S_{22}^* = S_{22} \Gamma S_{12} + S_{23} \Gamma S_{23} (1-L) + S_{21} \Gamma S_{12} (1-L) + \Gamma S_{22}
\]

3.9

\[
S_{23} \Gamma S_{31} + S_{21} \Gamma S_{13} + \Gamma S_{32}
\]

3.9
The 2x2 S-matrix of element 4 is:

\[
\begin{pmatrix}
S_{12} & S_{11} \\
S_{22} & S_{21}
\end{pmatrix}
\]

\[
S_{22} = \frac{S_{22}(1-L_1-L_3+L_1L_3-L_4) + S_{23} \Gamma D S_{32}(1-L_1) + S_{21} \Gamma G S_{12}(1-L_3)}{\phi}
\]

\[
S_{21} = \frac{S_{21}(1-L_3) + S_{23} \Gamma D S_{31}}{\phi}
\]

\[
S_{12} = \frac{S_{12}(1-L_3) + S_{13} \Gamma D S_{32}}{\phi}
\]

\[
S_{11} = \frac{S_{11}(1-L_2-L_3+L_2L_3-L_4)+S_{13} \Gamma D S_{31}(1-L_2)+S_{12} \Gamma G S_{21}(1-L_3)}{\phi}
\]

\[
S_{12} \Gamma L S_{23} \Gamma D S_{31} + S_{13} \Gamma D S_{32} \Gamma L S_{21}
\]

3.3
In the equations above the variables $L_1$ through $L_6$ and $\Phi$ are used. They are explained in paragraph 9.2.

The equation for $\Gamma_D$ is: (see paragraph 9.2.3)

$$\Gamma_D = \frac{S_{21} - S_9 + \Gamma S_{21} S_9 + \Gamma S_{21} S_9 + \Gamma S_{21} S_9 S_9 - S_9 \Gamma S_{21} S_9 S_9}{\Gamma S_{21} S_9 - \Gamma S_{21} S_9 S_9 - S_9 \Gamma S_{21} S_9 S_9 + S_9 \Gamma S_{21} S_9 S_9} \ldots$$

3.3 The available gain

In order to be able to perform noise-figure calculations, it is necessary to know the available gain of all four segments. These four segments are:

- From the input-port to the package of the DBRT.
- The DBRT plus package.
- From the package of the DBRT to the output-port.
- From the output-port to the input-port.

These segments are schematically given in figure 3.7.
Fig 3.7 The circuit split up into four elements. The arrows depict the circulator forward direction.

It is possible to calculate the available gain of a two-port out of the 2x2 S-matrix and the input and output reflection coefficients. The formula for the available gain is given by formula 3.13.

\[
G_a = \frac{|S_{21}|^2 \left( 1 - |\Gamma_c|^2 \right)}{|1 - S_{11} \Gamma_c|^2 - |1 - \Delta \Gamma_c|^2} \quad 3.13
\]

The derivation of formula 3.13 is given in paragraph 9.3. In figure 3.5 the element labeled 2 is the DBRT. The available gain of the DBRT can be calculated with eq. 3.14

\[
G_0 = |\Gamma_D|^2 \quad 3.14
\]

Where \(G_0\) stands for the available gain of the DBRT and \(\Gamma_D\) is the reflection coefficient of the packaged DBRT. The other three elements are two-ports each with their own 2x2 S-matrices.
This chapter will deal with the noisy two-port. In paragraph 4.1 the noisy two-port will be transformed into a noiseless two-port and noise generators attached to the two-port (equivalent generator principle). Paragraph 4.2 will explain the calculation of the noise figure of the two-port. Paragraph 4.3 will deal with the adaptations necessary for the two-port to make the overall noise figure calculation less complex.

4.1 The equivalent generator principle

In [9] a generalization of Thévenin's theorem is used, it states:

If a two-port contains internal sources, then the equivalent circuit must be modified. By a generalization of Thévenin's theorem, the two-port may be separated into a source-free network and two generators.

In [14,17,22] Peterson's equivalent noise generator theorem is used, it states:

The performance of a noisy, linear two-port network may be described completely by the addition of two noise generators to a noiseless network which is equivalent otherwise.

See also [9 through 21]

Because the \( a_i \) and \( b_i \) waves have a direct relation with voltages and currents (eq. 4.1), the generators may be \( a_{ni} \) and \( b_{ni} \) noise generators.

\[
a_i = \frac{E_i + Z_i J_i}{2 \sqrt{\text{Re}(Z_i)}} \quad 4.1a
\]

\[
b_i = \frac{E_i + Z_i^* J_i}{2 \sqrt{\text{Re}(Z_i)}} \quad 4.1b
\]

Where \( E_i \) stands for the voltage across the \( i \)th port, \( J_i \) stands for the current flowing into the \( i \)th port, \( Z_i \) stands for the characteristic impedance of the transmission line connected to port \( i \).

There are several possible noise generator configurations. They are listed below with the proper equation. [15]
Fig 4.1 The first equivalent generator configuration

\[
\begin{bmatrix}
    b_1 \\
    b_2
\end{bmatrix} = \begin{bmatrix}
    S_{11} & S_{12} \\
    S_{21} & S_{22}
\end{bmatrix} \begin{bmatrix}
    a_1 \\
    a_2
\end{bmatrix} + \begin{bmatrix}
    S_{11} & 1 \\
    S_{21} & 0
\end{bmatrix} \begin{bmatrix}
    a_{n1} \\
    b_{n1}
\end{bmatrix}
\]

Fig 4.2 The second equivalent generator configuration

\[
\begin{bmatrix}
    b_1 \\
    b_2
\end{bmatrix} = \begin{bmatrix}
    S_{11} & S_{12} \\
    S_{21} & S_{22}
\end{bmatrix} \begin{bmatrix}
    a_1 \\
    a_2
\end{bmatrix} + \begin{bmatrix}
    S_{11} & S_{12} \\
    S_{21} & S_{22}
\end{bmatrix} \begin{bmatrix}
    a_{n1} \\
    a_{n2}
\end{bmatrix}
\]

27
Fig 4.3 The third equivalent generator configuration.

\[
\begin{bmatrix}
  b_1 \\
  b_2
\end{bmatrix} = \begin{bmatrix}
  S_{11} & S_{12} \\
  S_{21} & S_{22}
\end{bmatrix}
\begin{bmatrix}
  a_1 \\
  a_2
\end{bmatrix}
+ \begin{bmatrix}
  S_{11} & 0 \\
  S_{21} & 1
\end{bmatrix}
\begin{bmatrix}
  a_{n1} \\
  b_{n2}
\end{bmatrix}
\]  

4.4

Fig 4.4 The fourth equivalent generator configuration.

\[
\begin{bmatrix}
  b_1 \\
  b_2
\end{bmatrix} = \begin{bmatrix}
  S_{11} & S_{12} \\
  S_{21} & S_{22}
\end{bmatrix}
\begin{bmatrix}
  a_1 \\
  a_2
\end{bmatrix}
+ \begin{bmatrix}
  1 & S_{12} \\
  0 & S_{22}
\end{bmatrix}
\begin{bmatrix}
  b_{n1} \\
  a_{n2}
\end{bmatrix}
\]  

4.5
Fig 4.5 The fifth equivalent generator configuration.

\[
\begin{bmatrix}
    b_1 \\
    b_2 \\
\end{bmatrix}
= 
\begin{bmatrix}
    S_{11} & S_{12} \\
    S_{21} & S_{22} \\
\end{bmatrix}
\begin{bmatrix}
    a_1 \\
    a_2 \\
\end{bmatrix}
+ 
\begin{bmatrix}
    b_{n1} \\
    b_{n2} \\
\end{bmatrix}
\]

Fig 4.6 The sixth equivalent generator configuration.

\[
\begin{bmatrix}
    b_1 \\
    b_2 \\
\end{bmatrix}
= 
\begin{bmatrix}
    S_{11} & S_{12} \\
    S_{21} & S_{22} \\
\end{bmatrix}
\begin{bmatrix}
    a_1 \\
    a_2 \\
\end{bmatrix}
+ 
\begin{bmatrix}
    0 & S_{12} \\
    1 & S_{22} \\
\end{bmatrix}
\begin{bmatrix}
    b_{n2} \\
    a_{n2} \\
\end{bmatrix}
\]

If eq. 4.2 through eq. 4.7 are compared with each other, it is clear that eq. 4.6 is the simplest. All other equations consist of an addition of two matrix-vector multiplications. Only eq. 4.6 consists of only one matrix-vector multiplication and one vector addition. Figure 4.5 is chosen as the equivalent generator configuration, for the reason stated above.
4.2 Noise figure of a two-port

Figure 4.7 shows a noise free two-port with two external noise sources. All variables necessary for the calculation of the noise figure are defined in figure 4.7.

![Diagram of a two-port with noise sources](image)

Fig 4.7 The noiseless twoport with external noise sources.

By using the noise figure, the noise behaviour of a two-port can be described with only one variable. Eq. 4.8 gives the definition of the noise figure.

\[ F = \frac{P_c}{P_{nG}} \]

where \( P_c \) stands for the available signal power at the input port, \( P_{nG} \)
stands for the input available noise power equal to the available noise power coming from a resistor at a temperature of 290 K, \( P_G \) stands for the available signal power at the output port, \( P_{ng} \) stands for the output available noise power. The relation between \( P_G \) and \( P^n_G \) is given by eq. 4.9. (see paragraph 9.4)

\[
P_G^n = \frac{|1 - \Gamma G_1|^2 |1 - S_{22} \Gamma_L|^2 \left(1 - |\Gamma G_2|^2\right)}{|S_{21}|^2 |1 - \Gamma G_2|^2 \left(1 - |\Gamma G|^2\right)}
\]

The relation between \( P_{ng} \) and \( P^n_{ng} \) is more complex as can be seen in eq. 4.10 (see paragraph 9.5).

\[
\frac{P^n_{ng}}{P_{ng}} = \frac{|\gamma|2 \left(1 - |\Gamma G|^2\right)}{|\Gamma G|^2 \left(1 - |\Gamma G_2|^2\right)} + \frac{|\gamma|2 \left(1 - |\Gamma G|^2\right)}{\left(1 - |\Gamma G_2|^2\right)} \frac{|b_{n1}|^2}{|b_{nG}|^2} + \frac{|1 - \Gamma L \Gamma_2|^2 \left(1 - |\Gamma G|^2\right)}{|1 - S_{22} \Gamma_L|^2 \left(1 - |\Gamma G_2|^2\right)} \frac{|b_{n2}|^2}{|b_{nG}|^2} + \gamma (1 - \Gamma L \Gamma_2)^*(1 - |\Gamma G|^2) \frac{b_{n1}^* b_{n2}^*}{|b_{nG}|^2}
\]

\[
\gamma^*(1 - \Gamma L \Gamma_2)(1 - |\Gamma G|^2) \frac{b_{n1}^* b_{n2}^*}{(1 - |\Gamma G_2|^2) (1 - S_{22} \Gamma_L)}
\]

Now eq. 4.9 and 4.10 must be substituted into eq. 4.8, this yields (see paragraph 9.5.3)

\[
F = 1 + |\Gamma G|^2 \frac{|b_{n1}|^2}{|b_{nG}|^2} + |\Gamma G|^2 \frac{1 - \Gamma L \Gamma_2}{|\gamma|^2 \left(1 - S_{22} \Gamma_L|^2\right)} \frac{|b_{n2}|^2}{|b_{nG}|^2} + \frac{\left(1 - \Gamma L \Gamma_2\right)^* |\Gamma G|^2 \frac{b_{n1}^* b_{n2}^*}{|b_{nG}|^2}}{\left(1 - S_{22} \Gamma_L\right)^* \gamma^* \frac{|b_{nG}|^2}{\left(1 - S_{22} \Gamma_L\right)}} \frac{1 - \Gamma L \Gamma_2 |\Gamma G|^2 b_{n1}^* b_{n2}^*}{|b_{nG}|^2}
\]

4.11
4.3 Comparison with literature

In order to compare eq. 4.11 with noise figure formulas in literature, eq. 4.11 will be written as follows

\[ F = 1 + \frac{|b_{n1}|^2}{|b_{nc}|^2} + \varepsilon \frac{|b_{n2}|^2}{|b_{nc}|^2} + \delta \frac{b_{n1}^* b_{n2}}{|b_{nc}|^2} + \gamma \frac{b_{n1}^* b_{n1}}{|b_{nc}|^2} \]  \hspace{1cm} 4.12

In [12] the general form is

\[ F = 1 + A + B E E^* + C J J^* + D (E J^* + E^* J) + F J(E J^* + E^* J) \]  \hspace{1cm} 4.13

In [10] the general form is

\[ F = 1 + A j_{n2}^2 + B e_{n2}^2 + C e_{n1} j_{n1} \]  \hspace{1cm} 4.14

In [9] the general form is

\[ F = 1 + A j_{u1}^2 + B e_{u1}^2 \]  \hspace{1cm} 4.15

(The Italic capitals stand in each article for a different value)

First eq. 4.13 will be compared with eq. 4.12

- Eq. 4.13 has a parameter A that depicts a sort of admittance mismatch. In the calculation of eq. 4.12 it was assumed that the impedance of the circuit was the same for the whole circuit.
- The third and fourth term of eq. 4.13 correspond with the second and third term of eq. 4.12
- The last two terms of eq. 4.13 correspond with the last two terms of eq. 4.12

Next eq. 4.14 will be compared with eq. 4.12

- The second and third term of eq. 4.14 correspond with the second and third term of eq. 4.12
- The last term of eq. 4.14 corresponds with the last two term of eq. 4.12

Now eq. 4.15 will be compared with eq. 4.12

- The second and third term of eq. 4.15 correspond with the second and third term of eq. 4.12
- As stated in [9] \(i_u\) and \(e\) are uncorrelated. This means that there is no corresponding term for the last two terms of eq. 4.12.
4.4 The adapted two-port

Figure 4.8 shows the adapted "two-port".

![Diagram](image)

Fig 4.8 The signal circuit diagram of the adapted "two-port".

As can be seen in figure 4.8 the output incident wave branch $a_0$ is not present. This is based on the following assumption.

As can be seen in figure 4.9 one of the circuit elements is a circulator.

![Diagram](image)

Fig 4.9 The main circuit elements. The arrow depicts the circulator rotation direction.

The circulator forward direction is the direction in which the attenuation is 20 dB less than in the reverse direction. This means that the signal power flowing into the main rotation direction will be larger than the signal flowing into the other.
direction. This also means that the influence of these signals will be small compared to the influence of the signals in the main direction. This means that the parameters corresponding with the signals flowing against the main direction will be smaller than the parameters corresponding with the signals flowing into the main direction. Keeping this in mind, the incident wave branch going against the forward direction is omitted.

The subscript branch numbers are different from those of the two-port used before, because of the fact that the numbers given here are adjusted to the numbers given to the total network presented in chapter 5.

For the three-terminal device (the fourth terminal is not connected) shown in fig. 4.8 eq. 4.8 is valid

\[
F = \frac{P_G P_{nG}^-}{P_{G}^- P_{nG}^-}
\]

Eq. 4.8 can be transformed into eq. 4.16

\[
F = \frac{|b_G|^2 |b_{nG}^-|^2}{|b_{nG}^-|^2 |b_{nG}^-|^2}
\]

(see paragraph 9.6)

The relation between \(|b_G|^2\) and \(|b_{nG}^-|^2\) is given by eq. 4.17 (see paragraph 9.6)

\[
|b_{nG}^-|^2 = \frac{|S_{89}|^2 |b_G|^2}{|1 - \Gamma_L S_{99}|^2}
\]

The relation between \(|b_{nG}^-|^2\) and \(|b_{nG}^-|^2\) is given by eq. 4.18 (see paragraph 9.6)

\[
\frac{|b_{nG}^-|^2}{|b_{nG}^-|^2} = \frac{|S_{89} (b_{nG} + \Gamma_L b_{n2}) + b_{n1}|^2}{|1 - \Gamma_L S_{99}|^2}
\]

This leads to the following formula for the noise figure
\[ F = 1 + \left| \Gamma_L \right|^2 \frac{|b_{n2}|^2}{|b_{nG}|^2} + \frac{|1 - \Gamma_L S_{99}|^2}{|S_{89}|^2} \frac{|b_{n1}|^2}{|b_{nG}|^2} + \]

\[ \frac{G_L (1 - \Gamma_L S_{99})}{S_{89}} \frac{b_{n1} b_{n2}^*}{|b_{n2}|^2} + \frac{G_L^* (1 - \Gamma_L S_{99})}{S_{89}} \frac{b_{n1}^* b_{n2}}{|b_{nG}|^2} \]

4.4.1 Comparison between noise figures of the general two-port and the adapted two-port

A comparison of eq. 4.11 and eq. 4.19 yields

- The second term in both equations are the same except for the subscripts
- When eq. 9.72 (the equation for \( \gamma \)) is substituted into the third term of eq. 4.11 the result is

\[ \frac{|1 - \Gamma_1 \Gamma_G|^2}{|S_{21}|^2} \frac{|b_{n2}|^2}{|b_{nG}|^2} \]

Because of the fact that the incident wave branch at the output port is left out, the third term in both equations will differ.
- For the same reason the last two terms in both equations will differ. The last two terms in eq. 4.11 result from the substitution of eq. 9.72 (the equation for \( \gamma \))
5 THE AMPLIFIER CIRCUIT

5.1 The circuit diagram

As mentioned in chapter 3 the amplifier circuit is a reflection amplifier. The dominating circuit elements are:
- The circulator
- The DBRT

The waves that are reflected at the output port are transported by the circulator to the input port, thus creating feedback. The amplifier circuit will now be drawn as a configuration of two two-port elements.

![Amplifier Circuit Diagram](image)

**Fig 5.1** The amplifier circuit (The arrow depicts the circulator rotation direction).

The two-port (in figure 5.1) labeled 1 depicts the feedback effect. The two-port labeled 2 depicts the forward effect.

The feedback two-port consists only of the circulator part from the output port to the input port. The forward two-port consists of:
- The circulator part from the input port to the DBRT port.
- The circuit elements between DBRT and circulator
- The DBRT.
- The circuit elements between circulator and DBRT (in the direction from the DBRT towards the circulator.
- The circulator part from the DBRT port to the output port.

The circuit diagram given in figure 5.2 is the noise circuit diagram. This is done because the signal circuit diagram can be obtained from the noise circuit diagram by leaving out the noise sources and one of the double wave names created as a result.
Fig 5.2 The total circuit diagram (The arrow depicts the circulator direction)

In figure 5.2 two matrices are given. They are:

\[ S_1 = \begin{bmatrix} s_{88} & s_{89} \\ s_{98} & s_{99} \end{bmatrix} \quad 5.1 \]

\[ S_2 = \begin{bmatrix} s_{44} & s_{45} \\ s_{54} & s_{55} \end{bmatrix} \quad 5.2 \]

Equation 5.1 deals with a 2x2 S-matrix whose incident and emergent waves have the subscript numbers 8 and 9. Equation 5.2 deals also with a 2x2 S-matrix, but the subscripts are here 4 and 5.

In paragraph 4.4 an adapted two-port is introduced. At this point the reason for this introduction will be given.

As can be seen in figure 5.2, the incident wave \( a_2 \) splits up into \( a_5 \) and \( a_9 \). This splitting up of \( a_2 \) is done by the circulator. The incident wave \( a_9 \) travels in the main rotation direction. The incident wave \( a_5 \) travels against the main rotation direction, and
suffers an attenuation that is about 20 dB greater than the attenuation which \( a_8 \) encounters. This means that the influence of the S-parameters dealing with \( a_5 \) will be smaller than the S-parameters dealing with \( a_9 \).

The incident wave \( a_4 \) splits up into \( a_4 \) and \( a_8 \). The incident wave \( a_8 \) encounters similar effects as does \( a_5 \). The incident wave \( a_4 \) encounters similar effects as does \( a_9 \).

In figure 5.3 the branches labeled \( a_8 \) and \( a_5 \) are left out, since the influence is omitted.

It must be said here that the emergent waves \( b_4, b_3, b_n3, b_9, b_n2 \) and \( b_{10} \) are also traveling in the reverse direction. However the noise sources \( b_{n2} \) and \( b_{n3} \) are important for the noise behaviour of the amplifier, so they are taken into account, as well as the other emergent wave branches.

![Fig 5.3 The circuit diagram (The arrow depicts the circulator rotation direction).](image)
5.2 The noise figure

As stated in eq. 4.16

\[ F = \frac{|b_G|^2}{|b_{nG}|^2} \]

The relation between \( |b_G|^2 \) and \( |b_{G}^\sim|^2 \) is

(see paragraph 9.7.1)

\[ \frac{|b_G|^2}{|b_{G}^\sim|^2} = \frac{|1 - S_{44} \Gamma_G|^2}{|S_{54}|^2} \]

The relation between \( |b_{nG}^\sim|^2 \) and \( |b_{nG}|^2 \) is

(see paragraph 9.7.2)

\[ \frac{|b_{nG}^\sim|^2}{|b_{nG}|^2} = \frac{S_{54} b_{nG} + S_{54} \frac{\Gamma_G}{1 - S_{44} \Gamma_G} (b_{n1} + b_{n3}) + (b_{n4} + b_{n2})}{|b_{nG}|^2} \]

The substitution of eq. 5.3 and 5.4 into eq. 4.16 yields (see paragraph 9.7.3)

\[
F = \frac{1}{|1 - S_{44} \Gamma_G|^2} + \frac{|\Gamma_G|^2}{|b_{nG}|^2} \left[ \frac{|b_{n1}|^2}{|b_{nG}|^2} + \frac{|b_{n3}|^2}{|b_{nG}|^2} + \frac{|1 - S_{44} \Gamma_G|^2}{|S_{54}|^2} \frac{|b_{n4}|^2}{|b_{nG}|^2} \right] + \frac{|b_{n2}|^2}{|S_{54}|^2} \left[ \frac{(1 - S_{44} \Gamma_G)^* \Gamma_G^*}{|b_{nG}|^2} b_{n1} b_{n2}^* + \frac{(1 - S_{44} \Gamma_G)^* \Gamma_G^*}{|b_{nG}|^2} b_{n1}^* b_{n2} \right] + \frac{(1 - S_{44} \Gamma_G)^* \Gamma_G^*}{S_{54}^* \Gamma_G^*} \frac{b_{n3} b_{n4}^*}{|b_{nG}|^2} \frac{(1 - S_{44} \Gamma_G)^* \Gamma_G^*}{S_{54}^* \Gamma_G^*} \frac{b_{n3}^* b_{n4}}{|b_{nG}|^2}
\]
Each of the two-port elements in figure 5.4 is an adapted two-port as introduced in paragraph 4.4.

The adapted amplifier circuit (The arrow depicts the circulator rotation direction).

The noise figure of the element labeled 1 is given by eq. 5.6 (see paragraph 4.4 and paragraph 9.6).

\[
F = 1 + \left| \frac{\Gamma_L}{\Gamma_n} \right|^2 + \left| \frac{\Gamma_{n_1}}{\Gamma_n} \right|^2 + \left| \frac{\Gamma_{n_2}}{\Gamma_n} \right|^2 + \left| \frac{\Gamma_{n_3}}{\Gamma_n} \right|^2 + \left| \frac{\Gamma_{n_4}}{\Gamma_n} \right|^2 
\]

\[
\left( 1 - \Gamma_L S_{99} \right) \Gamma_L \frac{b_{n_1} b_{n_2}}{b_{n_1}^* b_{n_2}} + \left( 1 - \Gamma_L S_{99} \right) \Gamma_L \frac{b_{n_1} b_{n_2}}{b_{n_1}^* b_{n_2}} 
\]

The noise figure of the element labeled 2 is given by eq. 5.7.

\[
F = 1 + \left| \frac{\Gamma_G}{\Gamma_n} \right|^2 + \left| \frac{\Gamma_{n_3}}{\Gamma_n} \right|^2 + \left| \frac{\Gamma_{n_4}}{\Gamma_n} \right|^2 + \left| \frac{\Gamma_{n_3}}{\Gamma_n} \right|^2 + \left| \frac{\Gamma_{n_4}}{\Gamma_n} \right|^2 
\]

\[
\left( 1 - \Gamma_G S_{44} \right) \Gamma_L \frac{b_{n_3} b_{n_4}}{b_{n_3}^* b_{n_4}} + \left( 1 - \Gamma_G S_{44} \right) \Gamma_L \frac{b_{n_3} b_{n_4}}{b_{n_3}^* b_{n_4}} 
\]
First eq. 5.7 will be substituted into eq. 5.5 this yields

\[
F = |1 - S_{44} \Gamma_C|^2 + \left( F_{45} - 1 \right) + \frac{|b_{n1}|^2}{|b_{nG}|^2} \left| \frac{1 - S_{44} \Gamma_C}{|S_{45}|^2} \right|^2 + \frac{|b_{n2}|^2}{|b_{nG}|^2} \left| \frac{1 - S_{44} \Gamma_C}{|S_{45}|^2} \right|^2
\]

\[
\frac{(1 - S_{44} \Gamma_C)^* \Gamma_C}{S_{45}} \cdot \frac{b_{n1} b_{n2}^*}{|b_{nG}|^2} + \frac{(1 - S_{44} \Gamma_C)^* \Gamma_C}{S_{45}} \cdot \frac{b_{n1} b_{n2}^*}{|b_{nG}|^2}
\]

Now eq. 5.6 must be substituted into eq. 5.8 giving

\[
F = |1 - S_{44} \Gamma_C|^2 + \left( F_{45} - 1 \right) + \frac{|\Gamma_C|^2 |S_{89}|^2}{|1 - S_{99} \Gamma_C|^2} \left( F_{89} - 1 \right) +
\]

\[
\frac{|b_{n2}|^2}{|b_{nG}|^2} \left( \frac{|1 - S_{44} \Gamma_C|^2}{|S_{45}|^2} - \frac{|\Gamma_L|^2 |\Gamma_C|^2 |S_{89}|^2}{|1 - \Gamma_L S_{99}|^2} \right) +
\]

\[
\frac{b_{n1} b_{n2}^*}{|b_{nG}|^2} \left( \frac{(1 - S_{44} \Gamma_C)^* \Gamma_C}{S_{45}} - \frac{\Gamma_L |\Gamma_C|^2 S_{89}}{1 - S_{99} \Gamma_L} \right) +
\]

\[
\frac{b_{n1} b_{n2}^*}{|b_{nG}|^2} \left( \frac{(1 - S_{44} \Gamma_C)^* \Gamma_C}{S_{45}} - \frac{\Gamma_L |\Gamma_C|^2 S_{89}^*}{1 - S_{99} \Gamma_L^*} \right)
\]

Eq. 5.9 is the formula for the total noise figure.

5.3 measuring the noise figure with the two-attenuator method

Basically, all techniques for measuring noise rely on measuring the output power from the circuit under two conditions of input power. Input power changes can be accomplished in basically two ways.
The first involves a change in the operating conditions of the source (see paragraph 5.4).

The second technique employs changes inside the circuit. The second technique will be used in this paragraph, in the so-called two attenuator configuration. The basic configuration is sketched in figure 5.5.

![Diagram of two attenuator configuration](image)

**Fig 5.5** The two attenuator configuration (the elements 1 and 3 are variable attenuators)

The noise power at the detector follows eq. 5.10

\[
N_D = G_1 G_2 G_3 N_1 + G_2 G_1 N_2 + G_3 N_3 + N
\]

where \(N_D\) is the noise power at the detector, \(N_1\) is the input noise power coming from a resistor at a temperature of 290 K, \(N_{a1}\) is the noise power added by attenuator 1, \(N_2\) is the noise power added by attenuator 2, \(N_{a3}\) is the noise power added by attenuator 3, \(G_1\), \(G_2\), and \(G_3\) are the available gains of the circuit elements.

Eq. 5.11 is valid for a general two-port network

\[
N_o = G N_1 + M (G - 1) k T_0 B
\]

where \(N_o\) is the output noise power, \(k\) is the Boltzmann constant, \(T_0\) is the reference temperature (mostly room temperature), \(B\) is the bandwidth.

If the network is purely resistive eq. 5.11 can be simplified

\[
N_o = G N_1 + (1 - G) k T_1 B
\]

where \(T_1\) is the temperature of the element.

The circuit elements 1 and 3 (in figure 5.5) are purely resistive.

The substitution of eq. 5.11 and 5.12 in eq. 5.10 yields

42
\[ N = G \frac{G}{1} G N + G \frac{G}{2} (1-G_1)kT_B + G \frac{G}{3} M(G_2 -1)kT_B + (1-G_3)kT_B \] 
\[ 5.13 \]

\[ N \text{ can be written as} \]
\[ N_1 = kT_s B \]
\[ 5.14 \]

where \( T_s \) is the noise source temperature

Substitution of eq. 5.14 into eq. 5.13 yields

\[ N = G \frac{G}{1} G kT_B + G \frac{G}{2} (1-G_1)kT_B + G \frac{G}{3} M(G_2 -1)kT_B + (1-G_3)kT_B \]
\[ 5.15 \]

Now the measurement method will be discussed
- First the attenuators 1 and 3 are set to into a certain attenuation value
- Next \( N \) is measured with the detector
- The setting of attenuator 1 is changed
- Now the value of attenuator 3 is changed so that the value of \( N \) is the same as the starting value

The equation for the first setting is

\[ N = G \frac{G}{1} G kT_B + G \frac{G}{2} (1-G_1)kT_B + G \frac{G}{3} M(G_2 -1)kT_B + (1-G_3)kT_B \]
\[ 5.16 \]

The equation for the second setting is

\[ N = \hat{G} \frac{G}{1} \hat{G} kT_B + \hat{G} \frac{G}{2} (1-\hat{G}_1)kT_B + \hat{G} \frac{G}{3} M(G_2 -1)kT_B + (1-\hat{G}_3)kT_B \]
\[ 5.17 \]

Eq. 5.16 and 5.17 are equal, this means

\[ G \frac{G}{1} G kT_B + G \frac{G}{2} (1-G_1)kT_B + G \frac{G}{3} M(G_2 -1)kT_B + (1-G_3)kT_B = \]
\[ \hat{G} \frac{G}{1} \hat{G} kT_B + \hat{G} \frac{G}{2} (1-\hat{G}_1)kT_B + \hat{G} \frac{G}{3} M(G_2 -1)kT_B + (1-\hat{G}_3)kT_B \]
\[ 5.18 \]

Now all terms with \( M \) must be brought to the left side of the equality sign

\[ M \left( (G_2 -1)G_3 T - (G_2 -1)\hat{G}_3 T \right) = -G_2 \left( G \frac{G}{1} \hat{G} \right) (T - T_s) + T \frac{G}{2} (\hat{G} - \hat{G}) + T \frac{G}{3} (\hat{G} - \hat{G}) \]
The result is eq. 5.19

$$M = \frac{G_2}{G_2 - 1} \left[ \frac{T_3 - T_1}{T_0 G_2} - \frac{G_{C_1} - \hat{G}_{C_1}}{T_0} \right]$$

5.4 Measuring the noise figure with the HP 8970B noise figure meter

This paragraph is divided into three parts. In the first part a description will be given of the noise source and the noise figure meter. In the second part, the measurement method the noise figure meter uses in order to obtain the available gain and the noise figure will be dealt with. In the third part the calibration routine used by the noise figure meter in order to obtain the correct noise figure will be dealt with.

5.4.1 Description of the noise source and the noise figure meter

In this sub paragraph only some topics will be brought forward, for more detailed information see [36].

The noise source is a hot/cold noise source (HP 346A). This means that the noise source can give two different noise powers, one as a cold noise source and one as a hot noise source.

The on/off switching is done by the noise figure meter (HP 8970B).

It was not equipped with a mixer.

The noise figure meter without a mixer has the following specifications:

- Frequency band usable for measurement
  10 - 1600 MHz

- Instrumentation uncertainty
  0.15 dB (gain)
  0.10 dB (noise figure)

- Range
  0 to 30 dB (noise figure)
  -20 to +40 dB (gain)
5.4.2 The measurement method of the gain and $F$

First the gain measurement method will be given and than the noise figure measurement method.

The measurement configuration is shown in figure 5.6

![Figure 5.6 The measurement configuration](image)

Before measuring the gain and the noise figure of the DUT, a calibration measurement must be conducted. The calibration configuration is sketched in figure 5.7

![Figure 5.7 The calibration configuration](image)

The output noise power is now measured (calibration configuration) with the source at $T_c$ and $T_h$. If the output noise power is plotted versus the noise source temperature, the result is figure 5.8
Fig 5.8 The output power v.s. the noise source temperature (The calibration configuration)

The slope of figure 5.8 is equal to

\[ k \cdot G \cdot B = \frac{N'_2 - N'_1}{T'_h - T'_c} \] \hspace{1cm} 5.20

Now the noise power measurement is done again (measurement configuration). The output noise power is plotted again against the noise source temperature, the result is figure 5.9.
Fig 5.9 The output noise power v.s. the noise source temperature (the measurement configuration).

The slope of figure 5.9 is

\[ k_G G_B = \frac{N_2 - N_1}{T_h - T_c} \]  

5.21

The combination of eq. 5.20 and 5.21 yields
The noise figure meter does not measure $F$ directly. Instead it measures the output noise power and calculates out of that $Y$ following eq. 5.23

$$Y = \frac{N_2}{N_1}$$

where $N_2$ stands for the output noise power when the noise source is at $T_h$ and $N_1$ stands for the output noise power when the noise source is at $T_c$.

$F$ can be calculated when $Y, T_h, T_c$ and $T_0$ are known ($T_0 = 290$ K), see eq. 5.24

$$F = \frac{\left(\frac{T_h}{T_0} - 1\right) - Y \left(\frac{T_c}{T_0} - 1\right)}{Y - 1}$$

5.4.3 The calibration of $F$

The calibration procedure is designed for the following situation (figure 5.10a)

Fig 5.10a, b The calibration configuration
First the noise figure of the configuration shown in figure 5.10a is measured. This is the noise figure of everything that will be placed in the circuit behind the DUT. This noise figure will be referred to as $F_c$. The next step is to measure the noise figure of the configuration sketched in figure 5.10b. This is the overall noise figure. It will be referred to as $F_t$. The noise figure of the DUT will be referred to as $F_{dut}$. For $F_t$ eq. 5.25 holds [37]

$$F_t = F_{dis} + \frac{(F_c - 1)}{G_{dut}}$$

where $G_{dut}$ stands for the available gain of the DUT. Eq.5.25 can be rewritten as

$$F_{dis} = F_t - \frac{(F_c - 1)}{G_{dut}}$$

where $F_{dis}$ is equal to $F_{dut}$ in this configuration.

In the rest of this paragraph the feedback effect of the reflection amplifier will not be taken into account.

The calibration of the reflection amplifier was done with a short instead of the DBRT, see figure 5.11a.

![Fig 5.11a,b The reflection amplifier and its calibration circuit](image-url)
Figure 5.11b shows the reflection amplifier with DBRT. This leads for the calibration measurement to the configuration showed in figure 5.12

![Fig 5.12 The calibration configuration](image)

The circuit element 1 in figure 5.12 consists of:
- The circulator part from the input port to port 3
- The elements between circulator port 3 and the DBRT connection plane

The circuit element 3 in figure 5.12 consists of:
- The elements between the DBRT connection port and circulator port 3
- The circulator part from port 3 to the output port 2

The noise figure measured with this configuration will be referred to as $F_c$. It can be written as

$$F_c = F_1 + \frac{F_3 - 1}{G_1} + \frac{F_4 - 1}{G_1 G_3}$$

Now the short is replaced by the DBRT. The measurement configuration is now as depicted in figure 5.13

![Fig 5.13 The measurement configuration](image)

The noise figure measured with this configuration will be referred to as $F_c$. It can be written as

$$F_c = F_1 + \frac{F_3 - 1}{G_1} + \frac{F_4 - 1}{G_1 G_3}$$
\[ F_t = F_1 + \frac{F_{\text{dut}} - 1}{G_1} + \frac{F_3 - 1}{G_1 G_{\text{dut}}} + \frac{F_4 - 1}{G_1 G_{\text{dut}} G_3} \]  

Now eq. 5.27 and 5.28 must be substituted into eq. 5.26, this yields

\[ F_{\text{dis}} = F_1 + \frac{F_{\text{dut}} - 1}{G_1} + \frac{F_3 - 1}{G_1 G_{\text{dut}}} + \frac{F_4 - 1}{G_1 G_{\text{dut}} G_3} - \frac{F_1 + \frac{F_3 - 1}{G_1} + \frac{F_4 - 1}{G_1 G_3}}{G_{\text{dut}}} \]  

The problem is that \( F_{\text{dis}} \) can be read from the display, but \( F_{\text{dut}} \) is the variable that is needed. This means that the noise figure on the display of the noise figure meter (\( F_{\text{dis}} \)) is not the noise figure of the DBRT (\( F_{\text{dut}} \)).

When eq. 5.29 is rewritten the result is

\[ F_{\text{dut}} = G_1 \left( F_{\text{dis}} - F_1 + \frac{F_1 - 1}{G_{\text{dut}}} \right) + 1 \]  

When the DUT and element 1 are attenuators than eq. 5.31 is valid for both

\[ F = 1/G \]  

This means for eq. 5.29

\[ F_{\text{dis}} = F_1 + F_1 (F_{\text{dut}} - 1) - F_{\text{dut}} (F_1 - 1) \]

\[ F_{\text{dis}} = F_{\text{dut}} \]  

This leads to the conclusion that if all elements are resistive attenuators the value on the display is the value of the noise figure of the DUT.
6 Measurements

6.1 Noise figure measurement using the HP NF meter

Before the actual measurement, calibration measurements have to be conducted first. The quantities that have to be measured during the calibration are:

- The noise figure of the circuit part from the noise source connection to the DBRT connection ($F_1$)
- The gain of that circuit part ($G_1$)
- The total gain of the circuit without the DBRT and in its place a short ($G_c$)

$F_1$ and $G_1$ can be measured with the same calibration circuit (figure 6.1)

![Fig 6.1 The first calibration configuration](image)

The results are ($F = 1.5$ GHz)

$$F_1 = 0.85 \text{ dB}$$

$$G_1 = -0.66 \text{ dB}$$
The last variable is measured with the circuit configuration shown in figure 6.2

![Diagram of circuit configuration](image)

*Fig 6.2 The second calibration configuration*

The result is

\[ G_c = -1.28 \, \text{dB} \]

Now the actual measurement on a DBRT can be reformed. The measurement configuration is depicted in figure 6.3.
Fig 6.3 NF and gain measurement configuration with a DBRT diode.

The results are shown in table 10.1
For the calculation of $F_d$ and $G_d$ (The noise figure and the available gain of the DBRT), the following formulas are used

$$G_d = \frac{G_t}{G_c}$$  \hspace{1cm} (6.1)

where $G_t$ is the gain of the circuit with the DBRT

$$F_d = 1 - G_1 \left( F_{dis} - F_1 + \frac{F_1 - 1}{G_d} \right)$$  \hspace{1cm} (6.2)

where $F_{dis}$ is the noise figure on the display of the noise figure meter (the total noise figure of the circuit)
The DBRT that was used here was the DBRT with a triangular surface (see figure 2.1) the effective area was $380 \, \mu m^2$ described in [1].
Fig 6.4 The I-V curve of the DBRT plus package $T = 290\ K$.

Fig 6.5 The $F_d-V$ characteristic of the DBRT plus package using the HP NF meter $T = 290\ K$. 

55
The L-V characteristic of the DBRT plus package using the HP NF meter $T = 290\, K$

This noise figure measurement method can only be used if the DBRT is stable not only in the pass of the circulator but also outside that band. Outside of its passband the impedance of the circulator may be far from the impedance the DBRT needs to see to be stable. In this case adjustments must be made to the circuit.

6.2 Noise figure measurement using the network analyzer and the NF meter

The measurement method that uses the network analyzer is based on eq. 5.9
The variables that have to be measured are:

- \( F_t \)
- \( \Gamma_L \) and \( \Gamma_G \)
- \( S^9_{21} \)
- all three 2x2 S-matrices

The last three items are measured with the network analyzer. The measurement configurations are given below.
Fig 6.7 The $\Gamma_c$ (reflection coefficient of the noise source) measurement configuration

Fig 6.7 The $\Gamma_L$ (reflection coefficient of the noise figure meter) measurement configuration

Fig 6.9 The $S_{21}^g$ (transfer function of the DBRT reflection amplifier) measurement configuration
The last three terms of eq. 5.9 are not accessible for direct measurement. Therefore, a calibration measurement is conducted with a short instead of the DBRT.

Now the total noise figure is measured with the DBRT and with the short. The measurement configuration is shown in figure 6.11.
Fig 6.11 a,b The overall noise figure measurement configuration

a calibration configuration with short

b measurement configuration with DBRT

The formula for the total noise figure of the circuit with the short is given in eq. 6.3
\[ F = \left| 1 - S_{44}^{\ast} \Gamma_G \right|^2 + \left( F^\circ_{45} - 1 \right) + \frac{|\Gamma_G|^2 |S_{89}|^2}{|1 - S_{99}^{\ast} \Gamma_G|^2} \left( F_{89} - 1 \right) + \]

\[ \frac{|b_{n2}|^2}{|b_{n0}|^2} \left( \frac{|1 - S_{44}^{\ast} \Gamma_G|^2}{|S_{45}^\circ|^2} - \frac{|\Gamma_L|^2 |\Gamma_G|^2 |S_{89}|^2}{|1 - \Gamma_L S_{99}^{\ast}|^2} \right) + \]

\[ \frac{b_{n1}}{b_{n0}} \frac{b_{n2}^\ast}{b_{n0}^\ast} \left( \frac{|1 - S_{44}^{\ast} \Gamma_G|^2}{S_{45}^*} - \frac{\Gamma_L |\Gamma_G|^2 S_{89}}{1 - S_{99} \Gamma_L} \right) + \]

\[ \frac{b_{n1}^\ast}{b_{n0}^\ast} \frac{b_{n2}}{b_{n0}} \left( \frac{|1 - S_{44}^{\ast} \Gamma_G|^2}{S_{45}^*} - \frac{\Gamma_L^* |\Gamma_G|^2 S_{89}^*}{1 - S_{99}^* \Gamma_L^*} \right) \]

Before eq. 6.3 is subtracted from eq. 5.9, \( F_{45} \) and \( F^\circ_{45} \) must be defined

\[ F_{45} = F_{1} + \frac{F_{d} - 1}{G_{1}} + \frac{F_{3} - 1}{G_{1} G_{d}} \]

\[ F^\circ_{45} = F_{1} + \frac{F_{3} - 1}{G_{1}} \]

The result of the subtraction is
\[ F_t - F_c = |1 - S_{44} \Gamma_G|^2 + |1 - S_{44}^* \Gamma_G|^2 + \frac{F_d - 1}{G_1} + \frac{F_3 - 1}{G_1} \left[ \frac{1}{G_d} - 1 \right] + \]

\[
\left| b_{n2} \right|^2 \frac{\left| 1 - S_{44} \Gamma_G \right|^2}{\left| S_{45} \right|^2} - \left| 1 - S_{44}^* \Gamma_G \right|^2 \frac{\left| S_{45}^* \right|^2}{\left| S_{45} \right|^2} \right] + \]

\[
\left( \frac{1 - S_{44} \Gamma_G}{S_{45}^*} \right) \Gamma_G - \left( \frac{1 - S_{44}^* \Gamma_G}{S_{45}} \right) \Gamma_G^* \right) + \]

\[
\left( \frac{1 - S_{44} \Gamma_G}{S_{45}^*} \right) \Gamma_G^* - \left( \frac{1 - S_{44}^* \Gamma_G}{S_{45}} \right) \Gamma_G \right) \]

Because of the fact that \( \Gamma_G \) is very small (see table 10.3) the last two terms are neglectable.

\[ F_{89} = 1 + \frac{|b_{n2}|^2 + |b_{n1}|^2}{G_{89} |b_{nG}|^2} \]

If \( |b_{n2}|^2 = |b_{n1}|^2 \) then

\[
\left( \frac{F_{89} - 1}{2} \right) \frac{G_{89}}{2} \approx 0.06 \]

\[
\left( \frac{1 - S_{44} \Gamma_G}{S_{45}^*} - \frac{1 - S_{44}^* \Gamma_G}{S_{45}} \right) \approx \left( \frac{1}{|S_{45}|^2} - \frac{1}{|S_{45}^*|^2} \right) \approx 0.08 \]

This means that an estimate value for the fifth term is 0.005. The
other terms are of the order $10^{-1}$ which is at least two orders of magnitude larger. This means that the fifth term is also neglectable.

The resulting equation for $F_d$ is

\[
F_d = 1 + G_d \left[ F_t - F_c + |1 - S_{44} \Gamma_0|^2 + |1 - S_{44} \Gamma_0|^2 + \frac{F_3 - 1}{G_1} \left( \frac{1}{G_d} - 1 \right) \right]
\]

Equation 6.9 is the basis for the calculation of the noise figure of the DBRT. This calculation is performed by the program "nfcalc" (see paragraph 9.8).

This formula is tested with two attenuators in the circuit shown in figure 6.12.

![Diagram of test circuit](image)

*Fig 6.12 The test circuit*

The differences with the known values were in the same order as the measurement uncertainties.

The results of the DBRT measurements are shown in table 10.2.
Fig 6.13 The I-V curve of the DBRT plus package using the network analyzer and the NF meter \( T = 290 \, \text{K} \).

Fig 6.14 The \( F_d - V \) characteristic of the DBRT plus package using the network analyzer and the NF meter \( T = 290 \, \text{K} \).
Fig 6.15 The $L_d$-$V$ characteristic of the DBRT plus package using the network analyzer and the NF meter $T = 290 \text{ K}$. 

Just like the noise figure measurement method of paragraph 6.1, there must be a stable DBRT in the circuit in order to be able to perform the measurements with this method.

6.3 The comparison between both methods

First the attenuation curve of both measurements will be compared with each other. (see figure 6.16)
Fig 6.16  The attenuation characteristics

The curves can be split into three parts:
- the first positive edge
- the negative edge
- the second positive edge

In the first part the attenuation values from the network analyzer method are lower than those from the noise figure meter method. Also the top value of the network analyzer method is lower than the value of the noise figure meter method.

In the second part the values correspond very well. Small changes in the measurement conditions may give considerable changes in the results.

In the third part the difference between the results of both methods are very small.

Now the noise figure curves of both measurements will be compared with each other.
Fig 6.17 The noise figure characteristics of the triangular DBRT.

Again the curves can be split into three parts
- the first positive edge
- the negative edge
- the second positive edge

In the first part the difference between the results of both methods is small. The difference in the top values is also small. In the second part the difference is even smaller. In the third part the results are almost equal.

The conclusion is that both methods give results with only a small difference although they are based on totally different calculations.
6.4 The noise measure of the DBRT

The measurement routine is the same as in paragraph 6.1. The DBRT is the same triangular DBRT [1]. The measurement results and the results of the calculations are given in table 10.20. The last column specifies the noise measure $M$ [38,39].

The definition for the noise measure is

\[
M = \frac{\overline{V_n}^2}{4 |\text{Re } Z| k T_0 B}
\]

Where $T_0$ is the standard temperature of 290 K, $\overline{V_n}^2$ and Re Z are given in figure 6.19 [5].

\[\text{Fig 6.18 The noise source and impedance}\]

The noise measure can also be written as

\[
M = \frac{F - 1}{1 - 1/G}
\]
Fig 6.19 The I-V curve of the DBRT plus package using the NF meter $T = 290$ K.

Fig 6.20 The $F_s$-V characteristic of the DBRT plus package using the NF meter $T = 290$ K.
Fig 6.21 The $L_d - V$ characteristic of the DBRT plus package using the NF meter $T = 290$ K.

Fig 6.22 The $M - V$ characteristic of the DBRT plus package using the NF meter $T = 290$ K.
In the last curve there are two areas (Area 1: $V \in [0, 300 \text{ mV}]$, Area 2: $V \in [1100 \text{ mV}, 1200 \text{ mV}]$) can be seen where $M$ is about constant. Now the following assumption is made:

All the noise power is coming from an equivalent noise resistance ($R_n$) and $R_n \geq 0$

This means for eq. 6.10, that $\overline{V^2}_n$ obeys eq. 6.12

$$\overline{V^2}_n = 4kT_0 R_n B$$ \hspace{1cm} 6.12

And $|\text{Re } Z|$ (eq. 6.10) is the differential resistance of the DBRT.

The combination of eq. 6.10 and 6.12 results in

$$M = \frac{4kT_0 R_n B}{4kT_0 R_d B} = \frac{R_n}{R_d}$$

$$R_n = |M R_d|$$ \hspace{1cm} 6.13

This gives for both areas the following $R_n$

<table>
<thead>
<tr>
<th>Table 6.1 The equivalent noise resistance</th>
</tr>
</thead>
<tbody>
<tr>
<td>Area 1</td>
</tr>
<tr>
<td>$V$(mV)</td>
</tr>
<tr>
<td>51</td>
</tr>
<tr>
<td>102</td>
</tr>
<tr>
<td>147</td>
</tr>
<tr>
<td>200</td>
</tr>
<tr>
<td>251</td>
</tr>
<tr>
<td>300</td>
</tr>
</tbody>
</table>

These results are depicted in figure 6.23
Fig 6.23 The equivalent noise resistance v.s. $V_{\text{bias}}$ of the DBRT.

If all the noise is shot noise then eq. 6.14 is valid

$$\frac{4 k T}{R_n} = 2 q I_{\text{dbrt}}$$

6.14

The results are
Table 6.2 The noise powers

<table>
<thead>
<tr>
<th>V(mV)</th>
<th>( \frac{4kT_0}{R_n} )</th>
<th>( 2qI_{dbrt} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>51</td>
<td>2.16 ( 10^{-24} )</td>
<td>5.44 ( 10^{-24} )</td>
</tr>
<tr>
<td>102</td>
<td>3.96 ( 10^{-24} )</td>
<td>1.28 ( 10^{-23} )</td>
</tr>
<tr>
<td>147</td>
<td>5.48 ( 10^{-24} )</td>
<td>2.27 ( 10^{-23} )</td>
</tr>
<tr>
<td>200</td>
<td>9.09 ( 10^{-24} )</td>
<td>4.03 ( 10^{-23} )</td>
</tr>
<tr>
<td>251</td>
<td>1.21 ( 10^{-23} )</td>
<td>6.53 ( 10^{-23} )</td>
</tr>
<tr>
<td>300</td>
<td>1.38 ( 10^{-23} )</td>
<td>9.76 ( 10^{-23} )</td>
</tr>
<tr>
<td>1.10 ( 10^3 )</td>
<td>1.93 ( 10^{-23} )</td>
<td>5.28 ( 10^{-22} )</td>
</tr>
<tr>
<td>1.15 ( 10^3 )</td>
<td>2.77 ( 10^{-23} )</td>
<td>6.53 ( 10^{-22} )</td>
</tr>
<tr>
<td>1.20 ( 10^3 )</td>
<td>3.36 ( 10^{-23} )</td>
<td>7.90 ( 10^{-22} )</td>
</tr>
</tbody>
</table>

The following conclusion can be taken

The shot noise power of \( I_{dbrt} \) is larger than the total noise power. This means that only a part of \( I_{dbrt} \) contributes to the shot noise power.

6.5 The noise figure for two frequencies

The first item of this paragraph is the measurement configuration. The basic configuration is the same as in paragraph 6.1. The difference is in the test fixture. The DBRT used here was not the triangular DBRT but a circular DBRT with a mesa diameter of 20 \( \mu \)m. Here it is possible to almost measure directly at the DBRT terminals without a circuit in between. The measurement procedure is the same as in paragraph 6.1, only the connectors in the DBRT branch differ. The measurement configuration is given in figure 6.24. With this DBRT in a special fixture the DBRT can be biased stable throughout the entire NDR region without relaxation oscillations and microwave oscillations. This is a requirement to conduct noise parameter measurements.
Fig 6.24 The measurement configuration for the unpackaged DBRT diode.

The measurements were conducted at two frequencies namely 1.0 GHz and 1.5 GHz.
Fig 6.25 The I-V curve of the DBRT using the NF meter $T = 290 \text{ K}$.

Figure 6.26 is the noise figure characteristic.
Before the noise figure reaches its local maximum, both curves differ only slightly. In the part between the local maximum and the local minimum both curves are almost identical. This is an important feature because it means that the noise figure in the amplification range is only slightly dependent on the frequency (in this frequency range). After the local minimum both curves start to differ.
The attenuation curves are almost equal except for high bias voltages. In the amplification area the frequency dependence is very small (in this frequency range).
The noise measure curves can be split in three parts, namely:
- The first negative part
- The positive part
- The second negative part

In the first part there is a difference in M for both frequencies. A feature that can be found here as well as in figure 6.22 is a small increase of the noise measure before the fast decrease. The explanation could lie in the numerical features of eq 6.11. In the second part both curves are almost identical. This is to be expected because the noise figure as well as the attenuation curves were here also identical. In the third part there is only a slight difference between both curves. The difference does not increase as in the noise figure and the attenuation curves.
It is possible to measure the noise figure of a DBRT

The noise figure of a DBRT has never been measured before. I haven't found an article that deals with this subject.

The DBRT must be stable in order to perform noise figure measurements. The impedance of the circulator outside of its pass band can be seen by the DBRT as a impedance which may give rise to oscillations.

The dependence of the noise measure on the frequency is in the 1.0 GHz - 1.5 GHz range very small.

In the NDR region the noise figure and the attenuation are independent of the frequency in the 1.0 GHz - 1.5 GHz range.

The measurement method using the Network Analyzer and the NF meter accounts for the feedback noise behaviour of the circulator as well as the internal reflections, the attenuation and the noise addition of the circulator.

The measurement method using the NF meter accounts for the attenuation of the circulator and the noise addition of the circulator.

The comparison between the Network Analyzer method and the Noise Figure Meter method shows that both methods give results that lay close together.

The comparison of the resulting equation (eq. 4.12) for the noise figure of a two-port is similar with equations in literature based on current-voltage noise sources.

The noise figure meter method can be conducted in less time and is easier than the network analyzer method.

In the noise figure v.s. bias voltage characteristic an area where the
noise figure decreases with increasing voltage can be detected. In the part of the noise measure v.s. voltage characteristic where $M > 0$, the decreasing edge has not the same shape as the increasing edge. The minimum of the noise measure characteristic corresponds with the largest negative differential resistance. The decreasing edge of the $M_V$ curve corresponds with the increasing negative resistance area in the I-V curve. Here in both curves the changes are larger with changes in the voltage than in the following parts. The increasing edge of the $M$-V curve corresponds with the decreasing negative differential resistance area of the I-V curve.

When $G \approx 1$ the noise measure value cannot be trusted. There are large changes in $M$ with small changes in $G$. This is due to the form of eq. 6.11.
8 Recommendations for further investigations

8.1 The theory

In chapters 4 and 5 adapted circuits have been presented. The complete circuit without these adaptations should be taken into account.

In chapter 4 one out of six possible noise source configurations is chosen. The other five configurations should be checked for a less complex resulting equation 5.9.

The theory of the noise of a DBRT should be investigated to see if the curves in chapter 6 can be supported by theory.

8.2 The measurements

The measurements of paragraph 6.4 and 6.5 should be done with several DBRTs. DBRTs equal to the one's used here, as well as DBRTs with different sizes and structures.

The stabilization of DBRTs in connection with a circulator should be investigated.

Measurements at more and higher frequencies should be done in order to gather more information on the frequency dependence of the noise figure of the DBRT.

Measurements to determine the temperature dependence of the noise behaviour of a DBRT.
Chapter 9 is a combination of the derivations of equations given in other chapters. After the definitions (paragraph 9.1), the derivation of the S-matrices is given. Out of the results of these derivations, the reflection-coefficient of the DBRT can be calculated. The next paragraph (9.3) gives the derivation of the available gain of a general two-port. This is done because noise calculations use available powers and available gains. The following paragraphs (9.4 through 9.7) deal with the noise figure calculations of a general two-port and of the reflection amplifier. The last paragraph (9.8) gives the flow diagram of the program that has been written to calculate the noise figure of the DBRT out of the noise figure of the reflection amplifier. This program is called "nfcalc"

9.1 Definitions

In this paragraph a new type of block diagram will be introduced.

![Diagram](image)

**Fig 9.1 a,b The new type of block diagram**

Figure 9.1a shows a circuit element connected at two places to transmission lines. In each of these transmission lines there are two waves, each of them traveling in the opposite direction. The waves that travel towards the element are labeled $a_1$, those traveling away from the element are labeled $b_1$. 
In the new block diagram the a and b waves are being treated as if they were separate waves, and are each assigned to a branch. In this way a two-port is derived.

For an element with more than two connections a similar block diagram can be derived. The number of ports equals the numbers of transmission lines connected.

In the next paragraphs it is assumed that the characteristic impedance is the same in all the connected transmission lines.

In this chapter the term generator plane will be used. This term can be defined as follows:

The generator plane is a fictive plane which divides a circuit into two parts, one part is the generator the other part is the load.

9.2 The 2x2 S-matrix

This paragraph will deal with the calculation of the three 2x2 S-matrices and $\Gamma_D$. The variables that are necessary for this calculation are:

- The 3x3 S-matrix
- The reflection coefficients $\Gamma_G$ and $\Gamma_L$
- $S_{21}^G$

where $S_{21}^G$ stands for the S-parameter $S_{21}$ measured between ports 1 and 2. The measurement is done with the DBRT attached to port 3. The DBRT is biased in its working point.

9.2.1 The principle

The S parameters can be regarded as a transfer function from an input node to an output node. Figure 9.2 shows the signal flow graph that can be constructed this way.
Fig 9.2 The signal flow graph of the reflection amplifier.

In [25,26] some properties of signal flow graphs are derived. In [26,27,28] Mason's gain rule is explained. In the calculation of the 2x2 S-matrices Mason's gain rule is used.

As an example the calculation of $S_{11}$ will be given here step by step.

There are eight loops in the signal flow graph shown in figure 9.2 (see paragraph 9.2.2) [29,30,31]

The first path is: $S_{11}$
The nontouching loops of this path are: $L_2$, $L_3$ and $L_6$
The loops $L_2$ and $L_3$ are also nontouching.

This yields the first numerator term:

$$S_{11} \left( 1 - L_2 - L_3 + L_2 L_3 - L_6 \right) \quad 9.1$$

The second path is: $S_{31} \Gamma_D S_{13}$
The nontouching loop is: $L_2$

This gives the second numerator term

$$S_{31} \Gamma_D S_{13} \left( 1 - L_2 \right) \quad 9.2$$

The third path is: $S_{21} \Gamma L S_{12}$
The nontouching loop is: $L_3$

This yields the third numerator term
The fourth path is: $S_{21} \Gamma_{L} S_{32} \Gamma_{D} S_{13}$
This path has only touching loops
This gives the fourth numerator term

The fifth path is: $S_{31} \Gamma_{D} S_{23} \Gamma_{L} S_{12}$
This path has only touching loops
This yields the last numerator term

eq. 9.1 through eq. 9.5 added together give the numerator. The denominator is $\phi$ (see paragraph 9.2.2). This yields for $S^*_{11}$:

\[
S^*_{11} = \frac{S_{11} (1-L_{23}+L_{32}L_{6})+ S_{31} \Gamma_{D} S_{13} (1-L_{23})+ S_{21} \Gamma_{L} S_{12} (1-L_{3})}{\phi} + \frac{S_{21} \Gamma_{L} S_{32} \Gamma_{D} S_{13} + S_{31} \Gamma_{D} S_{23} \Gamma_{L} S_{12}}{\phi}
\]
9.2.2 The calculation

In figure 9.2 eight loops can be found.

\[ L_1 = \Gamma_G S_{11} \]  \hspace{1cm} 9.6

\[ L_2 = \Gamma_L S_{22} \]  \hspace{1cm} 9.7

\[ L_3 = \Gamma_D S_{33} \]  \hspace{1cm} 9.8

\[ L_4 = \Gamma_G S_{31} \Gamma_D S_{13} \]  \hspace{1cm} 9.9

\[ L_5 = \Gamma_G S_{21} \Gamma_L S_{12} \]  \hspace{1cm} 9.10

\[ L_6 = \Gamma_D S_{23} \Gamma_L S_{32} \]  \hspace{1cm} 9.11

\[ L_7 = \Gamma_G S_{31} \Gamma_D S_{23} \Gamma_L S_{12} \]  \hspace{1cm} 9.12

\[ L_8 = \Gamma_G S_{21} \Gamma_L S_{32} \Gamma_D S_{13} \]  \hspace{1cm} 9.13

The determinant of the signal flow graph is:

\[ \phi = 1 - L_1 - L_2 - L_3 - L_4 - L_5 - L_6 - L_7 + L_1 L_2 + L_1 L_3 + L_2 L_3 + L_2 L_4 + L_3 E + L_1 + L_1 + L_1 + L_1 \]  \hspace{1cm} 9.14

eq. 9.6 up to and included eq. 9.14 yield

\[
S_{11}^* = \frac{S_{11} (1-L_1 + L_1 L_2 - L_1 L_3 + L_1 L_4 - L_1 L_5 + L_1 L_6 - L_1 L_7 + L_1 L_8)}{\phi} + \frac{S_{31} \Gamma_D S_{32} \Gamma_D S_{13}}{\phi} + \frac{S_{31} \Gamma_D S_{23} \Gamma_D S_{12}}{\phi} + S_{21} \Gamma_L S_{32} \Gamma_L S_{12}
\]  \hspace{1cm} 3.3
This calculation is performed by the program called "nfcalc". Paragraph 9.8 shows the flowdiagram of this program. This calculation is performed in the part labeled "calculation of the noise figure".
9.2.3 The calculation of \( \Gamma_D \)

The calculation of \( \Gamma_D \) is done as follows:

\( S_{21}^* \) is calculated as stated in paragraph 9.2.2. Furthermore \( S_{21}^q \) is measured, being the \( S \) parameter with the biased DBRT connected to port 3. \( S_{21}^* \) and \( S_{21}^q \) must be equal as stated in eq. 9.15

\[
S_{21}^* = S_{21}^q \tag{9.15}
\]

If eq. 3.11 is substituted into eq. 9.15 the result is eq. 9.16

\[
S_{21}^q = \frac{S_{21} \left[ 1 - L_3 \right] + S_{23} \Gamma_D S_{31}}{\phi} \tag{9.16}
\]

Now eq. 9.8 and eq. 9.14 are substituted into eq. 9.16

\[
S_{21}^q = \frac{S_{21} - S_{21} \Gamma_D S_{33} + S_{23} \Gamma_D S_{31}}{1 - \Gamma_G S_{11} - \Gamma_L S_{22} - \Gamma_D S_{33} - \Gamma_G S_{13} \Gamma_D S_{31} - \Gamma_G S_{12} \Gamma_L S_{21}} \tag{9.17}
\]

Eq. 9.17 also can be written as follows

88
Now all terms with $\Gamma_D$ are brought to the right side of the equality sign.

$$\Gamma_D \left( \Gamma_G S_{11} S_{33} S_{21}^g + \Gamma_L S_{22} S_{33} S_{21}^g - \Gamma_L S_{22} S_{13} S_{31} S_{21}^g + \right)$$

$$S_{33} \Gamma_G S_{12} S_{21} - S_{32} \Gamma_G S_{13} S_{32} S_{21}^g - \Gamma_G S_{12} S_{23} S_{31} S_{21}^g = S_{21} - S_{21} + \Gamma_G S_{11} S_{21}$$

Eq. 9.19 results in Eq. 3.12

$$\Gamma_D = \frac{S_{21} - S_{21}^g + \Gamma_G S_{11} S_{21}^g + \Gamma_G S_{11} S_{21}^g - \Gamma_G S_{11} S_{21}^g}{\Gamma_G S_{11} S_{21}^g + \Gamma_L S_{22} S_{33} S_{21}^g - \Gamma_G S_{11} S_{21}^g + \Gamma_G S_{11} S_{21}^g}$$

This calculation is performed by the program called "nfcalc" The flowdiagram of this program is given in paragraph 9.8 The calculation is performed in the part labeled "calculation of the noise figure".
9.3 The available gain

In this paragraph the available gain will be calculated from the 2x2 S-matrix and the reflection coefficients. This calculation is performed by the program "nfcalc". The calculation is performed by the flow diagram part labeled "calculation of the noise figure".

9.3.1 The principle

\[ b_G \rightarrow \Gamma_G \rightarrow G \rightarrow L \rightarrow a \]

\[ \Gamma_G, \Gamma_1 \rightarrow a_1 \rightarrow G \rightarrow a_2 \rightarrow L \rightarrow \Gamma_2, \Gamma_L \]

\[ b_1 \rightarrow G \rightarrow b_2 \rightarrow L \rightarrow \Gamma_G, \Gamma_L \]

Fig 9.3 a, b, c The shifted generator plane

Figure 9.3a shows a generator connected to a load, the available power that this generator delivers is (see [31,32,33]):
\[ P_G = \frac{|b_G|^2}{1 - |\Gamma_G|^2} \frac{1}{Z_{01}} \]

Where \( P_G \) stands for the available power of the generator, \( b_G \) stands for the generator wave amplitude, \( \Gamma_G \) stands for the reflection coefficient of the generator and \( Z_{01} \) stands for the characteristic impedance of the generator.

Figure 9.3b shows the same generator and load as in figure 9.3a. The difference is that here a two-port element is connected between the generator and the load.

Figure 9.3c shows a new generator connected to the same load as before. This new generator consists of the old generator and the two-port. This means that the generator plane has shifted from the input side of the two-port to the output side of the two-port.

The available power delivered by this new generator is:

\[ P^\sim_G = \frac{|b^\sim_G|^2}{1 - |\Gamma_G|^2} \frac{1}{Z_{02}} \]

Where \( P^\sim_G \) stands for the available power of the new generator, \( b^\sim_G \) stands for the new generator wave amplitude, \( \Gamma_G^\sim \) stands for the reflection coefficient of the new generator and \( Z_{02} \) stands for the characteristic impedance of the new generator.

The assumption that is made here is that eq. 9.23 holds for the characteristic impedances.

\[ Z_{01} = Z_{02} \]

The available gain can be calculated out of eq 9.24

\[ G_a = \frac{P^\sim_G}{P_G} \]

In order to loose all generator wave amplitudes \( b^\sim_G \) has to be expressed as a function of \( b_G \) according to eq. 9.25.

\[ b^\sim_G = C \ b_G \]
9.3.2 The calculation

In figure 9.4 the important variables are defined.

\[ r_1 \sim a_1 \quad a_2 \]

\[ S_{11} \quad S_{12} \quad S_{21} \quad S_{22} \]

\[ b_1 \quad b_2 \]

\[ \Gamma_1 \quad \Gamma_2 \]

\[ \Gamma_c \quad \Gamma_L \]

Fig 9.4 The two-port block diagram

If eq. 9.21, 9.22 and eq. 9.23 are substituted into eq. 9.24 eq 9.26 comes out.

\[
G_a = \frac{|b_1|}{1 - |\Gamma_2|^2} \frac{1}{Z_{02}} = \frac{|b_1|^2}{|b_0|^2} \frac{1}{1 - |\Gamma_c|^2} \]

9.26

The incident wave \( a_1 \) can be derived by eq. 9.27

\[ a_1 = b_G + \Gamma_c b_1 \]

9.27

This means that the incident wave \( a_1 \) is an addition of the generator wave \( b_G \) and the reflected part of the emergent wave \( b_1 \). The emergent wave \( b_1 \) is the reflected part of \( a_1 \).

\[ b_1 = \Gamma_1 a_1 \]

9.28

If eq. 9.28 is substituted into eq. 9.27, the result is eq. 9.29

\[ a_1 = b_G + \Gamma_c \Gamma_1 a_1 \]

9.29
The emergent wave $b_2$ can be derived by eq. 9.31

$$b_2 = b_{\sim} + \Gamma_2 a_2$$

This means that the emergent wave $b_2$ is an addition of the new generator wave $b_{\sim}$ and the reflected part of the incident wave $a_2$.

The incident wave $a_2$ is the reflected part of $b_2$.

$$a_2 = \Gamma_L b_2$$

If eq. 9.32 is substituted into eq. 9.31, the result is eq. 9.33

$$b_2 = b_{\sim} + \Gamma_2 \Gamma_L b_2$$

$$b_{\sim} = b_2 \left( 1 - \Gamma_2 \Gamma_L \right)$$

The S-matrix equation is given in eq. 9.35

$$b_1 = S_{11} a_1 + S_{12} a_2$$

$$b_2 = S_{21} a_1 + S_{22} a_2$$

When eq. 9.30 and 9.32 are substituted into eq. 9.35b, eq. 9.36 results

$$b_2 = \frac{S_{21} b_c}{1 - \Gamma_c \Gamma_1} + S_{22} \Gamma_L b_2$$

$$b_2 = \frac{S_{21} b_c}{\left( 1 - \Gamma_c \Gamma_1 \right) \left( 1 - S_{22} \Gamma_L \right)}$$

93
Now eq. 9.37 is substituted into eq. 9.34

\[
\hat{b}_G = \frac{S_{21} \left( 1 - \Gamma_2 \Gamma_G \right) b_G}{\left[ 1 - \Gamma_G \Gamma_1 \right] \left[ 1 - S_{22} \Gamma_L \right]} \quad 9.38
\]

In eq. 9.38 the variables \( \Gamma_1 \) and \( \Gamma_2 \) have to be replaced. \( \Gamma_1 \) is defined according to eq. 9.39

\[
\Gamma_1 = \frac{b_1}{a_1} \quad 9.39
\]

Now eq. 9.35a must be substituted into eq. 9.39

\[
\Gamma_1 = \frac{S_{11} a_1 + S_{12} a_2}{a_1} \quad 9.40
\]

Eq. 9.35b combined with eq. 9.32 gives

\[
b_2 = S_{12} a_1 + S_{22} \Gamma_L b_2
\]

\[
b_2 = \frac{S_{21} a_1}{1 - S_{22} \Gamma_L} \quad 9.41
\]

The combination of eq. 9.32, 9.40 and 9.41 gives

\[
\Gamma_1 = S_{11} + S_{12} \frac{\Gamma_L b_2}{a_1} = S_{11} + \frac{\Gamma_L S_{12} S_{21} a_1}{\left( 1 - S_{22} \Gamma_L \right) a_1}
\]
\[
\Gamma_1 = S_{11} + \frac{\Gamma_L S_{12} S_{21}}{1 - S_{22} \Gamma_L}
\]

\[
\Gamma_1 = \frac{S_{11} - \Delta \Gamma_L}{1 - S_{22} \Gamma_L}
\]

Where \( \Delta = S_{11} S_{22} - S_{12} S_{21} \)

\( \Gamma_2 \) is defined according to eq. 9.43

\[
\Gamma_2 = \frac{b_2}{a_2}
\]

Now eq. 9.35b must be substituted into eq. 9.43

\[
\Gamma_2 = \frac{S_{21} a_1 + S_{22} a_2}{a_2}
\]

\[
\Gamma_2 = S_{22} + S_{21} \frac{a_1}{a_2}
\]

Eq. 9.35a combined with

\[
a_1 = \Gamma_C b_1
\]

yields

\[
b_1 = S_{11} \Gamma_C b_1 + S_{12} a_2
\]

\[
b_1 = \frac{S_{12} a_2}{1 - S_{11} \Gamma_C}
\]

the combination of the eq. 9.44, 9.45 and 9.46 yields
\[ \Gamma_2 = S_{22} + \frac{S_{21} \Gamma_G S_{12}}{1 - S_{11} \Gamma_G} \]

\[ \Gamma_2 = \frac{S_{22} - \Delta \Gamma_G}{1 - S_{11} \Gamma_G} \quad 9.47 \]

Substituting eq. 9.42 and 9.47 into eq. 9.38 yields

\[ b\sim_G = \frac{S_{21} \left( 1 - \Gamma_L \left( S_{22} - \Delta \frac{\Gamma_G}{1 - \Gamma_G S_{11}} \right) \right) b_G}{1 - \Gamma_G \left( S_{11} - \Delta \frac{\Gamma_L}{1 - \Gamma_L S_{22}} \right)} \left( 1 - S_{22} \Gamma_L \right) \quad 9.48 \]

\[ b\sim_G = \frac{S_{21} \left( 1 - \Gamma_L \left( S_{22} - \Delta \frac{\Gamma_G}{1 - \Gamma_G S_{11}} \right) \right) b_G}{1 - S_{22} \Gamma_L - \Gamma_G \left( S_{11} - \Delta \frac{\Gamma_L}{1 - \Gamma_L S_{22}} \right)} \quad 9.49 \]

At this point the situation of eq. 9.25 is reached.
Substituting eq. 9.47 and 9.49 into eq. 9.26 yields
\[
Ga = \frac{S_{21} \left[ 1 - \left( \frac{\Gamma_L S_{22} - \Delta \Gamma_L \Gamma_G}{1 - S_{11} \Gamma_G} \right) \right]^2}{\left( 1 - S_{22} \Gamma_L - \Gamma_G S_{11} \Delta \Gamma_L \Gamma_G \right) \left( 1 - \left| b_G \right|^2 \left( 1 - \left| \Gamma_G \right|^2 \right) \right)}
\]

\[
Ga = \frac{S_{21} \left[ 1 - S_{11} \Gamma_G - \Gamma_L S_{22} - \Delta \Gamma_G \Gamma_L \right]^2}{\left( 1 - S_{11} \Gamma_G - \Gamma_L S_{22} - \Delta \Gamma_G \Gamma_L \right)^2 \left( 1 - \left| \Gamma_G \right|^2 \right)}
\]

\[
Ga = \frac{|S_{21}|^2 \left( 1 - \left| \Gamma_G \right|^2 \right)}{|1 - S_{11} \Gamma_G|^2 - |S_{22} - \Delta \Gamma_G|^2}
\]

Eq. 3.13 can also be found in [31,32]
9.4 The relation between $P_c$ and $\tilde{P}_c$

In figure 9.5 the signal circuit diagram is shown.

![Signal circuit diagram](image)

**Fig 9.5 The signal circuit diagram.**

The available output signal power is given by eq. 9.50

$$P_c^* = \frac{|b_c^*|^2}{1 - |\Gamma_2|^2} \quad 9.50$$

As stated in eq. 9.38 there is a relation between $b_c^*$ and $b_G$

$$b_c^* = \frac{S_{21} (1 - \Gamma_2 \Gamma_c) b_G}{(1 - \Gamma_c \Gamma_1)(1 - S_{22} \Gamma_L)} \quad 9.51$$

When eq. 9.51 is substituted into eq. 9.50 the result is

$$P_c^* = \frac{|S_{21}|^2 |1 - \Gamma_2 \Gamma_c|^2 |b_G|^2}{|1 - \Gamma_c \Gamma_1|^2 |1 - S_{22} \Gamma_L|^2 (1 - |\Gamma_2|^2)} \quad 9.52$$

The available input power is given by
\[
P_G = \frac{|b_G|^2}{1 - |\Gamma_G|^2}
\]

Now eq. 9.53 must be substituted into eq. 9.52 this yields

\[
\frac{P^-}{P_G} = \frac{|1 - \Gamma_G \Gamma_1|^2 | 1 - S_{22} \Gamma_L|^2 (1 - |\Gamma_2|^2)}{|S_{21}|^2 |1 - \Gamma_2 \Gamma_G|^2 (1 - |\Gamma_G|^2)}
\]

9.5 The relation between \(P_{nG}\) and \(P^-\)

9.5.1 The principle

The calculation of the relation between \(P_{nG}\) and \(P^-\) resembles the calculation of the available gain in paragraph 9.3 and paragraph 9.4. The main difference is that in paragraph 9.3 and paragraph 9.4 the calculations were with "normal" signals, here the calculations are with noise signals.

Figure 9.6 shows the two-port with generator and load.

![Two-port with generator and load](image)

**Fig 9.6** The two-port with generator and load.

The available generator noise power \(P_{nG}\) is defined by eq. 9.54

\[
P_{nG} = \frac{|b_{nG}|^2}{1 - |\Gamma_G|^2}
\]
The available output noise power $P_{n^G}$ (new generator noise power) is defined by eq. 9.55

$$P_{n^G} = \frac{|b_{n^G}|^2}{1 - |\Gamma_2|^2} \quad 9.55$$

First $b_{n^G}$ will be expressed as a function of $b_{n^G}$, $b_{n1}$ and $b_{n2}$ as shown in eq. 9.56

$$b_{n^G} = \Re b_{n^G} + \Im b_{n1} + \Omega b_{n2} \quad 9.56$$

Because of the fact that the measurement will be conducted in a small frequency band, the noise signals may be treated as if they were deterministic signals [Th. van de Roer private communication].

It must be stated here that $\Gamma_{n^G}$ is equal to and identical with $\Gamma_2$.

### 9.5.2 The calculation

In this paragraph all variables used in the calculation are noise variables. Figure 9.7 shows the noise circuit diagram.

![Noise Circuit Diagram](image)

**Fig 9.7 The noise circuit diagram**

For the incident wave $a_1$, eq. 9.57 holds

$$a_1 = b_{n^G} + \Gamma_{n^G} b_1 \quad 9.57$$

The emergent wave $b_1$ is an addition of two other waves, as stated in
eq. 9.58

\[ b_1 = b_3 + b_{n_1} \]

Now eq. 9.58 must be substituted into eq. 9.59

\[ a_1 = b_{nG} + \Gamma_G (b_3 + b_{n_1}) \]

Eq. 9.28 has to be revised to the changed situation because of the addition of two noise sources, it becomes:

\[ b_3 = \Gamma_1 a_1 \]

Now eq. 9.60 must be substituted into eq. 9.59, this yields

\[ a_1 = b_{nG} + \Gamma_G b_{n_1} \]

\[ a_1 = \frac{b_{nG} + \Gamma_G b_{n_1}}{1 - \Gamma_G \Gamma_1} \]

Figure 9.8 shows the new generator.

---

**Fig 9.8** The new noise generator with load.

For the emergent wave \( b_2 \) eq. 9.62 yields

\[ b_2 = b_{\sim G} + a_2 \Gamma_2 \]

This means that \( b_2 \) is an addition of the internal noise wave \( b_{\sim G} \) plus the reflected part from the incident noise wave \( a_2 \). The internal noise wave \( b_{\sim G} \) is built up out of \( b_{nG} \) plus the nonreflected part of \( a_1 \).
Furthermore \( b_2 \) is an addition of two waves (see figure 9.6)

\[
b_2 = b_4 + b_{n2}
\]

As eq. 9.32 states

\[
a_2 = \Gamma L b_2
\]

When eq. 9.62, 9.63 and 9.32 are combined the result is

\[
b_4 + b_{n2} = b_n^\sim + b_4 \Gamma L \Gamma_2 + b_{n2} \Gamma L \Gamma_2
\]

\[
b_4 = \frac{b_n^\sim - b_{n2} (1 - \Gamma L \Gamma_2)}{1 - \Gamma L \Gamma_2}
\]

Eq. 9.64 can also be written as in eq. 9.65

\[
b_n^\sim = b_4 (1 - \Gamma L \Gamma_2) - b_{n2} (1 - \Gamma L \Gamma_2)
\]

Just like eq. 9.28, eq. 9.35 has to be revised, it becomes

\[
b_3 = S_{11} a_1 + S_{12} a_2 \quad 9.66a
\]

\[
b_4 = S_{21} a_1 + S_{22} a_2 \quad 9.66b
\]

The substitution of eq. 9.61, 9.63 and 9.32 into eq. 9.66b yields

\[
b_4 = \frac{S_{21} (b_n^\sim + \Gamma G b_{n1})}{1 - \Gamma G \Gamma_1} + S_{22} \Gamma L b_4 + S_{22} \Gamma L b_{n2}
\]

\[
b_4 = \frac{S_{21} (b_n^\sim + \Gamma G b_{n1})}{(1 - \Gamma G \Gamma_1)(1 - S_{22} \Gamma L)} + \frac{S_{22} \Gamma L b_{n2}}{1 - S_{22} \Gamma L}
\]

Now eq. 9.67 must be substituted into eq 9.65
\[ b_{nG}^- = \frac{(1 - \Gamma_L \Gamma_2) S_{21} (b_{nG} + \Gamma_G b_{n1})}{(1 - \Gamma_G \Gamma_1)(1 - S_{22} \Gamma_L)} + \frac{(1 - \Gamma_L \Gamma_2) S_{22} L b_{n2}}{(1 - S_{22} \Gamma_L)} + b_{n2} (1 - \Gamma_L \Gamma_2) \]

\[ b_{nG}^- = \frac{(1 - \Gamma_L \Gamma_2) S_{21} (b_{nG} + \Gamma_G b_{n1})}{(1 - \Gamma_G \Gamma_1)(1 - S_{22} \Gamma_L)} + \frac{(1 - \Gamma_L \Gamma_2) b_{n2}}{(1 - S_{22} \Gamma_L)} \]

Eq. 9.64 can also be written as

\[ b_{nG}^- = \frac{(1 - \Gamma_L \Gamma_2) S_{21}}{(1 - \Gamma_G \Gamma_1)(1 - S_{22} \Gamma_L)} b_{nG} + \frac{(1 - \Gamma_L \Gamma_2) \Gamma_G S_{22}}{(1 - \Gamma_G \Gamma_1)(1 - S_{22} \Gamma_L)} b_{n1} + \]

\[ \frac{(1 - \Gamma_L \Gamma_2)}{(1 - S_{22} \Gamma_L)} b_{n2} \]

Eq. 9.69 has the form predicted in eq. 9.56. When eq. 9.69 is substituted into eq. 9.55 the result is

\[ P_{nG}^- = \frac{(1 - \Gamma_L \Gamma_2) S_{21}}{(1 - \Gamma_G \Gamma_1)(1 - S_{22} \Gamma_L)} b_{nG} + \frac{(1 - \Gamma_L \Gamma_2) \Gamma_G S_{22}}{(1 - \Gamma_G \Gamma_1)(1 - S_{22} \Gamma_L)} b_{n1} + \frac{(1 - \Gamma_L \Gamma_2)}{(1 - S_{22} \Gamma_L)} b_{n2} \]

\[ 1 - |\Gamma_2|^2 \]

9.70

Before the numerator is written out, a few assumptions are made. These assumptions are:

- The equivalent noise generator \( b_{n1} \) and \( b_{n2} \) are correlated.
- The equivalent noise generator \( b_{n1} \) and the noise generator \( b_{nG} \) are uncorrelated.
- The equivalent noise generator \( b_{n2} \) and the noise generator \( b_{nG} \) are uncorrelated.

The first assumption comes from the fact that both equivalent noise sources have their origin in the same noisy two-port. The other two
assumptions come from the fact that the external generator is totally independent from the two-port. Now eq. 9.70 will be written out.

\[
P_{\text{ng}}^{-} = \frac{|1 - \Gamma_L \Gamma_2|^2 |S_{21}|^2}{|1 - \Gamma_1 \Gamma_2|^2 |1 - S_{22} \Gamma_L|^2 (1 - |\Gamma_2|^2)} |b_{\text{ng}}|^2 + \]

\[
\frac{|1 - \Gamma_1 \Gamma_2|^2 S_{21} \Gamma_2}{|1 - S_{22} \Gamma_L|^2 (1 - |\Gamma_2|^2) (1 - \Gamma_1 \Gamma_2)} b_{n1}^* b_{n2} + \frac{|1 - \Gamma_1 \Gamma_2|^2 S_{21} \Gamma_2^*}{|1 - S_{22} \Gamma_L|^2 (1 - |\Gamma_2|^2) (1 - \Gamma_1 \Gamma_2)} b_{n1} b_{n2}
\]

9.71

Here a new variable will be introduced in order to increase the survey-ability of eq. 9.74

\[
\gamma = \frac{S_{21} \Gamma_2 (1 - \Gamma_2 \Gamma_L)}{(1 - \Gamma_1 \Gamma_2)(1 - S_{22} \Gamma_L)}
\]

9.72

When eq. 9.72 is substituted into eq. 9.71 the result is

\[
P_{\text{ng}}^{-} = \frac{|\gamma|^2}{|\Gamma_2|^2 (1 - |\Gamma_2|^2)} |b_{\text{ng}}|^2 + \frac{|\gamma|^2}{(1 - |\Gamma_2|^2)} |b_{n1}|^2 + \]

\[
\frac{|1 - \Gamma_L \Gamma_2|^2}{|1 - S_{22} \Gamma_L|^2 (1 - |\Gamma_2|^2) (1 - \Gamma_1 \Gamma_2)} \frac{\gamma (1 - \Gamma_L \Gamma_2)^*}{(1 - |\Gamma_2|^2) (1 - \Gamma_1 \Gamma_2)} b_{n1} b_{n2}^*
\]

\[
\frac{\gamma^* (1 - \Gamma_L \Gamma_2)}{(1 - |\Gamma_2|^2) (1 - S_{22} \Gamma_L)} b_{n1}^* b_{n2}
\]

9.73
The difference in the first two terms can be explained by the fact that only a part of \( b_{n1} \) namely \( \Gamma_G b_{n1} \) will add to the noise power at the input node of port 1.

The fact that not all of \( b_{n2} \) adds to the output noise power is originating from the reflections at the load and the output port of the two-port.

Both \( P_{nG} \) and \( P_{nG}^\sim \) are now known, so the relation between both can be calculated.

\[
\frac{P_{nG}^\sim}{P_{nG}} = \frac{|\gamma|^2 (1 - |\Gamma_G|^2)}{|\Gamma_G|^2 (1 - |\Gamma_2|^2)} + \frac{|\gamma|^2 (1 - |\Gamma_G|^2)}{(1 - |\Gamma_2|^2)} \frac{|b_{n1}|^2}{|b_{nG}|^2} + \frac{|b_{n2}|^2}{|b_{nG}|^2}
\]

\[
\frac{|1 - \Gamma_L \Gamma_2|^2 (1 - |\Gamma_G|^2)}{|1 - S_{22} \Gamma_L|^2 (1 - |\Gamma_2|^2)} \frac{|b_{n2}|^2}{|b_{nG}|^2} + \frac{\gamma^*(1 - \Gamma_L \Gamma_2)(1 - |\Gamma_G|^2)}{(1 - |\Gamma_2|^2)(1 - S_{22} \Gamma_L)} \frac{b_{n1}^* b_{n2}}{|b_{nG}|^2} + \frac{\gamma (1 - \Gamma_L \Gamma_2)^*(1 - |\Gamma_G|^2)}{(1 - |\Gamma_2|^2)(1 - S_{22} \Gamma_L)^*} \frac{b_{n1}^* b_{n2}}{|b_{nG}|^2}
\]

9.5.3 The noise figure

First eq. 9.72 is substituted into eq. 4.9 the result is

\[
\frac{|P_G|^2}{|P_{nG}|^2} = \frac{|\Gamma_G|^2 (1 - |\Gamma_2|^2)}{|\gamma|^2 (1 - |\Gamma_G|^2)} \frac{F}{P_{nG}} \frac{P_{nG}}{P_{nG}^\sim} \frac{P_{nG}^\sim}{P_{nG}^\sim} 9.74
\]

As mentioned in chapter 4 the noise figure can be written as

\[
F = \frac{P_G / P_{nG}}{P_{nG} / P_{nG}^\sim} \frac{P_{nG}^\sim}{P_{nG}} 4.8
\]
Now eq. 9.74 and 4.10 can be substituted into eq. 4.8 resulting in

\[ F = 1 + \frac{|\Gamma_c|^2}{|b_{n1}|^2} + \frac{|\Gamma_c|^2}{|\gamma|^2} \frac{1 - \Gamma_L \Gamma_2}{1 - S_{22} \Gamma_L^2} + \frac{|b_{n2}|^2}{|b_{nG}|^2} \]

\[ \frac{\left(1 - \Gamma_L \Gamma_2\right)^* |\Gamma_c|^2 b_{n1} b_{n2}^*}{\left(1 - S_{22} \Gamma_L\right)^* |\gamma|^2 |b_{nG}|^2} + \frac{\left(1 - \Gamma_L \Gamma_2\right) |\Gamma_c|^2 b_{n1} b_{n2}}{\left(1 - S_{22} \Gamma_L\right) |\gamma| |b_{nG}|^2} \]

4.11

9.6 The noise figure of the adapted two-port

Figure 9.9 shows the adapted two-port

\[ \Gamma_c \quad \Gamma_1 \]

\[ \text{Fig 9.9 The adapted "two-port" (three-terminals the fourth is not connected).} \]

Eq. 4.8 is also valid for this adapted two-port. If eq. 9.50, 9.53, 9.54 and 9.55 are substituted into eq. 4.8 the result is

\[ F = \frac{|b_c|^2}{1 - |\Gamma_c|^2} \frac{|b_n|}{1 - |\Gamma_2|^2} \frac{|b_{nG}|^2}{1 - |\Gamma_c|^2} \frac{|b_{nG}|^2}{1 - |\Gamma_2|^2} \]

106
Next the relation between $|b_c|^2$ and $|b_{c^*}|^2$ is calculated. Both $b_c$ and $b_{c^*}$ are signal waves, so figure 9.10 shows the signal circuit diagram.

$$F = \frac{|b_c|^2 |b_{c^*}|^2}{|b_{c^*}|^2 |b_c|^2}$$  \hspace{1cm} \text{(4.16)}$$

Fig 9.10 The signal circuit diagram of the Three-terminal device.

Eq. 9.11 is no longer valid it becomes

$$b_c = b_{c^*}$$  \hspace{1cm} \text{(9.75)}$$

In eq. 9.27 only the subscripts have changed

$$a_g = b_c + \Gamma_\text{L} b_g$$  \hspace{1cm} \text{(9.76)}$$

Eq. 9.35 also changes

$$b_g = S_{g9} a_g$$  \hspace{1cm} \text{(9.77a)}$$

$$b_8 = S_{89} a_g$$  \hspace{1cm} \text{(9.77b)}$$

$S_{g9}$ and $S_{89}$ are not elements of a 9x9 S-matrix, but they are elements of a 2x2 S-matrix whose port numbers are 8 and 9. In order to be able to distinguish between the 8-9 two-port and the 4-5 two-port (see paragraph 9.7) the matrix numbers are identical with the branch numbers. Now eq. 9.77a can be substituted into eq. 9.76

$$a_g = b_c + \Gamma_\text{L} S_{g9} a_g$$
\[
a_g = \frac{b_c}{1 - \Gamma_L S_{gg}}
\]

Now eq. 9.78 must be substituted into eq. 9.77b

\[
b_7 = \frac{S_{gg} b_c}{1 - \Gamma_L S_{gg}}
\]

Eq. 9.79 must be substituted into eq. 9.75

\[
b_{\sim c} = \frac{S_{gg} b_c}{1 - \Gamma_L S_{gg}}
\]

This means for \(|b_{\sim c}|^2\)

\[
|b_{\sim c}|^2 = \frac{|S_{gg}|^2 |b_c|^2}{1 - \Gamma_L S_{gg}}
\]

The noise circuit diagram of the adapted two-port is given by figure 9.11

![Noise Circuit Diagram](image)

**Fig 9.11** The noise circuit diagram of the adapted "two-port".

Eq. 9.62 is no longer valid it becomes
\[ b_7 = b_{nG} \]  

9.81

\( b_7 \) is an addition of two waves

\[ b_7 = b_8 + b_{n1} \]  

9.82

Substituting eq. 9.77b and 9.82 into eq. 9.81 yields

\[ b_{nG} = S_{99} a_9 + b_{n1} \]  

9.83

The only changes that have to be made in eq. 9.57 are the subscripts, it becomes

\[ a_g = b_{nG} + \Gamma_L b_{10} \]  

9.84

The emergent wave \( b_{10} \) is an addition of two waves

\[ b_{10} = b_{n2} + b_9 \]  

9.85

Now eq. 9.85 and 9.77a are substituted into eq. 9.84

\[ a_g = b_{nG} + \Gamma_L b_2 + \Gamma_L S_{99} a_9 \]  

\[ a_g = \frac{b_{nG} + \Gamma_L b_{n2}}{1 - \Gamma_L S_{99}} \]  

9.86

Now eq. 9.86 must be substituted into eq. 9.83 the result is

\[ b_{nG} = \frac{S_{99} (b_{nG} + \Gamma_L b_{n2})}{1 - \Gamma_L S_{99}} + b_{n1} \]  

9.87

This yields for \( |b_{nG}^*|^2 \)

\[ |b_{nG}^*|^2 = \left| \frac{S_{99} (b_{nG} + \Gamma_L b_{n2})}{1 - \Gamma_L S_{99}} + b_{n1} \right|^2 \]  

9.88

109
This means for the relation between \( |b_{\text{nc}}^\sim|^2 \) and \( |b_{\text{nc}}|^2 \):

\[
\frac{|b_{\text{nc}}^\sim|^2}{|b_{\text{nc}}|^2} = \frac{S_{99}(b_{\text{nc}} + \Gamma_L b_{n2})}{1 - \Gamma_L S_{99}} + b_{n1}^2
\]

4.18

Substituting eq. 4.17 and 4.18 into eq. 4.16 yields

\[
F = \frac{|1 - \Gamma_L S_{99}|^2}{|S_{89}|^2} \left( \frac{|S_{99}|^2 |b_{\text{nc}}|^2}{|1 - \Gamma_L S_{99}|^2 |b_{\text{nc}}|^2} + \frac{|S_{89}|^2 |\Gamma_L|^2 |b_{n2}|^2}{|1 - \Gamma_L S_{99}|^2 |b_{nc}|^2} \right)
\]

9.89

The assumptions made in paragraph 9.5.2 are also valid in this configuration. The assumptions where

\[
\mathbb{E} [ b_{\text{nc}}(t) b_{n1}(t+\tau) ] = 0 \quad 9.90a
\]

\[
\mathbb{E} [ b_{\text{nc}}(t) b_{n2}(t+\tau) ] = 0 \quad 9.90b
\]

\[
\mathbb{E} [ b_{n1}(t) b_{n2}(t+\tau) ] \neq 0 \quad 9.90c
\]

Now eq. 9.89 can be written as follows

\[
F = \frac{|1 - \Gamma_L S_{99}|^2}{|S_{89}|^2} \left( \frac{|S_{99}|^2 |b_{\text{nc}}|^2}{|1 - \Gamma_L S_{99}|^2 |b_{\text{nc}}|^2} + \frac{|S_{89}|^2 |\Gamma_L|^2 |b_{n2}|^2}{|1 - \Gamma_L S_{99}|^2 |b_{nc}|^2} \right)
\]
\[
\left( \frac{|b_{n2}|^2}{|b_{nc}|^2} + \frac{S_{99} \Gamma_L}{1 - S_{99} \Gamma_L} \frac{b_{n1} b_{n2}^*}{|b_{nc}|^2} + \frac{S_{99} \Gamma_L^*}{(1 - S_{99} \Gamma_L)^*} \frac{b_{n1}^* b_{n2}}{|b_{nc}|^2} \right)
\]

9.91

Eq. 9.91 can be simplified

\[
F = 1 + |\Gamma_L|^2 \left( \frac{|b_{n2}|^2}{|b_{nc}|^2} + \frac{|1 - \Gamma_L S_{99}|^2}{|S_{99}|^2} \frac{|b_{n1}|^2}{|b_{nc}|^2} + \frac{\Gamma_L (1 - \Gamma_L S_{99})^*}{S_{99}} \frac{b_{n1} b_{n2}^*}{|b_{nc}|^2} + \frac{\Gamma_L^* (1 - \Gamma_L S_{99})}{S_{99}} \frac{b_{n1}^* b_{n2}}{|b_{nc}|^2} \right)
\]

4.19
9.7 The total noise figure

9.7.1 The relation between $|b_{c}|^2$ and $|b_{c}'|^{2}$

The signal circuit diagram is given in figure 9.12

![Signal Circuit Diagram](image)

**Fig 9.12 The signal circuit diagram (The arrow depicts the circulator rotation direction).**

For the emergent wave $b_2$ eq. 9.92 holds

$$b_2 = b_{c}' + \Gamma_b a_g$$  \hspace{1cm} 9.92

Eq. 9.32 has to be revised it becomes

$$a_g = \Gamma_L b_2$$  \hspace{1cm} 9.93

The substitution of eq. 9.93 into eq. 9.92 yields

$$b_2 = b_{c}' + \Gamma_b \Gamma_L b_2$$

$$b_{c}' = b_2 (1 - \Gamma_b \Gamma_L)$$  \hspace{1cm} 9.94

112
For the incident wave $a_1$ eq. 9.95 holds

$$a_4 = b_0 + \Gamma_a b_1$$  \hspace{1cm} 9.95

Eq. 9.60 has to be revised it becomes

$$b_1 = \Gamma_a a_4$$  \hspace{1cm} 9.96

The combination of eq. 9.95 and 9.96 yields

$$a_4 = b_0 + \Gamma_a \Gamma_a a_4$$

$$a_4 = \frac{b_0}{1 - \Gamma_a \Gamma_a}$$  \hspace{1cm} 9.97

The emergent wave $b_2$ is an addition of 2 waves

$$b_2 = b_9 + b_5$$  \hspace{1cm} 9.98

For $b_9$ the following equation holds

$$b_9 = S_{99} a_9$$  \hspace{1cm} 9.99

The combination of eq. 9.93 and eq. 9.99 yields

$$b_9 = S_{99} \Gamma b_2$$  \hspace{1cm} 9.100

For $b_5$ eq. 9.101 holds

$$b_5 = S_{54} a_4$$  \hspace{1cm} 9.101

Substituting eq. 9.97 into eq. 9.101 yields
\[ b_5 = \frac{S_{54} b_G}{1 - \Gamma_a \Gamma_G} \] 9.102

Now eq. 9.100 and 9.102 must be filled in into eq. 9.98

\[ b_2 = S_{99} \Gamma_L b_2 + \frac{S_{54} b_G}{1 - \Gamma_a \Gamma_G} \]

\[ b_2 = \frac{S_{54} b_G}{(1 - \Gamma_a \Gamma_G)(1 - \Gamma_L S_{99})} \] 9.103

Now eq. 9.103 must be substituted into eq. 9.94

\[ b_c^- = \frac{S_{54} b_G (1 - \Gamma_a \Gamma_G)}{(1 - \Gamma_a \Gamma_G)(1 - S_{99} \Gamma_L)} \] 9.104

This means for the relation between \( |b_c^-|^2 \) and \( |b_c|^2 \)

\[ \frac{|b_c|^2}{|b_c^-|^2} = \frac{|1 - \Gamma_a \Gamma_G|^2 |1 - S_{99} \Gamma_L|^2}{|S_{54}|^2 |1 - \Gamma_a \Gamma_L|^2} \] 9.105

The next step is the elimination of \( \Gamma_a \) and \( \Gamma_b \). The procedure is similar to the procedure used to eliminate \( \Gamma_2 \) and \( \Gamma_1 \), in paragraph 9.3.2.

\( \Gamma_a \) is defined as

\[ \Gamma_a = \frac{b_1}{a_4} \] 9.106

The emergent wave \( b_1 \) is an addition of two waves
\[ b_1 = b_4 + b_8 \quad 9.107 \]

For \( b_4 \) eq. 9.108 holds

\[ b_4 = S_{44} a_4 \quad 9.108 \]

For \( b_8 \) eq. 9.109 holds

\[ b_8 = S_{89} a_9 \quad 9.109 \]

Substituting eq. 9.108 and 9.109 into eq. 9.107 yields

\[ b_1 = S_{44} a_4 + S_{89} a_9 \quad 9.110 \]

The combination of eq. 9.98, 9.99 and 9.100 yields

\[ b_2 = S_{99} \Gamma_L b_2 + S_{54} a_4 \]

\[ b_2 = \frac{a_4 S_{54}}{1 - S_{99} \Gamma_L} \quad 9.111 \]

The substitution of eq. 9.93 and 9.111 into eq 9.110 gives

\[ b_1 = S_{44} a_4 + \frac{S_{89} \Gamma_L S_{54} a_4}{1 - S_{99} \Gamma_L} \]

\[ b_1 = \left\{ S_{44} + \frac{S_{89} \Gamma_L S_{54}}{1 - S_{99} \Gamma_L} \right\} a_4 \quad 9.112 \]

Substitution of eq. 9.112 into eq 9.106 yields
\[ \Gamma_a = S_{44} + \frac{S_{89} \Gamma L S_{54}}{1 - S_{99} \Gamma L} \]  
\[ \text{9.113} \]

\( \Gamma_b \) is defined as

\[ \Gamma_b = \frac{b_2}{a_9} \]  
\[ \text{9.114} \]

The combination of eq. 9.98, 9.99 and 9.101 gives

\[ b_2 = S_{99} a_9 + S_{54} a_4 \]  
\[ \text{9.115} \]

Eq. 9.116 holds for the incident wave \( a_4 \)

\[ a_4 = \Gamma_c b_1 \]  
\[ \text{9.116} \]

The substitution of eq. 9.116 into eq. 9.115 gives

\[ b_2 = S_{99} a_9 + S_{54} \Gamma_c b_1 \]  
\[ \text{9.117} \]

The combination of eq. 9.110 and 9.116 yields

\[ b_1 = S_{44} \Gamma_c b_1 + S_{99} a_9 \]

\[ b_1 = \frac{S_{89} a_9}{1 - S_{44} \Gamma_c} \]  
\[ \text{9.118} \]

Now eq. 9.118 must be substituted into eq. 9.117

\[ b_2 = S_{99} a_9 + \frac{S_{54} \Gamma_c S_{89} a_9}{1 - S_{44} \Gamma_c} \]
\[ b_1 = \left( S_{99} + \frac{S_{89} \Gamma_G S_{54}}{1 - S_{44} \Gamma_G} \right) a_0 \]  

9.119

To find \( \Gamma_b \), eq. 9.119 must be substituted into eq. 9.114

\[ \Gamma_b = S_{99} + \frac{S_{89} \Gamma_G S_{54}}{1 - S_{44} \Gamma_G} \]  

9.120

Substituting eq. 9.113 and 9.120 into eq. 9.105 gives

\[
\frac{|b_G|^2}{|b_G^\sim|^2} = \frac{1 - S_{44} \Gamma_G \frac{S_{89} \Gamma_G \Gamma L S_{54}}{1 - S_{99} \Gamma_L}}{|S_{54}|^2} \left| 1 - S_{99} \Gamma_L - \frac{S_{89} \Gamma_G \Gamma L S_{54}}{1 - S_{44} \Gamma_G} \right|^2
\]

\[
\frac{|b_G|^2}{|b_G^\sim|^2} = \frac{(1 - S_{44} \Gamma_G)(1 - S_{99} \Gamma_L) - S_{89} \Gamma_G \Gamma L S_{54}}{|S_{54}|^2 (1 - S_{44} \Gamma_G)(1 - S_{99} \Gamma_L) - S_{89} \Gamma_G \Gamma L S_{54}} \left| 1 - S_{44} \Gamma_G \right|^2
\]

\[
\frac{|b_G|^2}{|b_G^\sim|^2} = \frac{1 - S_{44} \Gamma_G}{|S_{54}|^2} = 5.3
\]

9.7.2 The relation between \( |b_{\text{nc}}|^2 \) and \( |b_{\text{nc}}^\sim|^2 \)

The noise circuit diagram is given in figure 9.13
The incident wave $a_1$ is given by eq. 9.121

$$a_1 = b_{nG} + \Gamma_c b_1 \quad 9.121$$

The emergent wave $b_1$ is an addition of two waves each of which is also an addition of two waves.

$$b_1 = b_3 + b_7$$

$$b_1 = b_4 + b_{n3} + b_8 + b_{n1} \quad 9.122$$

The emergent wave $b_4$ is

$$b_4 = S_{44} a_4 = S_{44} a_1$$

$$b_4 = S_{44} \Gamma_c b_1 \quad 9.123$$

The substitution of eq. 9.123 into eq. 9.122 yields
\[ b_1 = S_{44} \Gamma_G b_1 + b_{n1} + b_{n3} + b_8 \]

\[ b_1 = \frac{b_8 + b_{n1} + b_{n3}}{1 - S_{44} \Gamma_G} \tag{9.124} \]

The emergent wave \( b_2 \) is an addition of two waves each of which is also an addition of two waves.

\[ b_2 = b_6 + b_{10} \]

\[ b_2 = b_5 + b_{n4} + b_9 + b_{n2} \tag{9.125} \]

The emergent wave \( b_9 \) is

\[ b_9 = S_{99} a_9 = S_{99} a_2 \tag{9.126} \]

\[ b_9 = S_{99} \Gamma_L b_2 \tag{9.127} \]

Substitution of eq 9.127 into eq 9.125 gives

\[ b_2 = b_5 + b_{n4} + b_{n2} + S_{99} \Gamma_L b_2 \]

\[ b_2 = \frac{b_5 + b_{n4} + b_{n2}}{1 - S_{99} \Gamma_L} \tag{9.128} \]

For \( b_5 \) eq. 9.129 holds

\[ b_5 = S_{54} a_1 \tag{9.129} \]

The substitution of eq. 9.129 into eq. 9.128 yields

\[ b_2 = \frac{S_{54} a_1 + b_{n4} + b_{n2}}{1 - S_{99} \Gamma_L} \tag{9.130} \]

The emergent wave \( b_8 \) is

\[ b_8 = S_{89} a_2 \tag{9.131} \]
The combination of eq. 9.131 and 9.127 gives
\[ b_8 = S_{99} \Gamma_L b_2 \] 9.132

The substitution of eq. 9.130 and 9.132 into eq. 9.124 yields
\[ b_1 = \frac{S_{99} \Gamma_C (S_{54} a + b_{n4} + b_{n2}) + (b_{n1} + b_{n3})(1 - S_{99} \Gamma_L)}{(1 - S_{99} \Gamma_L)(1 - S_{44} \Gamma_C)} \] 9.133

Now eq. 9.133 must be substituted into eq. 9.121
\[ a_1 = \frac{\Gamma_C S_{99} (S_{54} a + b_{n4} + b_{n2}) + \Gamma_C (1 - S_{99} \Gamma_L)(b_{n1} + b_{n3})}{(1 - S_{99} \Gamma_L)(1 - S_{44} \Gamma_C)} \] 9.134

Now eq. 9.134 must be substituted into eq. 9.130
\[ b_2 = \frac{b_{n4} + b_{n2}}{1 - S_{99} \Gamma_L} \] 9.135

For \( b_2 \) eq. 9.136 holds
\[ b_2 = b_2 + a_2 \Gamma_b \] 9.136

For \( a_2 \) eq. 9.137 is valid
\[ a_2 = \Gamma_L b_2 \]

The combination of eq. 9.136 and 9.137 gives

\[ b_2 = b_{nG}^\sim + \Gamma_L \Gamma_b b_2 \]

\[ b_{nG}^\sim = b_2 (1 - \Gamma_b \Gamma_L) \]

Now eq. 9.120 must be substituted into eq. 9.138

\[ b_{nG}^\sim = b_2 (1 - \Gamma_L \left( S_{99} - \frac{S_{44} \Gamma_G}{1 - S_{44} \Gamma_G} \right)) \]

\[ b_{nG}^\sim = \frac{b_2}{1 - S_{44} \Gamma_G} \left( (1 - S_{99} \Gamma_L)(1 - S_{44} \Gamma_G) - \Gamma_L \Gamma_G S_{99} S_{54} \right) \]

The next step is the substitution of eq. 9.135 into eq. 9.139

\[ b_{nG}^\sim = \frac{S_{54}(b_n(1-S_{99} \Gamma_L)(1-S_{99} \Gamma_G)+\Gamma_L \Gamma_G (b_{n1} + b_{n2}) + \Gamma_L (1-S_{99} \Gamma_G)(b_{n1} + b_{n2})}{(1-S_{99} \Gamma_L)(1-S_{44} \Gamma_G)} + (b_{n1} + b_{n2})(1 - S_{99} \Gamma_L)(1 - S_{44} \Gamma_G) - \Gamma_L \Gamma_G S_{99} S_{54} \]

\[ (1-S_{99} \Gamma_L)(1-S_{44} \Gamma_G) \]

\[ b_{nG}^\sim = S_{54} b_{nG} + \frac{S_{54} \Gamma_G}{1 - S_{44} \Gamma_G} (b_{n1} + b_{n2}) + (b_{n4} + b_{n2}) \]

The relation between \(|b_{nG}^\sim|^2\) and \(|b_{nG}|^2\)

\[ \frac{|b_{nG}^\sim|^2}{|b_{nG}|^2} = \frac{S_{54} b_{nG} + \frac{S_{54} \Gamma_G}{1 - S_{44} \Gamma_G} (b_{n1} + b_{n2}) + (b_{n4} + b_{n2})}{|b_{nG}|^2} \]

121
9.8 The total noise figure

The substitution of eq. 5.3 and eq. 5.4 into eq. 4.16 yields

\[
F = \frac{|1-S_{44}^{\Gamma G}|^2}{|S_{64}|^2} \left| S_{54}^{b} n_G + \frac{S_{54}^{\Gamma G}}{1 - S_{44}^{\Gamma G}} (b_{n1} + b_{n3}) + (b_{n4} + b_{n2}) \right|^2
\]

9.141

Before eq. 9.141 can be simplified the following assumptions must be made.

In accordance with the assumptions made in paragraph 9.5.2 and 9.6 (eq.9.90) the following assumptions are made.

\[
E \left[ b_{nG}(t) b_{n1}(t+\tau) \right] = 0
\]

9.142a

\[
E \left[ b_{nG}(t) b_{n2}(t+\tau) \right] = 0
\]

9.142b

\[
E \left[ b_{nG}(t) b_{n3}(t+\tau) \right] = 0
\]

9.142c

\[
E \left[ b_{nG}(t) b_{n4}(t+\tau) \right] = 0
\]

9.142d

\[
E \left[ b_{n1}(t) b_{n2}(t+\tau) \right] \neq 0
\]

9.142e

\[
E \left[ b_{n1}(t) b_{n3}(t+\tau) \right] = 0
\]

9.142f

\[
E \left[ b_{n1}(t) b_{n4}(t+\tau) \right] = 0
\]

9.142g

\[
E \left[ b_{n2}(t) b_{n3}(t+\tau) \right] = 0
\]

9.142h

\[
E \left[ b_{n2}(t) b_{n4}(t+\tau) \right] = 0
\]

9.142i
\[ \mathbb{E} \left[ b_{n3}(t) b_{n4}^*(t+\tau) \right] = 0 \]  

Now eq. 9.141 can be simplified but eq. 9.142 must be taken into account

\[ F = \left| 1 - S_{44} \Gamma_G \right|^2 + \left| \Gamma_G \right|^2 \frac{|b_{n1}|^2}{|b_{nG}|^2} + \left| \Gamma_G \right|^2 \frac{|b_{n3}|^2}{|b_{nG}|^2} + \left| 1 - S_{44} \Gamma_G \right|^2 \frac{|b_{n4}|^2}{|S_{S4}|^2} \]

\[ \frac{|1 - S_{44} \Gamma_G|^2 |b_{n2}|^2}{|S_{S4}|^2} + \frac{(1 - S_{44} \Gamma_G)^* \Gamma_G}{|b_{nG}|^2} \frac{b_{n1}^* b_{n2}}{S_{S4}} + \frac{(1 - S_{44} \Gamma_G)^* \Gamma_G^*}{|b_{nG}|^2} \frac{b_{n1}^* b_{n2}}{|S_{S4}|^2} \]

\[ \frac{(1 - S_{44} \Gamma_G)^* \Gamma_G}{S_{S4}} \frac{b_{n3}^* b_{n4}}{|b_{nG}|^2} + \frac{(1 - S_{44} \Gamma_G)^* \Gamma_G^*}{S_{S4} \Gamma_G} \frac{b_{n3}^* b_{n4}}{|b_{nG}|^2} \]
9.8 The flow diagram of "nfcalc"

The following is a short description of the main features of the flow diagram:

First the number of measurements has to be specified. After this the input and output configuration of the noise figures must be specified. Then the three 2x2 S-matrices are read from file. After this $\Gamma_L$, $\Gamma_C$, and $S_{21}^0$ must be specified. Next the input configuration is set. Now the noise figures must be given. At this point the DBRT noise figure is calculated. Then the output configuration is set. And the program brings the results out. Now a check is made if all measurements are processed, if not the new S-matrices are being read. If all measurements are processed the program ends.
Begin

INPUT: number of measurements

INPUT: choice input configuration $f \lor k \lor p$

input correct ?

N

choice = $p$

Y

INPUT: first choice is $p$, next choice configuration $f \lor k$

input correct ?

N

Y

INPUT: choice output configuration $f \lor \text{scr}$

input correct ?

N

Y

INPUT: S parameters out of 3 files one, two and three

OUTPUT: error message fout3

OUTPUT: error message fout4

OUTPUT: error message fout5

B

A
### Table 10.1 The measurement using the HP NF meter (f=1.5 GHz)

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<th>I(μA)</th>
<th>( F_t ) (dB)</th>
<th>( G_t ) (dB)</th>
<th>( F_d ) (dB)</th>
<th>( G_d ) (dB)</th>
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<td>0.98 \times 10^3</td>
<td>7.26</td>
<td>1.16</td>
<td>6.66</td>
<td>2.44</td>
</tr>
<tr>
<td>766</td>
<td>0.95 \times 10^3</td>
<td>7.30</td>
<td>0.04</td>
<td>6.72</td>
<td>1.68</td>
</tr>
<tr>
<td>850</td>
<td>0.90 \times 10^3</td>
<td>7.74</td>
<td>-2.12</td>
<td>7.23</td>
<td>-0.84</td>
</tr>
<tr>
<td>900</td>
<td>0.95 \times 10^3</td>
<td>8.13</td>
<td>-2.76</td>
<td>7.64</td>
<td>-1.48</td>
</tr>
<tr>
<td>950</td>
<td>1.03 \times 10^3</td>
<td>8.74</td>
<td>-3.47</td>
<td>8.25</td>
<td>-2.19</td>
</tr>
<tr>
<td>1000</td>
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<td>9.63</td>
<td>-4.42</td>
<td>9.15</td>
<td>-3.14</td>
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<td>10.9</td>
<td>-5.62</td>
<td>10.4</td>
<td>-4.34</td>
</tr>
<tr>
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<td>12.3</td>
<td>-7.00</td>
<td>11.8</td>
<td>-5.72</td>
</tr>
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</table>

### Table 10.2 The measurement using the network analyzer (f=1.5 GHz)

<table>
<thead>
<tr>
<th>V(mV)</th>
<th>I(μA)</th>
<th>( F_t ) (dB)</th>
<th>( F_d ) (dB)</th>
<th>( G_d ) (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>1.67</td>
<td>0.89</td>
<td>-0.16</td>
</tr>
<tr>
<td>100</td>
<td>39</td>
<td>2.00</td>
<td>1.21</td>
<td>-0.30</td>
</tr>
<tr>
<td>250</td>
<td>200</td>
<td>3.98</td>
<td>3.15</td>
<td>-1.22</td>
</tr>
<tr>
<td>400</td>
<td>611</td>
<td>7.22</td>
<td>6.38</td>
<td>-2.86</td>
</tr>
<tr>
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<td>14.2</td>
<td>13.4</td>
<td>-9.37</td>
</tr>
<tr>
<td>600</td>
<td>2.16 \times 10^3</td>
<td>16.4</td>
<td>15.6</td>
<td>-10.3</td>
</tr>
<tr>
<td>620</td>
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<td>14.7</td>
<td>13.9</td>
<td>-6.91</td>
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<tr>
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<tr>
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<td>6.67</td>
<td>1.94</td>
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<td>-4.78</td>
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Table 10.3 The two-port S-parameter measurement (between input-port and DBRT-port)

<table>
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<th>FREQW. (GHz)</th>
<th>$S11$</th>
<th>$S12$</th>
<th>$S21$</th>
<th>$S22$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.9000</td>
<td>0.161</td>
<td>294.19</td>
<td>0.150</td>
<td>278.73</td>
</tr>
<tr>
<td>0.9500</td>
<td>0.093</td>
<td>280.48</td>
<td>0.093</td>
<td>218.48</td>
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<tr>
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<td>0.050</td>
<td>300.16</td>
<td>0.047</td>
<td>149.09</td>
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<tr>
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<td>0.059</td>
<td>324.15</td>
<td>0.025</td>
<td>46.78</td>
</tr>
<tr>
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<td>0.073</td>
<td>318.67</td>
<td>0.055</td>
<td>310.60</td>
</tr>
<tr>
<td>1.1500</td>
<td>0.077</td>
<td>305.06</td>
<td>0.047</td>
<td>248.29</td>
</tr>
<tr>
<td>1.2000</td>
<td>0.072</td>
<td>288.96</td>
<td>0.053</td>
<td>198.28</td>
</tr>
<tr>
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<td>0.065</td>
<td>240.29</td>
<td>0.054</td>
<td>154.87</td>
</tr>
<tr>
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<td>0.060</td>
<td>206.18</td>
<td>0.053</td>
<td>80.40</td>
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<tr>
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<td>0.070</td>
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<td>0.062</td>
<td>18.81</td>
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<td>341.98</td>
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<td>0.098</td>
<td>63.34</td>
<td>0.081</td>
<td>300.61</td>
</tr>
<tr>
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<td>0.106</td>
<td>37.13</td>
<td>0.089</td>
<td>257.90</td>
</tr>
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Table 10.4 The two-port S parameter measurement (between DBRT-port and output-port)

<table>
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<th>FREQW. (GHz)</th>
<th>$S11$</th>
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<th>$S21$</th>
<th>$S22$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.9000</td>
<td>0.282</td>
<td>272.63</td>
<td>0.251</td>
<td>30.14</td>
</tr>
<tr>
<td>0.9500</td>
<td>0.265</td>
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<td>0.180</td>
<td>329.45</td>
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<td>0.209</td>
<td>192.99</td>
<td>0.115</td>
<td>266.87</td>
</tr>
<tr>
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<td>193.63</td>
</tr>
<tr>
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<td>0.105</td>
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<td>89.91</td>
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<tr>
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<td>0.053</td>
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<td>0.070</td>
<td>292.18</td>
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<td>91.18</td>
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<td>237.57</td>
</tr>
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<td>0.120</td>
<td>61.01</td>
<td>0.081</td>
<td>188.39</td>
</tr>
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<td>143.50</td>
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<tr>
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<td>357.56</td>
<td>0.068</td>
<td>103.09</td>
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<td>334.03</td>
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<td>342.24</td>
<td>0.057</td>
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<td>321.71</td>
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<tr>
<td>1.7000</td>
<td>0.127</td>
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<td>258.21</td>
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129
<table>
<thead>
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<th>Frequency (GHz)</th>
<th>$S_{11}$ MOD. ARG.</th>
<th>$S_{12}$ MOD. ARG.</th>
<th>$S_{21}$ MOD. ARG.</th>
<th>$S_{22}$ MOD. ARG.</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.0000</td>
<td>0.205 + 322.30</td>
<td>0.221 + 235.74</td>
<td>0.886 + 3.02</td>
<td>0.233 + 325.91</td>
</tr>
<tr>
<td>0.0000</td>
<td>0.245 + 307.00</td>
<td>0.248 + 184.81</td>
<td>0.914 + 330.92</td>
<td>0.212 + 314.37</td>
</tr>
<tr>
<td>0.0000</td>
<td>0.193 + 283.92</td>
<td>0.173 + 133.44</td>
<td>0.941 + 299.99</td>
<td>0.156 + 293.96</td>
</tr>
<tr>
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<td>0.122 + 253.79</td>
<td>0.113 + 72.19</td>
<td>0.960 + 269.60</td>
<td>0.095 + 280.33</td>
</tr>
<tr>
<td>1.0000</td>
<td>0.061 + 258.63</td>
<td>0.081 + 252.64</td>
<td>0.968 + 239.91</td>
<td>0.054 + 295.91</td>
</tr>
<tr>
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<td>0.042 + 280.46</td>
<td>0.079 + 267.16</td>
<td>0.970 + 210.99</td>
<td>0.061 + 323.17</td>
</tr>
<tr>
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<td>0.096 + 209.29</td>
<td>0.969 + 182.43</td>
<td>0.072 + 316.22</td>
</tr>
<tr>
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<td>0.045 + 285.95</td>
<td>0.109 + 157.05</td>
<td>0.969 + 154.22</td>
<td>0.080 + 304.50</td>
</tr>
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<td>1.2000</td>
<td>0.047 + 277.64</td>
<td>0.105 + 107.43</td>
<td>0.970 + 126.47</td>
<td>0.076 + 288.56</td>
</tr>
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<td>0.084 + 65.01</td>
<td>0.969 + 98.53</td>
<td>0.076 + 267.54</td>
</tr>
<tr>
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<td>0.054 + 231.39</td>
<td>0.063 + 21.03</td>
<td>0.971 + 71.01</td>
<td>0.066 + 242.47</td>
</tr>
<tr>
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<td>0.057 + 155.05</td>
<td>0.023 + 321.95</td>
<td>0.971 + 43.19</td>
<td>0.063 + 288.46</td>
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<tr>
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<td>0.017 + 125.57</td>
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<td>0.970 + 348.08</td>
<td>0.070 + 129.47</td>
</tr>
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<td>0.968 + 320.38</td>
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<td>0.098 + 71.93</td>
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<tr>
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<td>0.125 + 314.74</td>
<td>0.962 + 265.45</td>
<td>0.101 + 37.92</td>
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Table 10.6 The reflection-coefficient measurement (noise source)

<table>
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<th>MOD.</th>
<th>ARG.</th>
<th>VSNF</th>
<th>REAL (%)</th>
<th>IMAG (%)</th>
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<tbody>
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<td>1000</td>
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<tr>
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<td>0.007</td>
<td>144.47</td>
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<td>9.882E-01</td>
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<td>136.16</td>
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<td>9.898E-01</td>
<td>9.646E-03</td>
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<td>9.905E-01</td>
<td>1.451E-02</td>
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<tr>
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<td>1.237E-02</td>
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<tr>
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<td>107.87</td>
<td>1.015E+00</td>
<td>9.955E-01</td>
<td>1.237E-02</td>
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<tr>
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<td>106.60</td>
<td>1.015E+00</td>
<td>1.001E+00</td>
<td>1.467E-02</td>
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<td>IM(Z')</td>
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<td>------</td>
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<td>--------</td>
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Table 10.8 The two-port S parameter measurement (between input-port and output-port with the DBRT biased at 0 mV)

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Table 10.9 The two-port S parameter measurement (between input-port and output-port with the DBRT biased at 100 mV)

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Table 10.10 The two-port S parameter measurement (between input-port and output-port with the DBRT biased at 250mV)

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Table 10.11 The two-port S parameter measurement (between input-port and output-port with the DBRT biased at 400mV)

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<tr>
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<td>0.926 4.81</td>
<td>0.081 304.28</td>
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Table 10.12 The two-port S parameter measurement (between input-port and output-port with the DBRT biased at 550mV)

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Table 10.13 The two-port S Parameter measurement (between input-port and output-port with the DBRT biased at 600 mV)

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<th>MOD. ARG.</th>
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135
### Table 10.14 The two-port S parameter measurement (between input-port and output-port with the DBRT biased at 620 mV)

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### Table 10.15 The two-port S parameter measurement (between input-port and output-port with the DBRT biased at 630 mV)

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Table 10.16 The two-port $S$ parameter measurement (between input-port and output-port with the DBRT biased at 766 mV)

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Table 10.17 The two-port $S$ parameter measurement (between input-port and output-port with the DBRT biased at 850 mV)

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Table 10.18 The two-port S parameter measurement (between input-port and output-port with the DBRT biased at 950 mV)

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Table 10.19 The two-port S parameter measurement (between input-port and output-port with the DBRT biased at 1100 mV)

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### Table 10.22 The second frequency measurement (f = 1.0 GHz)

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<th>$F_t$ (dB)</th>
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<th>$F_d$ (dB)</th>
<th>$G_d$ (dB)</th>
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### Continuation of Table 10.22 ($f = 1.0 \text{ GHz}$)

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11 Literature


[34] M. Sucher and J. Fox, "Handbook of microwave measurements", Polytechnic institute of Brooklyn, New York, 1963


