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FEL

T.D. 90/726

TNO-report

report no. FEL-90-B176

copy no. 8

title

A design method for an X-band diplexer

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classification	
title	: unclassified
abstract	: unclassified
report	: unclassified
appendices 1/2	: unclassified
no. of copies	: 22
no. of pages	: 48 (incl. titlepage & appendices excl. RDP & distributionlist)
appendices	: 2
date	: June 1990

TNO

report no.	: FEL-90-B176
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institute	: TNO Physics and Electronics Laboratory
date	: June 1990
NDRO no.	:
no. in pow '90	: 712.3

ABSTRACT (UNCLASSIFIED)

This report provides a detailed overview of a sequential method for the design of a passive X-band diplexer in order to support future design activities in a similar field. Also design steps and difficulties for diplexer design are dealt with in case only the computer program TOUCHSTONE® [4] is used. Predicted and measured performance of a realized diplexer are given.
An overview of FORTRAN 77 programs, developed at FEL-TNO is included. This software also can be used for separate synthesis of low- and highpass microstrip filters.

Keywords: Netherlands, (KR)

rapport no. : FEL-90-B176
titel : Een ontwerpmethode voor een X-band diplexer

auteur(s) : Ir. R. Overduin, Ir. J.C. Henkus, Ir. P.J. Koomen
instituut : Fysisch en Elektronisch Laboratorium TNO

datum : juni 1990
hdo-opdr.no. :
no. in iwp '90 : 712.3

SAMENVATTING (ONGERUBRICEERD)

Dit rapport geeft een gedetailleerd overzicht van een sequentiële methode voor het ontwerp van een passieve X-band diplexer teneinde ondersteuning te bieden voor soortgelijke toekomstige ontwerp-activiteiten.
Eveneens worden de ontwerpstappen en problemen besproken voor het geval men alleen het computerprogramma TOUCHSTONE® [4] gebruikt. Het voorspelde en gemeten gedrag van een gerealiseerde diplexer zijn in dit rapport opgenomen.
Een overzicht wordt gegeven van FORTRAN 77 programma's die bij het FEL-TNO zijn ontwikkeld. Deze programmatuur kan ook worden gebruikt voor afzonderlijk ontwerp van laag- en hoogdoorlatende microstrip filters.

1 INTRODUCTION

This report deals with the design of an X-band microstrip diplexer. Although the most important (numerical) results are given in this report, its main goal is to provide a complete overview of the design approaches and backgrounds, since the methods applied are rather generic. This implies a potential usefulness for similar activities in the future. For this purpose design steps which have appeared not to be successful for this particular diplexer design also are dealt with.
One approach is referred to as the sequential method. It is based on convenient microstrip topologies and implies separate design in two optimization steps. The initial optimization is preceded by a systematic global search and simple microstrip element models are used. The final optimization is local and uses the program TOUCHSTONE with non-ideal element models. There is a limited freedom in the choice of the topology, which means that a successful design is not guarantied for all design objectives.
Hence, an alternative straightforward approach, using the TOUCHSTONE program only, also is presented to be used in case the topology limitations of the sequential approach prevent a solution which is physically realizable. However, serious disadvantages are involved using this "TOUCHSTONE approach". Hence, use of the sequential method is highly recommended and the TOUCHSTONE approach is considered as a method to force a realizable microstrip diplexer if the sequential method fails.

This report focuses on the sequential design method.
In chapter 2, after a theoretical analysis, the fundamental design problem is explained, using a simple diplexer notation. Goals and conditions are given in chapter 3. The sequential approach is briefly discussed in chapter 4 and worked out in detail in chapters 5 and 6. Chapter 7 explains which design steps are involved if only the TOUCHSTONE program is used. Essential differences with the sequential

approach are given and the preference for the sequential method is motivated. Chapter 8 gives the most important intermediate and final numerical results. Finally, in chapter 9 the conclusions and recommendations are given.
Appendix 2 yields an overview of FORTRAN programs which are developed at FEL-TNO. This software can be used also for more or less separate synthesis of low- and highpass microstrip filters.

The diplexer design project has been a collaboration between two divisions of FEL-TNO, division 3 (radar and communication) and division 5 (technical development).

2 PROBLEM ANALYSIS

The diplexer design problem is analyzed in this chapter in order to explain the most fundamental design considerations for both the sequential and the TOUCHSTONE approach.

A diplexer is modelled in figure 2.1 as a passive, lossless three-port with one input and two output ports [3]. This device consists of a T-junction and a lowpass/ highpass complementary filter pair.

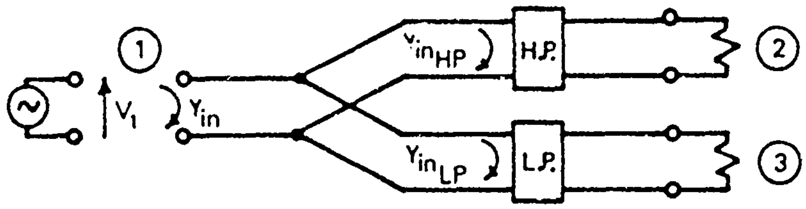

Fig. 2.1: Diplexer notation

The aim of the diplexer is to transmit the power, delivered to the input port (1) with a minimum insertion loss to one of the output ports (2) and (3), depending on the frequency. Since it is desired that a maximum of the input power should pass to the output port, the input admittance must be matched. Let the complex input admittances of the lowpass filter and the highpass filter be denoted by Y_{lp} and Y_{hp}, respectively, and the complex source admittance by Y_{in}, then we can write (in normalized form)

$$Y_{in} = Y_{lp} + Y_{hp} = 1 \qquad (2.1)$$

Equation (2.1) is the condition for the filters to be complementary and hence indicates the dependence of the two filters. Separating real and imaginary parts in (2.1) yields

$$\mathrm{Re}(Y_{lp}) + \mathrm{Re}(Y_{hp}) = 1 \qquad (2.2)$$

$$\mathrm{Im}(Y_{lp}) + \mathrm{Im}(Y_{hp}) = 0 \qquad (2.3)$$

Without prove it is stated that if in the usual case the filters are both of the minimum susceptive type, meaning that the imaginary part and the real part of the admittance are related through the Hilbert transform, equations (2.1)..(2.3) will hold.
The properties of an individual filter are examined to interpret equations (2.2) and (2.3).
For an arbitrary filter, let the complex input admittance be Y_f, the complex input voltage V_1, the input power and output power at unit termination P_{in} and P_{out}, respectively. We can write

$$P_{in} = |V_1|^2 \, \mathrm{Re}(Y_f) = P_{out} \qquad (2.4)$$

The filter is terminated with unit impedance, which means that

$$P_{out} = |V_2|^2 \qquad (2.5)$$

Substitution of (2.4) in (2.5) finally yields

$$|V_2|^2/|V_1|^2 = \mathrm{Re}(Y_f) \qquad (2.6)$$

Equation (2.6) implies that a specified power response, either low- or highpass, determines the form of the real part of the filter admittance.

Summarized, design of a diplexer is equivalent to satisfy all of the following relations:

- equation (2.2);
- equation (2.3);
- the imaginary and real parts are related through the Hilbert transform.

Furthermore, it is desired that the lowpass and the highpass filter have initial determined, realizable characteristics. These determine the real parts of the corresponding input admitances as function of the frequency (2.6).

3 DESIGN GOALS AND CONDITIONS

Specifications (figure 3.1):

-3 dB (cutoff) frequency	f_c = 7.5 GHz
Pass band highpass filter	= 7.5 - 11.5 GHz
Pass band ripple	< 0.5 dB
Att. of lowpass filter at 8.5 GHz	< -15 dB
Input port reflection coefficient	$\|\Gamma_{in}\|$ < -10 dB, or
SWR at the input port	< 6 dB

Constraints:

Source impedance =	
Load impedance (both filters)	Z_0 = 50 Ω
Relative electr. permittivity	ϵ_r = 2.33 (Duroid)
Substrate height	h = 508 μm
Strip height	t = 10 μm
Loss tangent	TAN(δ) = 0.001

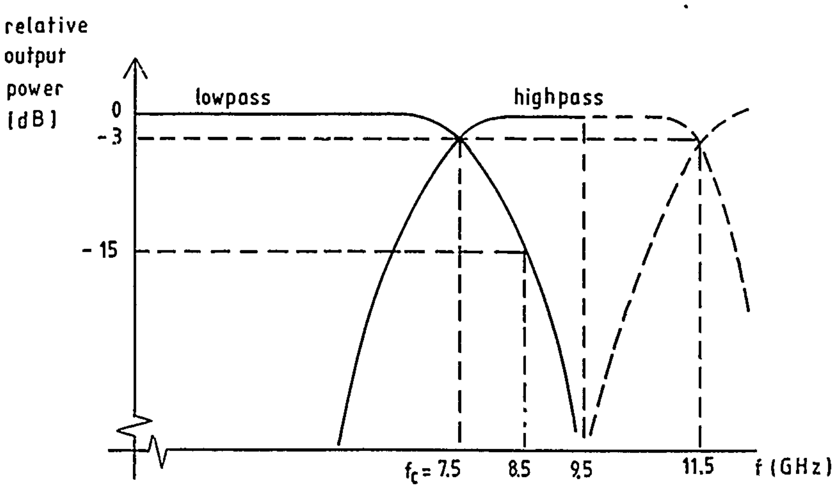

Fig. 3.1: Diplexer specifications

The microstrip realization of the filters implies that the power transfer functions subsequently mirror around a multiple of the resonance frequency f_0=9.5 GHz, which corresponds with four times the electrical length of the microstrip lines

4 THE SEQUENTIAL APPROACH

This chapter gives the design objective in mathematical terms and briefly gives the individual steps for the sequential approach.

An attractive approach would be to separate the design of the diplexer into two independent steps:

- synthesis of the lowpass filter;
- synthesis of the highpass filter.

This is possible if prototype filter responses are assumed for both filters. The diplexer design specifications are conveniently met using Butterworth prototype power transfer functions (maximally flat transfer in the pass band; moderate slope performance):

Lowpass filter:

$$|V_{21p}|^2/|V_{11p}|^2 = \mathrm{Re}(Y_{1p}) = 1/(\ 1 + \Omega^{2N}\ ) \tag{4.1}$$

Highpass filter:

$$|V_{2hp}|^2/|V_{1hp}|^2 = \mathrm{Re}(Y_{hp}) = \Omega^{2N}/(\ 1 + \Omega^{2N}\ ) \tag{4.2}$$

In which N is the order of the circuit and -for discrete element circuits- $\Omega = f/f_c$,
with f the signal frequency and f_c the cutoff frequency.
It is easily verified that (4.1) and (4.2) satisfy all relations, given in chapter 2.

To describe the "mirror" effect of the power transfer function in case of a microstrip realization, in chapter 5.2 to Ω an alternative frequency variable will be assigned.

The following successive design steps are executed:

a) Determine the microstrip topologies for both filters
b) Initial optimization: for each filter separately element values are calculated in such a way that the prototype responses are matched as accurate as possible. For each filter the optimization is preceded by a systematic global search for suitable initial values.
c) Element extraction: calculation of the dimensions of the microstrip lines from the resulting wave impedance values
d) Second optimization: the dimensions, mentioned above are used as input (initial) values for the computer program TOUCHSTONE, which utilizes non-ideal element models in order to simulate dispersion, junction effects, substrate losses, a finite small stripheight, etc.. The optimization objective is now to optimize the reflection coefficient at the source, and vary the microstrip line dimensions rather than the element values. This is a local optimization.

Items a), b) and c) are applicable to the low- and the highpass filter. Therefore, these are separately dealt with in chapter 5 for the highpass filter and chapter 6 for the lowpass filter. The extraction procedure, mentioned under item c) is considered as a minor problem and was solved, using software available at the establishment [4],[5]. Theoretical backgrounds on this subject are considered not to be within the scope of this report.

5 HIGHPASS FILTER DESIGN

In this chapter a topology for the highpass section is proposed for which a model is propounded which is applied using the sequential method. Further, the initial optimization of the network is dealt with. In the last two paragraphs particular difficulties for two realizations are discussed. Also the experiences with these two topologies, applied to the specific design problem (chapter 3), are included in these paragraphs.

5.1 Topology determination and modelling

The highpass filter is realized in parallel coupled line structures. In order to use this structure as a two-port filter component, two diagonal connections are left open. The advantage of this realization is that inconvenient perforation and bonding through the substrate is avoided. A disadvantage however, is that the non-ideal coupling of the lines is not included in the model which is used for the initial optimization.
The number of coupled sections is determined as follows: a Butterworth lowpass power response of the fourth order yields an attenuation of -25 dB at 8.5 GHz. Since the order of the complementary highpass filter should equal the lowpass filter order, three coupled sections will probably be sufficient to achieve the desired attenuation of -15 dB at 8.5 GHz (chapter 3).

The parallel coupled line section is modelled as proposed by Ozaki and Ishii [1]. The model consists of two capacitors and one lossless transmission line, which is denoted in the equivalent circuit diagram of figure 5.1.1 as a unit element (u.e.).

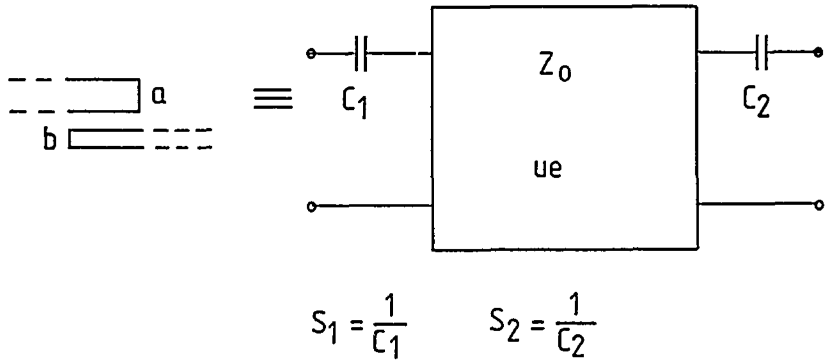

Fig. 5.1.1: Equivalent circuit diagram of a coupled section

Let the individual lines be labelled a and b, then the even and odd mode wave impedances are denoted by Z^i_{0e} and Z^i_{0o}, respectively (i=a,b). S_1 and S_2 are the reciprocal capacitor values in the equivalent circuit diagram. Z_0 is the wave impedance of the unit element. All the elements of the equivalent circuit diagram are s-plane elements, which means that the real frequency f is transformed to Richard's frequency variable

$$s = jTAN(\tfrac{1}{2}\pi f/f_0) \tag{5.1}$$

in which f_0 is the resonance frequency, corresponding with four times the electrical length of the microstrip lines and $j^2 = -1$. Introduction of s allows convenient linear analysis and synthesis for distributed circuits, comparable with "low-frequency" techniques under the restriction that all electrical line lengths are equal.

The elements of the equivalent circuit diagram are related to the even and odd impedances as follows:

$$Z^a_{0e} + Z^a_{0o} = 2Z_0 + 2S_1 \tag{5.2}$$

$$Z^{a}_{0o} + Z^{b}_{0o} = S_1 + S_2 \tag{5.3}$$

$$Z^{b}_{0e} + Z^{b}_{0o} = 2Z_0 + 2S_2 \tag{5.4}$$

To model the highpass filter, the equivalent circuits of the coupled sections are cascaded. In the equivalent circuit of the resulting filter two adjacent capacitors of the individual coupled sections can be replaced by an equivalent capacitor. For a total number of sections N_{sec}, there are N_{sec}-1 capacitor pairs to be substituted. The final equivalent circuit for the highpass filter is given in figure 5.1.2.

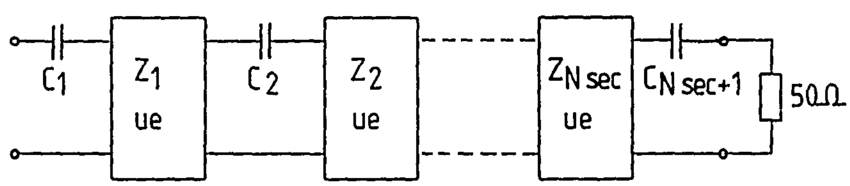

Fig. 5.1.2: Equivalent circuit diagram for an N_{sec} section highpass filter

It is possible to show that the equivalent circuit, shown in figure 5.1.2, can further be reduced to a topology with only N_{sec} unit elements and one capacitor. For the generic approach to be discussed in paragraph 5.2 and further, this latter representation is not suitable since element conversions are necessary, which do not contribute to a simpler design method.

5.2 General method for the initial optimization

Modelling the highpass filter with the equivalent circuit diagram as shown in figure 5.1.1 results in a redundant rational form of the real part Y_r of the complex admittance function Y_{hp}

$$\begin{aligned} Y_r(s^2) &= \mathrm{Re}\{Y_{hp}(s)\} = \mathrm{Re}\{Y_{hpn}(s)/Y_{hpd}(s)\} \\ &= [Y_{hpn}(s)*Y_{hpd}(-s)]/[Y_{hpd}(s)*Y_{hpd}(-s)] \end{aligned} \qquad (5.5)$$

The variable s is purely imaginary, which means that $s^* = -s$.

It can be shown that after cancellation of the common factor s^{Nsec-1} in numerator and denominator the following generic rational form for Y_r results:

$$Y_r(s^2) = \frac{-s^2(1 - s^2)^{Nsec}}{y_0 + y_1 s^2 + \ldots + y_{(Nsec+1)} s^{2(Nsec+1)}} \qquad (5.6)$$

The parameters y_i ($i = 0..N_{sec}+1$) all depend on the element values S_1, S_2 and Z_0 of the individual line sections. Since the number of independent elements in the non-redundant case is equal to $N_{sec}+1$, and the total number of denominator parameters is equal to $N_{sec}+2$, the parameters y_i are mutual dependent. The set of simultaneous, non-linear equations in y_i can not be found easily for an arbitrary value of N_{sec}. This means that the parameters y_i can not be chosen as independent inputs for a fitting procedure to meet the Butterworth response function (4.2) and extract the element values after the optimization process.
Alternatively, the elements of all sections can be optimized directly. To generate the values of y_i, it appears most convenient to evaluate the ABCD matrix of the highpass filter by multiplication of the ABCD matrices of the individual sections (appendix 1). In this calculation, it is not necessary to involve the common factor $1/\surd(1-s^2)$ in equations (al.3) and (al.4). This means that the operations are mainly restricted

to subtraction, addition and (scalar) multiplication, which do not lay a high claim on computer resources.

Advantages of this latter method are that

- after optimization the optimum element values are directly available;
- during optimization restrictions can easily be laid on the element values to obtain realizable strip dimensions (see paragraphs 5.3 and 5.4).

A disadvantage of this method is that the calculation of the parameters y_i from the element values increases computation time considerably since this analysis has to be carried out at each iteration.
The solution for the initial optimization was obtained using a least-squares method, applying a three section topology. The residue R_i for each frequency sample number i, denoting the criterion for an optimum solution, was determined as

$$R_i = (f_i - f_{obj,i})/f_{obj,i} \qquad (5.7)$$

in which $f_{obj,i}$ is the objective- and f_i is the actual power transfer function for each sample/observation. Hence, for the low- and the highpass case $f_{obj,i}$ is the ith sample of (4.1) and (4.2), respectively. The division by $f_{obj,i}$ in (5.7) accomplishes a weighting of the absolute error f_i-$f_{obj,i}$, in such way that for each sample the squared values of the residues are likely to be in the same order of magnitude and hence contribute about equally to the residual sum of squares R_s

$$R_s = \Sigma R_i^2 \text{ over all samples} \qquad (5.8)$$

The objective power transfer functions are generated for 20 samples, adapting the Butterworth prototype transfer function (4.2) for the microwave distributed circuit topologies, introducing

$$\Omega = \mathrm{TAN}(\tfrac{1}{2}\pi f/f_0) \tag{5.9}$$

$$\Omega_c = \mathrm{TAN}(\tfrac{1}{2}\pi f_c/f_0) \tag{5.10}$$

yielding the frequency variable

$$X = \Omega/\Omega_c \tag{5.11}$$

in the modified power transfer function

$$|V_{2hp}|^2/|V_{1hp}|^2 = X^{2N}/(\ 1 + X^{2N}\ ) \tag{5.12}$$

with $N = N_{sec}+1 = 4$.
It is easily verified that for $f \ll f_0$, and $f_c \ll f_0$ (5.12) becomes (4.2).
Numerical results are discussed in chapter 8.

5.3 Symmetrical coupled line filters

For symmetrical coupled lines, which means that lines a and b (figure 5.1.1) have an equal linewidth, there are commonly known techniques and computer programs available to calculate the even and odd wave impedance from the line width and the gap. Hence, it was attempted to first use this structure for the highpass filter realization.

Symmetrical coupled lines means that

$$S_1 = S_2 = S \tag{5.13}$$

$$Z^a_{0o} = Z^b_{0o} = Z_{0o} \tag{5.14}$$

$$Z^a_{0e} = Z^b_{0e} = Z_{0e} \tag{5.15}$$

which simplifies equations (5.2)..(5.4) to

$$Z_{0e} + Z_{0o} = 2Z_0 + 2S \tag{5.16}$$

$$Z_{0o} = S = 1/C \tag{5.17}$$

Referring to the equivalent circuit diagram (figure 5.1.1), condition (5.13) implies that the mutual capacitor values of the cascaded sections are no longer independent. The degree of design freedom has decreased with one, which again makes direct optimization of the denominator coefficients y_i impossible. Therefore, computation of these parameters

from the S- and Z_0 values is performed as previously discussed in paragraph 5.1.

During optimization the element values are constrained in order to obtain realizable line dimensions- and gapwidth. With the use of the given material and the current production aids dimensions down to approximately 80 μm are considered feasible. The maximum strip width is fixed at 3 mm. The minimum and maximum values of Z_{0o} and Z_{0e} corresponding to the dimension limits are calculated using the program LINECALC® [5], from which limits to Z_0 and S (or rather C) can be determined.
From equations (5.15) and (5.16), however, the maximum value of Z_0 depends on the current value of C. For a number of values of C, distributed in the area where the corresponding wave impedance is realizable (32 to 175 ohms), the minimum and maximum values of Z_0 have been determined.
A curve fitting was executed to approach the minimum and maximum values of Z_0 by two power series in C. The results are given in figures 5.3.1. and 5.3.2. These figures are based on normalized values of $Z_{0min/max}$, C and S, denoted by $Z_0'_{min/max}$, C' and S', respectively.

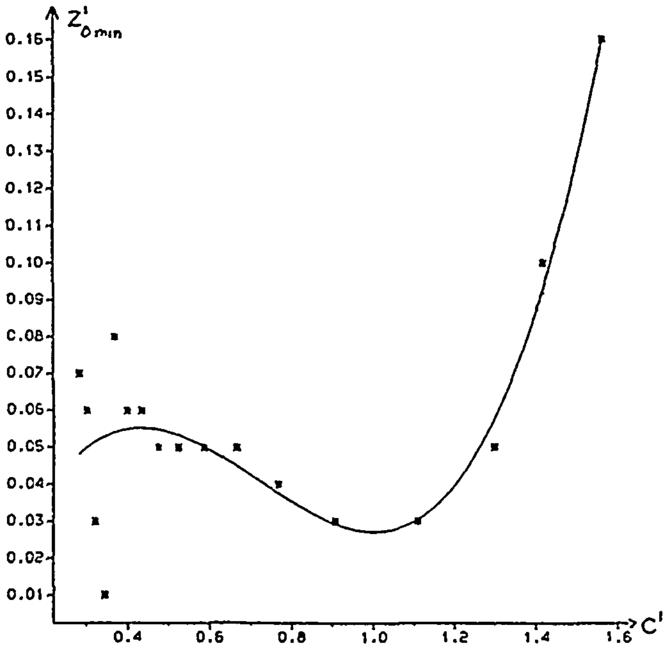

Fig. 5.3.1: Minimum values of Z_0' as a function of C

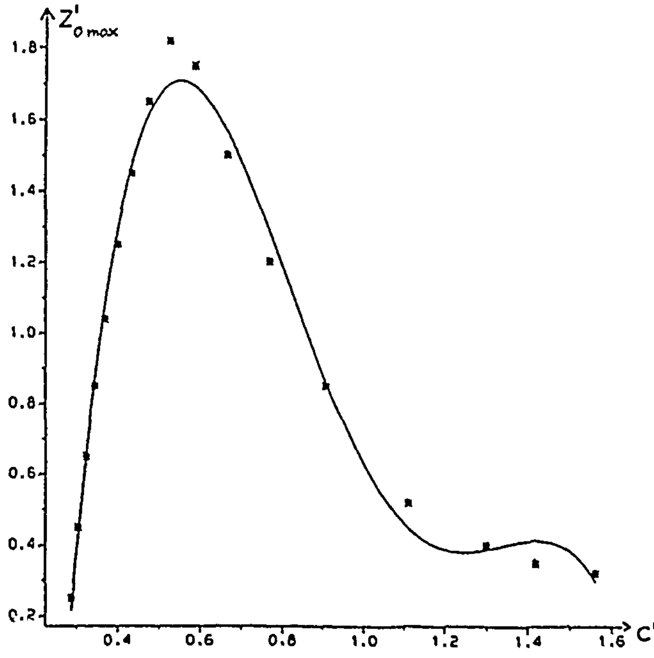

Fig. 5.3.2: Maximum values of Z_0' as a function of C'

Since all impedance values are normalized to 50 ohms, restriction of values of S between 32 and 175 ohms implies that values for S' are between 0.64 and 3.5. This means that values for C'=1/S' should be between 0.286 and 1.563, which is accomplished using the following function:

$$C' = \tfrac{1}{2}(C'_{max} + C'_{min} + (C'_{max} - C'_{min})SIN(X_i)) \qquad (5.18)$$
$$= 0.9245 + 0.6385 * SIN(X_i)$$

in which X_i is an element from the argument array which is directly used by the optimizer.
The "best fits" for the minimum impedance $Z_0'_{min}$ (figure 5.3.1) and the maximum impedance $Z_0'_{max}$ (figure 5.3.2) appear to be

$$Z_0'_{min} = -0.019 +0.399C' -0.656C'^2 +0.304C'^3 \qquad (5.19)$$

$$Z_0'_{max} = -7.578 +44.680C' -73.973C'^2 +48.867C'^3 -11.367C'^4 \qquad (5.20)$$

Finally, the current value of Z'_0 is determined analogous to equation (5.18):

$$Z'_0 = \tfrac{1}{2}(Z'_{0max} + Z'_{0min} + (Z'_{0max} - Z'_{0min})SIN(X_{i+1})) \qquad (5.21)$$

This completes the formal discussion of the approach for the highpass topology with symmetrical coupled lines. Although in essence this method is generic, the outcomes of the initial optimization as mentioned in chapter 4, sub b) for the symmetrical coupled line topology were not accepted for the specific design case. The design goals, mentioned in chapter 3, only were sufficiently met with dimensions, which are not realizable on the Duroid substrate.

Since this optimization was combined with a systematical global optimum search, and several initial promising values were tried without satisfaction, it is considered impossible to construct the desired highpass filter using symmetrical coupled microstrip lines.
The FORTRAN program for the synthesis of symmetrical coupled lines is mentioned in appendix 2.
The next paragraph describes design considerations and the final numerical results for a filter realization using asymmetrical coupled structures.

5.4 Asymmetrical coupled line filters

As previously remarked in paragraph 5.1, the total number of independent equivalent circuit parameters is equal to $N_{sec}+1$, which means that the degree of freedom compared with the symmetrical coupled line topology has increased with one.

For two reasons it is very difficult to limit each Z_0, C_1 and C_2 value in a way as discussed under the design approach for symmetrical coupled lines, where a relation between the limits and a relation between limits and a current value was derived:

- first, there are three elements to be optimized for each section instead of two. To find the relations mentioned above is a rather formidable task;
- second, no suitable software is available on element extraction to perform this job within a reasonable time span.

However, limitation of the values of Z_0, S_1 and S_2 on two sides has not appeared necessary. It was sufficient to obtain positive element values, so the x^2 function was used.
After the initial optimization, with the aid of the computer program TOUCHSTONE the line dimensions and the gap width can be derived indirectly from the output values of Z_0, S_1 and S_2, without explicit use of the relations for the even- and the odd mode impedances (5.2)..(5.4).
This element extraction procedure is executed as follows:

- calculate the hybrid parameters from the ABCD matrix of the asymmetrical coupled line section (appendix 1) for a few frequencies;
- use the $Im[h_{11}]$, $Re[h_{12}]$ and $Im[h_{22}]$ values as optimization objectives, since $Re[h_{11}] = Im[h_{12}] = Re[h_{22}] = 0$ and $h_{12} = -h_{21}$ for the parallel coupled section
- declare as variables W_1, W_2, the gap s and the linelength l.

The name and a brief description of the program for the design of asymmetrical coupled lines is included in appendix 2.

6 LOWPASS FILTER DESIGN

In this chapter the lowpass microstrip topology is given and two possible design approaches are discussed.

6.1 Topology determination

The most general lowpass microstrip filter topology consists of series transmission lines and lines, connected as open shunt stubs. The transmission lines are assumed lossless and are modelled in figure 6.1.1 as unit elements. The capacitors in this figure are s-plane elements, which model the open stubs. It is also possible to expand this lowpass realization with coupled lines as long as there is a DC-path from the input- to the output connection. However, the elementary topology is believed to offer sufficient design freedom to meet the objects (chapter 3).

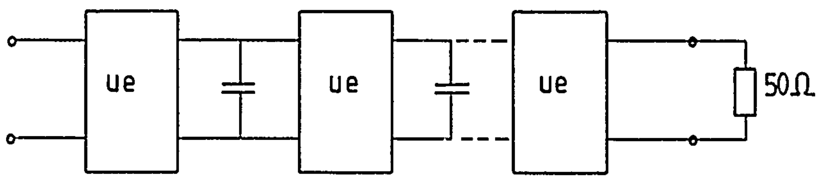

Fig. 6.1.1: General lowpass filter s-plane topology

Like the highpass filter, connections through the substrate to the earth plane are avoided if this topology is applied.

6.2 Redundant synthesis

A Butterworth prototype lowpass circuit, realized in microstrip as depicted in figure 6.1.1, can be designed using the traditional L-C ladder network as a starting point. For this topology the (Richard's) frequency and the impedance normalized component values which yield a Butterworth power transfer function, are explicitly given in the literature as a function of the network order (the number of L-C elements). The actual design problem is now restricted to the transformation of the L-C ladder to the distributed structure, modelled in figure 6.1.1. The problem is tackled using a Kuroda identity and its inverse identity (figure 6.2.1).

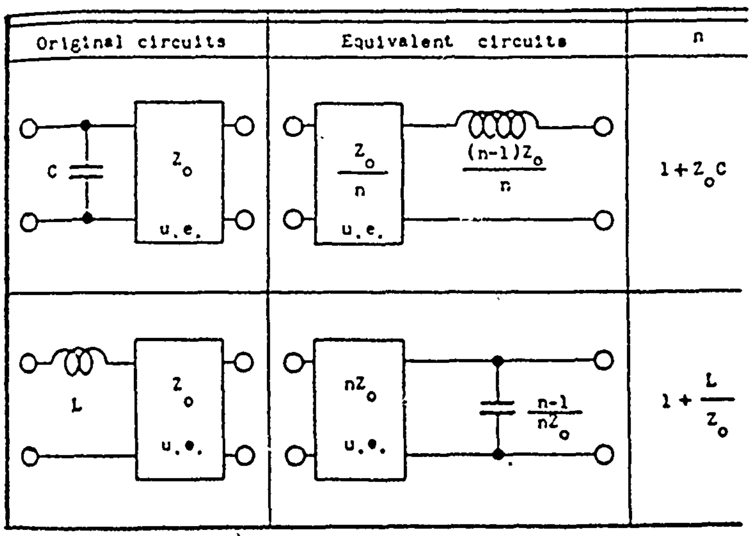

Fig. 6.2.1: Two Kuroda identities

The procedure is to subsequently add unit elements to the load connection of the ladder network and transfer each unit element towards the source, until there are no longer self-inductions present. The impedance value of each unit element added should equal unity, which

means that the power transfer function and the input admittance do not change.
In case one or more impedance values appear not to be convenient, more unit elements can be added in the way described above.
To eliminate all self-inductions in the fourth order network four unit elements have to be added. These components are redundant because they only transfer the topology and do not increase the order of the circuit. This approach is therefore called redundant synthesis. An advantage of this method is, that without much effort the final microwave circuit yields the exact Butterworth power transfer function.
Application of this approach to the design of the lowpass filter section for the diplexer has led to several difficulties:

- a total of eight transmission lines could hardly be realized on the restricted substrate surface
- the width of one or two lines was to small to be realized. A sensitivity analysis with the program TOUCHSTONE has shown that these lines could not be omitted without impact on the power transfer function.

The alternative approach, which yielded the input values for the final optimization with TOUCHSTONE, is briefly discussed in the next paragraph.

6.3 Minimum redundant synthesis

An alternative approach for the lowpass filter design is to optimize the susceptibility function analogous to the straightforward method discussed in paragraph 5.2. This means that the susceptibility function is calculated, evaluating the total ABCD matrix of the lowpass filter. For the lowpass section this function is comparable with (5.12) and has the general form

$$Y_r(s^2) = \frac{(1 - s^2)^{N_{ue}}}{1 + y_1 s^2 + \ldots + y_N s^{2N}} \qquad (6.1)$$

This form is to numerically meet the Butterworth lowpass power transfer function

$$|V_{21p}|^2 \,/\, |V_{11p}|^2 = 1 \,/\, (\, 1 + X^{2N} \,) \qquad (6.2)$$

in which, like in 5.2,

$$X = \Omega/\Omega_c, \text{ with } \Omega = \mathrm{TAN}(\tfrac{1}{2}\pi f/f_0) \text{ and } \Omega_c = \mathrm{TAN}(\tfrac{1}{2}\pi f_c/f_0)$$

Equation (6.2) is comparable with (5.20).
The parameters y_i (i= 1..N) all depend on the impedance values and are obtained for each iteration through straightforward analysis. N_{ue} is equal to the number of series transmission lines and N is equal to the total number of elements (series lines and stubs). Obviously, the form of the numerator functions in (6.1) and (6.2) differ considerably, which indicates the probable necessity for a redundant topology to numerically cancel the numerator in (6.1) as much as possible.

The design approach, described in paragraph 5.2 is adapted slightly for the lowpass case on two points:

- The degree of design freedom is higher than with the highpass filter synthesis, because there is a limited freedom in the choice of the number of unit elements, the placement- and number of stubs. The initial optimization can therefore be divided into more attempts in order to obtain the best realizable configuration with the least number of lines.
- For sample frequencies near the resonance frequency f_0, the value of $f_{obj,i}$ (the desired power transfer function value) approaches zero rapidly. This means that if the residue were defined as in (5.7)

$$R_i = (f_i - f_{obj,i})/f_{obj} \qquad (6.3)$$

for these samples the residues would be very large and their contribution to the sum of residues would be disproportionate. Hence, the residue is for the lowpass case is determined as

$$R_i = f_i - f_{obs,i} \qquad (6.4)$$

A good trade-off between the number of microstrip lines and the approximation of the Butterworth function was achieved with three unit elements and two open stubs (see paragraph 5.1).
The corresponding FORTRAN program is mentioned in appendix 2.

7 DIPLEXER DESIGN USING TOUCHSTONE ONLY

It is believed that the topology restrictions of the sequential diplexer design method in most cases is not a limitation which prevents realizable results. Especially in the special design case, this method yielded satisfactory diplexer dimensions. However, if the topology restrictions of the sequential method form a limitation, one could try an alternative approach which is based on the use of only the program TOUCHSTONE. This chapter discusses the "TOUCHSTONE approach". Important differences with the sequential method and difficulties one can expect using the alternative approach are highlighted.

The fundamentals for the sequential approach also apply to the TOUCHSTONE method. This means that for the TOUCHSTONE approach diplexer design also is separated in a low- and a highpass filter synthesis. An important difference of this design approach with the sequential method, is the freedom of the topology choice. The microstrip structures for the low- as well as for the highpass filter may result from the redundant synthesis procedure, using discrete components and Kuroda identities (6.2). If the discrete filters are designed to match complementary Butterworth responses, the characteristics of the microstrip filters will be fully equivalent to the discrete prototypes.

The design procedure consists of the following steps:

a) Calculation of the ideal lossless single terminated low- and highpass microstrip filters with the redundant synthesis approach using Kuroda identities. These serve as the initial circuits for the optimization.

b) Determination of the limits for each microstrip dimension to be optimized

c) Local optimization of the initial filters with the non-ideal models of TOUCHSTONE. If the topology does not yield satisfactory results or -as often is the case for higher-order filters- the redundancy is considerable, it is adjusted and the optimization is repeated.

In some cases -with a suitable initial guess provided- this TOUCHSTONE approach may give good results, and avoids explicit element extraction. There are also, however, considerable disadvantages involved:

- The optimization with TOUCHSTONE in most cases is an extremely (computer)time-consuming procedure.
- The starting point usually is a highly redundant circuit because the redundancy is proportional to the order of the filter. Moreover, reduction of the redundancy or modification of the topology often is a tedious process of trial-and-error in case only TOUCHSTONE is used.
- The random-optimization feature of the program can not be applied properly for the diplexer design problem. This means that only a gradient-oriented local optimization can be selected, which implies the hazard of obtaining a quasi-optimum solution or convergence may not be obtained at all.

The difficulties of the TOUCHSTONE approach stated above, clearly show that a design tool like TOUCHSTONE with complex models is not ideal to use for diplexer design without any additional software. The TOUCHSTONE approach therefore is considered only as a means to force a microstrip diplexer realization.

8 NUMERICAL RESULTS

This chapter compares the element values of the initial optimization for both the low- and the highpass filter with the final values after the final optimization. In the last paragraph, the predicted and the measured performance of the total diplexer are given.

8.1 Highpass filter

In table 8.1 the results for the highpass filter are given. From the initial optimization, two different optimum solutions have resulted, yielding exactly the same residues. The values in the second column were more convenient and have therefore been chosen to derive the strip dimensions, given in the fourth column. These values are the actual input for the final optimization to be carried out with TOUCHSTONE. The final dimensions are given in the fifth column. The values of the corresponding equivalent circuit elements are found in column three. The section numbers increase in the load direction.

	circuit values [Ω]				dimensions [μm]		
Sec.		initial opt.		final		init. opt.	final
1	S_1	166.5	179.4	150.1	W_1	50.8	80.0
	S_2	145.1	90.9	107.3	W_2	255.1	169.3
	Z_0	69.6	56.6	50.4	S	187.0	208.5
					l	5527	5720
2	S_1	173.2	114.3	122.5	W_1	82.1	102.1
	S_2	182.3	48.1	65.4	W_2	328.6	319.0
	Z_0	110.7	68.6	61.6	S	64.4	68.5
					l	5820	6194
3	S_1	142.7	120.8	134.2	W_1	87.3	85.5
	S_2	44.5	64.5	50.5	W_2	271.1	469.8
	Z_0	84.2	64.2	60.7	S	87.2	69.2
					l	5661	5826

Table 8.1: Highpass filter results

As the value of W_1 for the first section from the initial optimization is considered too small, at the final optimization W_1 was set to the constant value of 80 μm, which is the minimum realizable value on the duroid substrate.
The residue values, resulting from the initial optimization (second column of table 8.1) are listed in table 8.2 for each sample frequency f_s. These values are determined according to equation (5.7), which means that these are errors, relative to the objective values. The residues for the optimization with TOUCHSTONE are not given in this table, because for this final optimization a different criterion was to be met (optimum SWR at the diplexer-input port rather than Butterworth approximation) and hence a comparison with the residues from the initial optimization is not sensible.

sample	f_s [GHz]	residue
1	4.75	0.41
2	5.00	0.17
3	5.25	0.04
4	5.50	-0.11
5	5.75	-0.19
6	6.00	-0.23
7	6.25	-0.24
8	6.50	-0.20
9	6.75	-0.10
10	7.00	0.05
11	7.25	0.16
12	7.50	0.03
13	7.75	-0.14
14	8.00	-0.16
15	8.25	-0.04
16	8.50	0.11
17	8.75	0.13
18	9.00	0.03
19	9.25	-0.07
20	9.50	0.12

Table 8.2: Residues from the highpass filter initial optimization

8.2 Lowpass filter

Table 8.3 gives the impedance values and the line dimensions from the initial and the final optimization, respectively.
The section numbers increase in the load direction.

	impedance [Ω]			dimensions [μm]		
Sec.	initial opt.		final	init. opt.		final
1	ue	131.8	66.1	W	202.7	951.8
				l	5868	5190
2	stub	79.0	87.7	W	687.5	556.7
				l	5727	5656
3	ue	66.2	70.9	W	949.6	841.1
				l	5674	5763
4	stub	168.0	111.7	W	83.3	317.4
				l	5922	5741
5	ue	56.4	57.8	W	1238.8	1194.0
				l	5627	5608

Table 8.3: Lowpass filter results

In table 8.4 the residues for the initial optimization are given for the lowpass case. These values correspond to absolute errors according to (6.4), and are not related to the objective values for the reason, mentioned earlier in paragraph 6.3.

sample	f_s [GHz]	residue
1	3.75	-0.016
2	4.05	0.005
3	4.36	0.014
4	4.66	0.011
5	4.96	0.002
6	5.26	-0.009
7	5.57	-0.016
8	5.87	-0.014
9	6.17	-0.003
10	6.47	0.014
11	6.78	0.021
12	7.08	-0.004
13	7.38	-0.031
14	7.68	0.010
15	8.00	0.044
16	8.30	0.029
17	8.60	0.010
18	8.90	0.002
19	9.20	0.000
20	9.50	0.000

Table 8.4: residues from the lowpass filter initial optimization

8.3 Diplexer performance

This paragraph gives the predicted diplexer insertion losses and input reflection coefficient in figure 8.5 and the measured performance in figure 8.6. The measured diplexer characteristics as used for the formulation of the design goals in chapter 3 are given in table 8.5. T-junctions and steps are defined assuming a width of 1464 μm. The strip height is set to 10 μm and the tangent of the loss angle to 0.001.
In figure 8.7 the circuit layout is given, enlarged four times.

As previously remarked in chapter 4 and denoted in figure 3.1, the high- and lowpass curve mirrors around the resonance frequency. This means that in figures 8.5 and 8.6 the highpass filter behaves as a bandpass with centre frequency of 9.5 GHz.

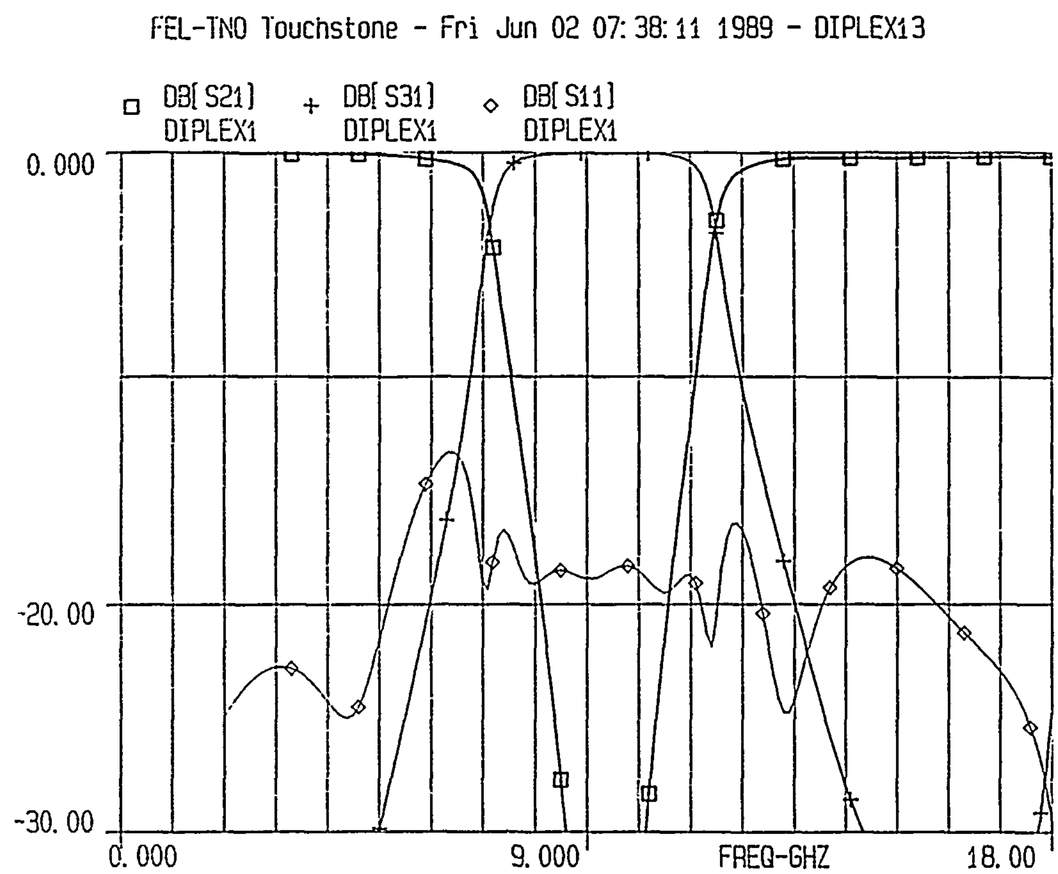

Fig. 8.5: Predicted diplexer performance

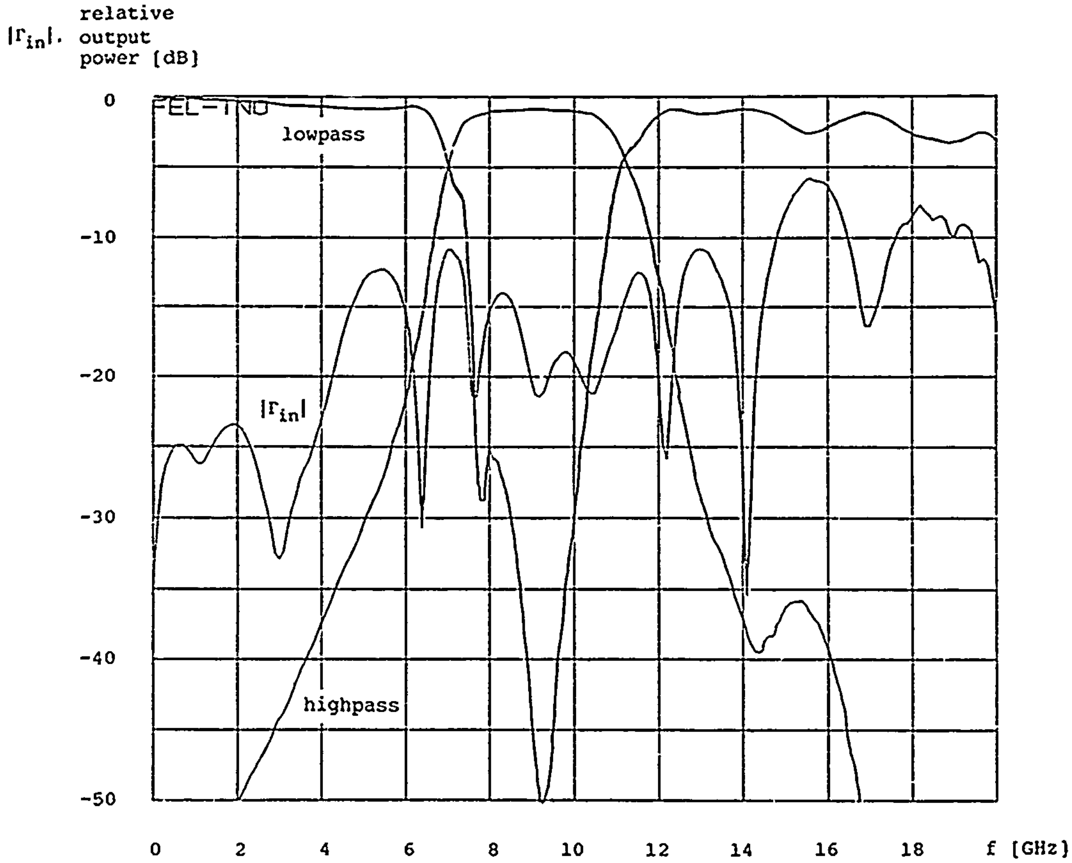

Fig. 8.6: Measured diplexer performance

	specs	measured	units
-3dB frequency	7.5	7.1	GHz
Pass band highpass	7.5 - 11.5	7.1 - 11.2	GHz
Pass band ripple	< 0.5	0.5	dB
Att. lowp. at 8.5 GHz	< -15	-25	dB
Input refl. coeff.	< -10	-11.5	dB

Table 8.5: Measured diplexer characteristics vrs. specs

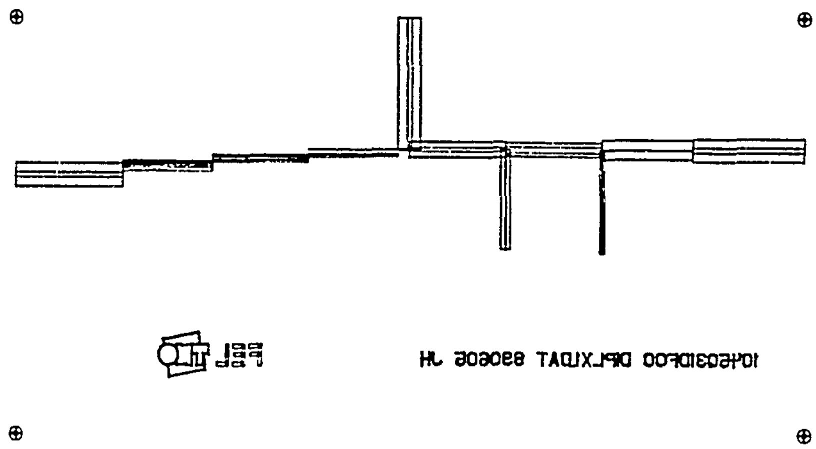

Fig. 8.7: Diplexer circuit lay-out (enlarged 4 times)

9 CONCLUSIONS AND RECOMMENDATIONS

It is feasible to design an X-band diplexer using conventional prototype power transfer functions as a starting point. This allows a convenient separate synthesis of the low- and the highpass filter.

It is highly recommended to use the sequential method. The first phase is the separate design of the low- and the highpass filter in a relative simple and fast initial optimization, using specific topologies. In this step simple transmission line models are used and impedance values are optimized rather than microstrip dimensions. It is possible to precede this first optimization by a fast systematical global search in order to increase the possibility of obtaining global optimum values and to speed up the initial optimization. These values can serve as suitable initial values for a final, local, optimization in which strip height, losses, dispersion- and junction effects, etc. are involved. The total diplexer configuration can be optimized in this final step to yield an input reflection coefficient less than -10 dB (SWR<6 dB).
All microstrip topologies applied, avoid perforation of the substrate. A highpass filter realization using symmetrical coupled lines has the advantage that extraction techniques- and computer programs are widely available which calculate dimensions from the even- and odd impedance values. The degree of design freedom, however, is somewhat limited, and in this case no realizable solution was achieved. Moreover, complex relations for constraints of the optimization variables are necessary. Asymmetrical coupled structures offer a higher degree of freedom, and more realizable solutions are possible, using simple constraint equations. A disadvantage is that element extraction has to take place indirectly, using the hybrid matrix output and the optimization features of the computer program TOUCHSTONE.
It is recommended to execute the lowpass filter initial optimization using a similar straightforward method as described above assuming a topology of series microstrip lines and open-circuited stubs. Exact synthesis with the aid of Kuroda identities can give a highly redundant

circuit, and does not guarantee realizable values. It is believed that a satisfactory solution with a minimum redundancy will result by straightforward analysis-in-the-loop design, imposing simple constraints on the impedance values to be optimized and altering the number of series elements and stubs.

Since the diplexer filters have been designed separately and the software has been developed accordingly, the programs can also be utilized with minor changes for generic low- and highpass microwave filter synthesis.

If this sequential synthesis does not yield satisfactory results, the approach which utilizes only the TOUCHSTONE program (chapter 7) can be applied. The initial circuits for the optimization are easily designed using discrete prototype circuits and redundant synthesis with Kuroda identities. The topology may be adjusted in the design procedure to better meet the design objectives or to reduce the redundancy, which may be considerable.

This method however, is a very time-consuming procedure and does not guarantee convergence or an optimum solution. Moreover, adjusting the topology usually is an intensive trial-and-error procedure. It is obvious that the TOUCHSTONE program is suited for use in the final design step only, in a local optimization mode. The use of this method is recommended only in case none of the options of the sequential method give results which are physically realizable and one has to "force" a microstrip realization.

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CALCULATION OF TWO-PORT MATRICES OF A PARALLEL COUPLED LINE SECTION

The general form of the ABCD matrix for two-port networks is

$$[ABCD] = \begin{vmatrix} a & b \\ c & d \end{vmatrix}$$

For a series capacitor in the s-plane the ABCD matrix is:

$$[ABCD]_C = \begin{vmatrix} 1 & 1/sC \\ 0 & 1 \end{vmatrix} = \begin{vmatrix} 1 & S/s \\ 0 & 1 \end{vmatrix} \qquad (a1.1)$$

and the ABCD matrix of a unit element:

$$[ABCD]_{ue} = \frac{1}{\surd(1-s^2)} \begin{vmatrix} 1 & sZ_0 \\ s/Z_0 & 1 \end{vmatrix} \qquad (a1.2)$$

Let the left side capacitor be denoted by C_1 or S_1 and the right hand side capacitor by C_2 or S_2 as in figure 5.1.1, then the resulting ABCD matrix for an asymmetrical parallel coupled section $[ABCD]_{asym}$ is given by

$$[ABCD]_{asym} = [ABCD]_{C1} * [ABCD]_{ue} * [ABCD]_{C2} =$$
$$\frac{1}{\surd(1-s^2)} \begin{vmatrix} 1 + S_1/Z_0 & (S_1+S_2+S_1S_2/Z_0)/s + Z_0s \\ s/Z_0 & 1 + S_2/Z_0 \end{vmatrix} \qquad (a1.3)$$

If $C_1 = C_2 = C$, or $S_1 = S_2 = S$, the expression for symmetrical coupled sections follows from (a1.3):

$$[ABCD]_{sym} = [ABCD]_C * [ABCD]_{ue} * [ABCD]_C =$$

$$\frac{1}{\sqrt{(1-s^2)}} \begin{vmatrix} 1 + S/Z_0 & (2S + S^2/Z_0)/s + Z_0 s \\ s/Z_0 & 1 + S/Z_0 \end{vmatrix} \quad \text{(a1.4)}$$

From (a1.3) or (a1.4) the hybrid matrix

$$[H] = \begin{vmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{vmatrix}$$

is determined, utilizing the general conversion relations for two-ports:

$$\begin{aligned} h_{11} &= b/d \\ h_{12} &= \det(ABCD)/d \\ h_{21} &= -1/d \\ h_{22} &= c/d \end{aligned} \quad \text{(a1.5)}$$

with det(ABCD) = ad - bc.

Since the parallel coupled line section is a reciprocal and -ideally- lossless device, it can be shown that

$$Re[h_{11}] = Im[h_{12}] = Re[h_{22}] = 0 \quad \text{(a1.6)}$$

$h_{12} = h_{21}$

INDEX OF FORTRAN PROGRAM SOURCES

Within FEL-TNO the following FORTRAN 77 programs have been developed:

HIGHPASS_SYM	:	calculation of the initial optimized impedance values of the symmetrical highpass filter
HIGHPASS_ASYM	:	calculation of the initial optimized impedance values of the asymmetrical highpass filter
LOWPASS	:	calculation of the initial optimized impedance values of the lowpass filter

UNCLASSIFIED

REPORT DOCUMENTATION PAGE (MOD-NL)

1. DEFENSE REPORT NUMBER (MOD-NL) TD90-1726	2. RECIPIENT'S ACCESSION NUMBER	3. PERFORMING ORGANIZATION REPORT NUMBER FEL-90-B176
4. PROJECT/TASK/WORK UNIT NO. 1045	5. CONTRACT NUMBER	6. REPORT DATE JUNE 1990
7. NUMBER OF PAGES 49 (INCL. RDP & TITLEPAGE & 2 APPENDICES, EXCL. DISTRIBUTIONLIST)	8. NUMBER OF REFERENCES 5	9. TYPE OF REPORT AND DATES COVERED FINAL REPORT

10. TITLE AND SUBTITLE
A DESIGN METHOD FOR AN X-BAND DIPLEXER

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12. PERFORMING ORGANIZATION NAME(S) AND ADDRESS(ES)
TNO PHYSICS AND ELECTRONICS LABORATORY
POBOX 96864, 2509 JG, THE HAGUE

13. SPONSORING/MONITORING AGENCY NAME(S)

14. SUPPLEMENTARY NOTES

15. ABSTRACT (MAXIMUM 200 WORDS, 1044 POSITIONS)
THIS REPORT PROVIDES A DETAILED OVERVIEW OF A SEQUENTIAL METHOD FOR THE DESIGN OF A PASSIVE X-BAND DIPLEXER IN ORDER TO SUPPORT FUTURE DESIGN ACTIVITIES IN A SIMILAR FIELD. ALSO DESIGN STEPS AND DIFFICULTIES FOR DIPLEXER DESIGN ARE DEALT WITH IN CASE ONLY THE COMPUTER PROGRAM TOUCHSTONE® [4] IS USED. PREDICTED AND MEASURED PERFORMANCE OF A REALIZED DIPLEXER ARE GIVEN.
AN OVERVIEW OF FORTRAN 77 PROGRAMS, DEVELOPED AT FEL-TNO IS INCLUDED. THIS SOFTWARE ALSO CAN BE USED FOR SEPARATE SYNTHESIS OF LOW- AND HIGHPASS MICROSTRIP FILTERS.

16. DESCRIPTORS	IDENTIFIERS
DESIGN	DIPLEXER
HIGH-PASS FILTER	MICROSTRIP
LOW-PASS FILTER	
COMPUTER PROGRAM	
FORTRAN COMPUTER AIDED DESIGN	

17a. SECURITY CLASSIFICATION (OF REPORT) UNCLASSIFIED	17b. SECURITY CLASSIFICATION (OF PAGE) UNCLASSIFIED	17c. SECURITY CLASSIFICATION (OF ABSTRACT) UNCLASSIFIED

18. DISTRIBUTION/AVAILABILITY STATEMENT UNLIMITED AVAILABILITY	17d. SECURITY CLASSIFICATION (OF TITLES) UNCLASSIFIED

UNCLASSIFIED