SELECTED NUMERICAL EXAMPLES OF DESIGN OPTIMIZATION OF MULTI-COUPLED CAVITY MICROWAVE FILTERS

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Abstract

Selected numerical examples of the design optimization of multi-coupled cavity microwave filters are presented. The dual-symmetrical coupling configuration is considered. Examples include asynchronously tuned filters realizing asymmetric characteristics and simultaneous optimization of amplitude and group delay responses resulting in nonminimum-phase filters. Network variables as well as optimization parameters are provided.

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I. INTRODUCTION

An approach to interactive design optimization of multi-coupled cavity microwave filters has been proposed in [1]. Implementing our approach on a CDC 170/730 digital computer, a computer program has been constructed and successfully tested for various filter design problems. An efficient method of filter simulation and sensitivity evaluation, namely the loaded filter approach, as presented in Section IV of [2], is utilized. Here, we select numerical examples of filters with dual-symmetrical coupling pattern and resistive terminations. A review of the configuration of such filters can be found in Section III of [2]. Examples presented include asynchronously tuned filters realizing asymmetric characteristics and simultaneous optimization of amplitude and group delay responses resulting in nonminimum-phase function filters. Besides the solutions, we also provide relevant information such as the starting points, original specifications, optimization parameters and transfer functions in order to facilitate case study, further investigations or independent verification.

In describing filter responses, the following abbreviations are used: RC – reflection coefficient, RL – return loss, IL – insertion loss and GD – relative group delay.

II. THE FILTER MODEL

All the examples presented are based on a 6-pole model, centered at 4000 MHz with 1% (40 MHz) bandwidth. The structure of the filter network is shown in Fig. 1. The nonzero elements of the coupling matrix **M**, as used for our examples, are illustrated in Fig. 2. The filter cavities are coupled in a dual-symmetrical pattern. This implies, in terms of the elements of **M**, that

$$M_{\ell k} = M_{k\ell} = M_{\sigma\tau} = M_{\tau\sigma}, \tag{1}$$

where $\sigma = n+1-\ell$ and $\tau = n+1-k$, n being the order of the filter.

The solutions of the examples are given in terms of the nonzero couplings and transformer ratios. Since the couplings are related by (1), only the distinct coupling values are given, e.g., for $M_{12} = M_{21} = M_{56} = M_{65}$, only M_{12} is given. The terminations, as shown in Fig. 1, are normalized such that we have the load resistor $R_L = 1\ \Omega$ and the voltage source $E = 1$ V with a resistor $R_S = 1\ \Omega$.

III. DESIGN 1

Filter Type

Symmetric optimum amplitude characteristics.

Solution

$$M_{12} = 0.81777$$

$$M_{23} = 0.51110$$

$$M_{34} = 0.82430$$

$$M_{16} = 0.09330$$

$$M_{25} = -0.35710$$

$$n_1^2 = n_2^2 = 0.98239$$

Simulated Responses

Lower stopband:	– 3976 MHz, minimum IL 34.2 dB
Passband:	3980 – 4020 MHz, minimum RL 20 dB
Upper stopband:	4024 – MHz, minimum IL 34.2 dB

The simulated responses are also shown in Figures 3 and 4.

Comment

Filters with electrically symmetric amplitude characteristics are most clearly defined and extensively studied in the conventional literature. An optimum solution of such a typical problem, besides having significance of its own, provides a basic starting point for our further investigation.

Starting Point*

$$M_{12} = 0.8101$$

$$M_{23} = 0.4894$$

$$M_{34} = 0.8450$$

$$M_{16} = 0.1197$$

$$M_{25} = -0.4010$$

$$n_1^2 = n_2^2 = 0.9600$$

* Source: Shamasundara et al. [3].

Table of Subinterval Data

Frequency edges of subinterval (MHz)	No. of sample points	Step-length of extrema location	Specification	Weighting factor
3950 – 3970	4	2.0	RC = 0.9993	−1.0
3970 – 3976	4	1.0	RC = 0.9993	−1.0
3980 – 4001	8	0.5	RC = 0.1	1.0

Optimization Parameters

Variables: $M_{12}, M_{23}, M_{34}, M_{16}, M_{25}, n_1^2, n_2^2$

Initial step-length: 0.01

Accuracy requirement: 1.0×10^{-6}

Solution obtained after 17 iterations with 5.7 CPU sec.

Values of the active objective functions: -5.1×10^{-4}

Transfer Function Analysis

The voltage transfer ratio $H(s) \triangleq c\,N(s)/D(s)$, where

$$D(s) = \prod_{j=1}^{6} (s-p_j)\ ,\quad N(s) = \prod_{i=1}^{4} (s-z_i)\ \text{ and }\ c = j\,0.091657$$

The poles p_j: $-0.062739 \pm j\,1.056104$, $-0.277563 \pm j\,0.942653$, $-0.642085 \pm j\,0.447049$

The zeros z_i: $0.000000 \pm j\,1.542819$, $0.000000 \pm j\,1.228307$

The pole-zero pattern of H(s) is shown in Fig. 4.

Real frequency loss poles (MHz): 3969.3, 3975.5, 4024.6, 4031.0.

Reflection zeros (MHz): 3980.4, 3984.1, 3993.5, 4006.5, 4016.0, 4019.7.

IV. DESIGN 2

Filter Type

Asynchronously tuned filter realizing asymmetric characteristics.

Solution

$$M_{11} = -0.03315$$

$$M_{22} = -0.05918$$

$$M_{33} = 0.69943$$

$$M_{12} = 0.83234$$

$$M_{23} = 0.46166$$

$$M_{34} = 0.29296$$

$$M_{16} = -0.07894$$

$$M_{25} = 0.33007$$

$$M_{15} = 0.03999$$

$$M_{24} = -0.31780$$

$$n_1^2 = n_2^2 = 0.99623$$

Simulated Responses

Lower stopband: – 3978 MHz, minimum IL 40.7 dB

Passband: 3980 – 4020 MHz, minimum RL 20 dB

Upper stopband: 4035 – MHz, minimum IL 34.9 dB

The simulated responses are also shown in Figures 6 and 7.

Comment

Asynchronously tuned filters realizing asymmetric characteristics, as demonstrated by Cameron [4], acquire increasing attention in the development of modern communication systems. Our software provides a fast and flexible approach to the design of such filters. Three examples, with different passband specifications, are presented in sequence.

Starting Point

The solution of Design 1.

Table of Subinterval Data

Frequency edges of subinterval (MHz)	No. of sample points	Step-length of extrema location	Specification	Weighting factor
3950 – 3974	4	2.0	RC = 0.99995	– 1.0
3974 – 3978	4	0.4	RC = 0.99995	– 1.0
3980 – 3983	4	0.5	RC = 0.1	1.0
3983 – 4020	7	2.0	RC = 0.1	1.0
4035 – 4050	4	3.0	RC = 0.9998	-1.0

Optimization Parameters

Variables: $M_{11}, M_{22}, M_{33}, M_{12}, M_{23}, M_{34}, M_{16}, M_{25}, M_{15}, M_{24}, n_1^2, n_2^2$

Initial step-length: 0.005

Accuracy requirement: 1.0×10^{-6}

Solution obtained after 283 iterations with 175 CPU sec.

Values of the active objective functions: -7.4×10^{-4}

Transfer Function Analysis

The voltage transfer ratio $H(s) \triangleq c\, N(s)/D(s)$, where

$$D(s) = \prod_{j=1}^{6} (s - p_j) \;, \quad N(s) = \prod_{i=1}^{4} (s - z_i) \text{ and } c = -j\,0.078642$$

The poles p_j: $-0.027647 - j\,1.025974$, $-0.121258 - j\,0.981043$, $-0.348255 - j\,0.800175$

$-0.678583 - j\,0.221800$, $-0.620329 + j\,0.674959$, $-0.196389 + j\,1.139832$

The zeros z_i: $0.000000 - j\,1.675102$, $0.000000 - j\,1.215036$, $0.000000 - j\,1.112455$

$0.000000 + j\,1.878787$.

The pole-zero pattern of H(s) is shown in Fig. 8.

Real frequency loss poles (MHz): 3966.6, 3975.8, 3977.8, 4037.8.

Reflection zeros (MHz): 3980.2, 3981.8, 3986.9, 3997.4, 4016.6, 4018.9.

Comment

The transfer function obtained assumes an asymmetric pole-zero pattern as expected. The sharp cut-off of the lower transition band corresponds to a dense distribution of poles and zeros. Also, notice that the zeros of H(s) no longer appear in conjugate pairs hence N(s) may have complex coefficients. Letting $s = j\omega$, however, we find $N(j\omega)$ to be a real function in ω, which is a necessary condition of realizability [5].

V. DESIGN 3

Filter Type

Asynchronously tuned filter realizing asymmetric characteristics.

Solution

$$M_{11} = -0.024386$$

$$M_{22} = -0.043917$$

$$M_{33} = 0.68523$$

$$M_{12} = 0.75282$$

$$M_{23} = 0.44064$$

$$M_{34} = 0.29111$$

$$M_{16} = -0.051903$$

$$M_{25} = 0.29976$$

$$M_{15} = 0.025149$$

$$M_{24} = -0.29633$$

$$n_1^2 = n_2^2 = 0.79421$$

Simulated Responses

Lower stopband:	– 3978 MHz,	minimum IL 45.9 dB
Passband:	3980 – 4020 MHz,	minimum RL 15 dB
Upper stopband:	4035 – 4037 MHz,	minimum IL 35 dB
	4037 – MHz,	minimum IL 42 dB

The simulated responses are also shown in Figures 9 and 10.

Starting Point

The solution of Design 2.

Table of Subinterval Data

Frequency edges of subinterval (MHz)	No. of sample points	Step-length of extrema location	Specification	Weighting factor
3950 – 3974	4	2.0	RC = 0.99995	– 1.0
3974 – 3978	4	0.4	RC = 0.99995	– 1.0
3980 – 3983	4	0.5	RC = 0.17783	1.0
3983 – 4020	7	2.0	RC = 0.17783	1.0
4035 – 4050	4	3.0	RC = 0.9998	-1.0

Optimization Parameters

Variables: $M_{11}, M_{22}, M_{33}, M_{12}, M_{23}, M_{34}, M_{16}, M_{25}, M_{15}, M_{24}, n_1^2, n_2^2$

Initial step-length: 0.005

Accuracy requirement: 1.0×10^{-6}

Solution obtained after 251 iterations with 115.6 CPU sec.

Values of the active objective functions: -3.7×10^{-5}

Transfer Function Analysis

The voltage transfer ratio $H(s) \triangleq c\, N(s)/D(s)$, where

$$D(s) = \prod_{j=1}^{6} (s-p_j)\ ,\quad N(s) = \prod_{i=1}^{4} (s-z_i) \text{ and } c = -j\,0.041222$$

The poles p_j: $-0.023772 - j\,1.015327$, $-0.102599 - j\,0.956950$, $-0.283025 - j\,0.749747$

$-0.525652 - j\,0.193357$, $-0.487414 + j\,0.609179$, $-0.165960 + j\,1.072348$

The zeros z_i: $0.000000 - j\,1.670345$, $0.000000 - j\,1.214280$, $0.000000 - j\,1.112421$

$0.000000 + j\,1.984880$.

The pole-zero pattern of H(s) is shown in Fig. 8.

Real frequency loss poles (MHz): 3966.7, 3975.8, 3977.8, 4040.0.

Reflection zeros (MHz): 3980.2, 3981.8, 3986.9, 3997.3, 4010.5, 4018.9.

V. DESIGN 4

Filter Type

Asynchronously tuned filter realizing asymmetric characteristics.

Solution

$$M_{11} = -0.05092$$

$$M_{22} = -0.07059$$

$$M_{33} = 0.76097$$

$$M_{12} = 0.91876$$

$$M_{23} = 0.44780$$

$$M_{34} = 0.24704$$

$$M_{16} = -0.08815$$

$$M_{25} = 0.41092$$

$$M_{15} = 0.01462$$

$$M_{24} = -0.31096$$

$$n_1^2 = n_2^2 = 1.18083$$

Simulated Responses

Lower stopband:	– 3978 MHz,	minimum IL 37 dB
Passband:	3980 – 4020 MHz,	minimum RL 24.3 dB
Upper stopband:	4035 – 4044 MHz,	IL 20 → 40 dB
	4044 – MHz,	minimum IL 40 dB

The simulated responses are also shown in Figures 11 and 12.

Starting Point

The solution here is obtained with the lower stopband specification weighted by –100.0. The result shows that the only active objective function in the specified upper stopband occurs at 4035 MHz, an edge point. Another solution with lower stopband weighted by –1.0 gives a non-equiripple response in the lower stopband. This may be explained by the realizability restriction on asymmetric characteristics and the tight passband specification (RL = 25 dB) of this case.

Starting Point

The solution of Design 2.

Table of Subinterval Data

Frequency edges of subinterval (MHz)	No. of sample points	Step-length of extrema location	Specification	Weighting factor
3950 – 3974	4	2.0	RC = 0.99995	–1.0
3974 – 3978	4	0.4	RC = 0.99995	–1.0
3980 – 3983	4	0.5	RC = 0.056234	1.0
3983 – 4020	7	2.0	RC = 0.056234	1.0
4035 – 4050	4	3.0	RC = 0.9999	-1.0

Optimization Parameters

Variables: $M_{11}, M_{22}, M_{33}, M_{12}, M_{23}, M_{34}, M_{16}, M_{25}, M_{15}, M_{24}, n_1^2, n_2^2$

Initial step-length: 0.005

Accuracy requirement: 1.0×10^{-6}

Solution obtained after 88 iterations with 40.9 CPU sec.

Values of the active objective functions: -4.95×10^{-3}

Transfer Function Analysis

The voltage transfer ratio $H(s) \triangleq c\,N(s)/D(s)$, where

$$D(s) = \prod_{j=1}^{6} (s-p_j)\ ,\quad N(s) = \prod_{i=1}^{4} (s-z_i)\ \text{ and }\ c = -j\,0.041222$$

The poles p_j: −0.029474 − j 1.035361, −0.130029 − j 1.003869, −0.384281 − j 0.857209

−0.791314 − j 0.293041, −0.767075 + j 0.683301, −0.259481 + j 1.227260

The zeros z_i: 0.000000 − j 1.658533, 0.000000 − j 1.212477, 0.000000 − j 1.112392

0.000000 + j 2.297883.

The pole-zero pattern of H(s) is shown in Fig. 8.

Real frequency loss poles (MHz): 3967.0, 3975.8, 3977.8, 4046.2.

Reflection zeros (MHz): 3980.2, 3981.8, 3986.7, 3996.9, 4010.1, 4018.8.

V. DESIGN 5

Filter Type

Nonminimum-phase, simultaneous optimization of the amplitude and group delay responses.

Solution

$$M_{12} = 0.84322$$

$$M_{23} = 0.60733$$

$$M_{34} = 0.62970$$

$$M_{16} = -0.11003$$

$$M_{25} = 0.03351$$

$$n_1^2 = n_2^2 = 0.99826$$

Simulated Responses

Lower stopband:	– 3973.5 MHz, minimum IL 25 dB
Passband:	3980 – 4020 MHz, minimum RL 16.5 dB
Upper stopband:	4026.5 – MHz, minimum IL 25 dB
Group Delay:	3985 – 4015 MHz, maximum GD 2 ns
	(relative to delay minima at 3989 and 4011 MHz)

The simulated responses are also shown in Figures 13 and 14.

Comment

The advantages of nonminimum-phase filters, such as favourable physical size and superior electrical behaviour, have been demonstrated by Atia and Williams [6] by a self-equalized, optimum amplitude 12-pole filter. Here, simultaneous optimization of amplitude and group delay responses on our 6-pole test example also yields positive results.

Starting Point

The solution of Design 1.

Comment

The final solution of this design is obtained by a two-stage optimization.

Table of Subinterval Data (1st Run)

Frequency edges of subinterval (MHz)	No. of sample points	Step-length of extrema location	Specification	Weighting factor
3952 – 3967	4	2.0	RC = 0.9999	–1.0
3967 – 3974	4	1.0	RC = 0.9999	–1.0
3980 – 4001	8	0.5	RC = 0.1	1.0
3984 – 4000	2	2 fixed points at 3984 and 4000	GD = 0.0	1.0

Comment

After 109 iterations an intermediate result was obtained as

$$M_{12} = 0.83147$$

$$M_{23} = 0.59920$$

$$M_{34} = 0.59950$$

$$M_{16} = -0.11060$$

$$M_{25} = 0.0285$$

$$n_1^2 = n_2^2 = 0.99464.$$

By simulating this intermediate design, it was found that the minimum of the group delay within the range 3984 – 4000 MHz shifted from the center frequency (4000 MHz) to 3990 MHz. Taking this into account, we modified the specifications and restarted optimization from the intermediate result given above.

Table of Subinterval Data (2nd Run)

Frequency edges of subinterval (MHz)	No. of sample points	Step-length of extrema location	Specification	Weighting factor
3950 – 3968	4	3.0	RC = 0.99992	−100.0
3968 – 3974	4	1.0	RC = 0.99992	−100.0
3980 – 4001	8	0.5	RC = 0.1	10.0
3985 – 4000	3	3 fixed points at 3985, 3990, 4000	GD = 0.0	0.3

Optimization Parameters

Variables: M_{12}, M_{23}, M_{34}, M_{16}, M_{25}, n_1^2, n_2^2 (for both the 1st and 2nd runs).

Initial step length: 0.005.

Accuracy requirement: 1.0×10^{-6} (for both runs).

The final solution was obtained after 109 + 20 iterations with 41.3 + 6.7 CPU sec.

Values of the active objective functions: 0.636.

Transfer Function Analysis

The voltage transfer ratio $H(s) \triangleq c\, N(s)/D(s)$, where

$$D(s) = \prod_{j=1}^{6} (s - p_j)\ , \quad N(s) = \prod_{i=1}^{4} (s - z_i) \quad \text{and} \quad c = j\,0.109839$$

The poles p_j: $-0.112037 \pm j\,1.123742$, $-0.426870 \pm j\,0.811198$, $-0.459353 \pm j\,0.240363$

The zeros z_i: $0.000000 \pm j\,1.413932$, $\pm\, 0.804559 + j\,0.000000$

The pole-zero pattern of H(s) is also shown in Fig. 15.

Real frequency loss poles (MHz): 3971.8, 4028.4.

Comment

One of the zeros of H(s), namely + 0.804559, is located at the right-half-plane of s. H(s) is a nonminimum-phase function as expected. From network theory we know that the number of finite zeros of H(s) $\leq$ n$-$2, n being the order of the filter. For the 6th-order filter considered here, we could at most have 4 finite zeros of H(s). In the case that only amplitude is optimized, we have seen that all 4 zeroes are on the $s=j\omega$ axis, so we have maximum number of loss poles at finite real frequency. For a nonminimum-phase design, however, some of the zeroes are removed from the $j\omega$ axis in order to improve the phase response of the filter. As in this example, we have only 2 loss poles at finite frequency.

VIII. CONCLUSION

Five examples of filter design, including one symmetric optimum amplitude design, three asynchronously tuned designs with different passband specifications and one nonminimum-phase design, have been presented. A powerful gradient-based, interactive, user-oriented computer program offers efficient minimax optimization of filter design. It directly solves for the coupling values of multi-coupled cavity filters and accommodates special filter characteristics of great interest to modern communication systems, such as asymmetric responses and optimal tradeoffs between amplitude and group delay responses. The convenience in the treatment of various engineering specifications using computer-aided design techniques is vigorously illustrated.

ACKNOWLEDGEMENT

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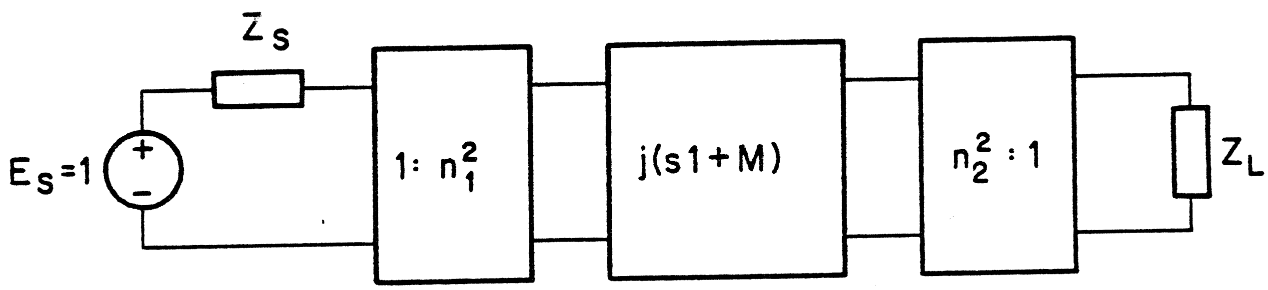

Fig. 1 Block representation of the overall filter network.

$$
\begin{bmatrix}
+ & * & & & + & * \\
* & + & * & + & * & + \\
 & * & + & * & + & \\
 & + & * & + & * & \\
+ & * & + & * & + & * \\
* & + & & & * & +
\end{bmatrix}
$$

Fig. 2 Nonzero elements of the coupling matrix
* – synchronously tuned filters
+,* – asynchronously tuned filters

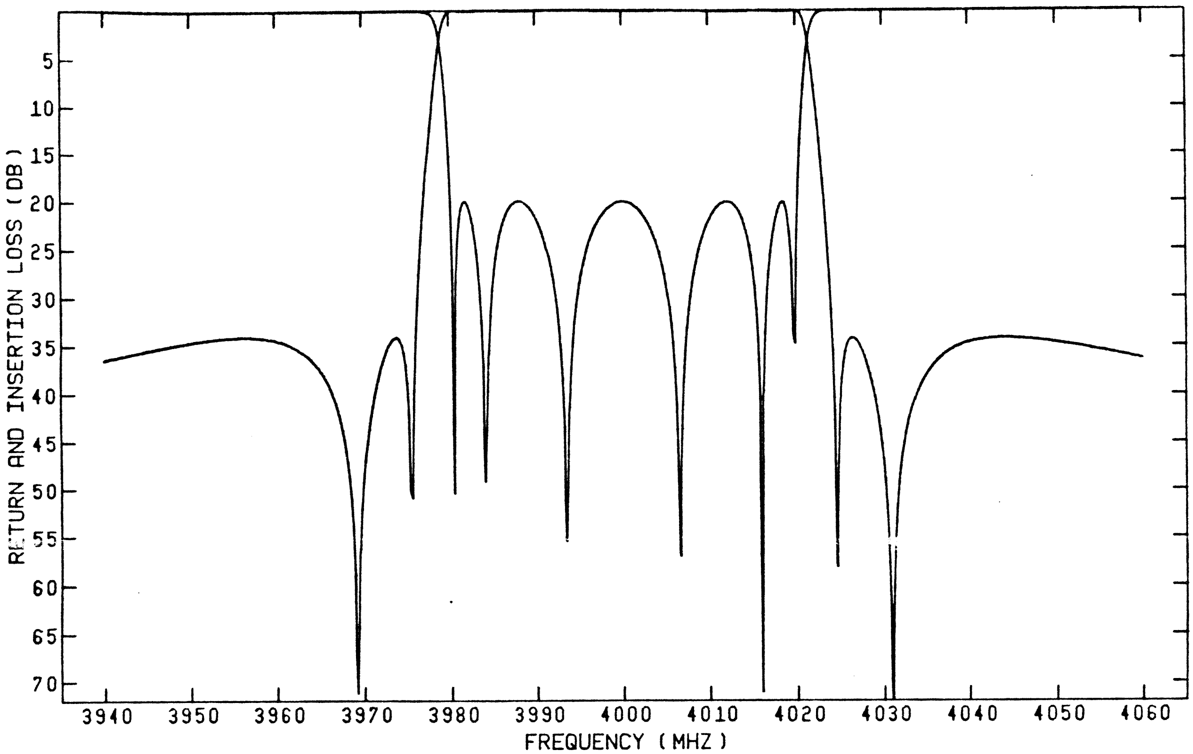

Fig. 3 Return loss and insertion loss of Design 1.

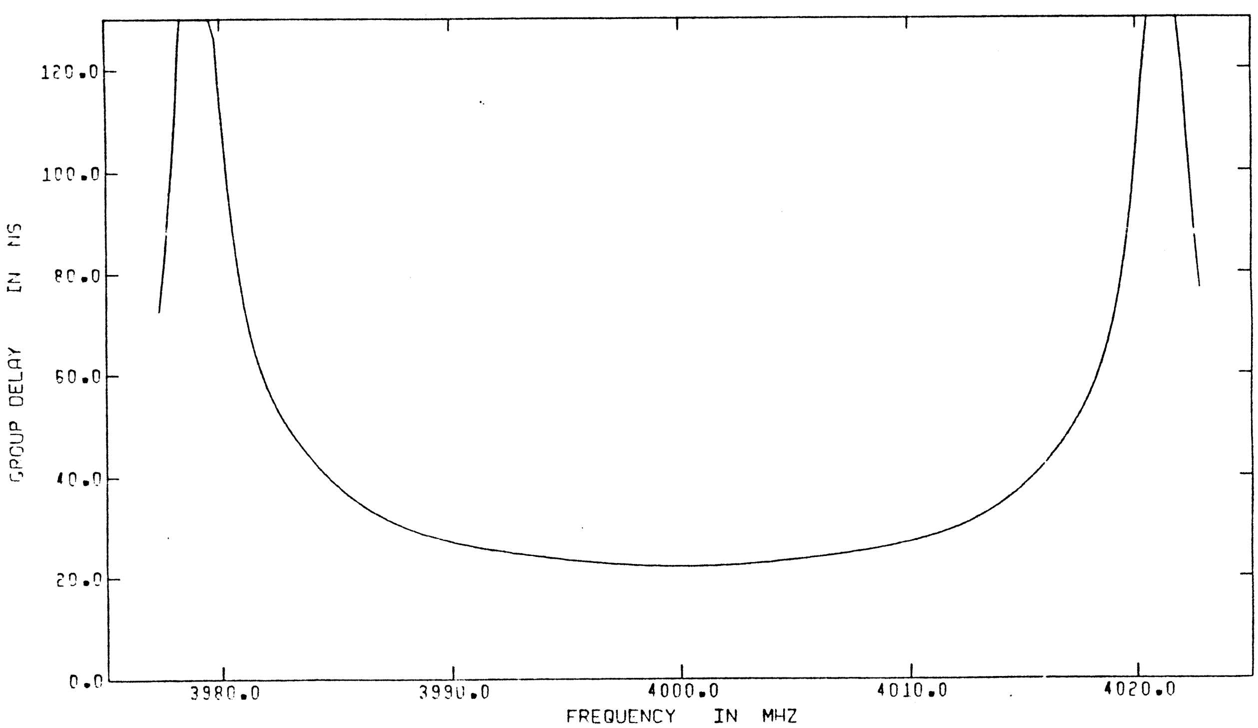

Fig. 4 Group delay of Design 1.

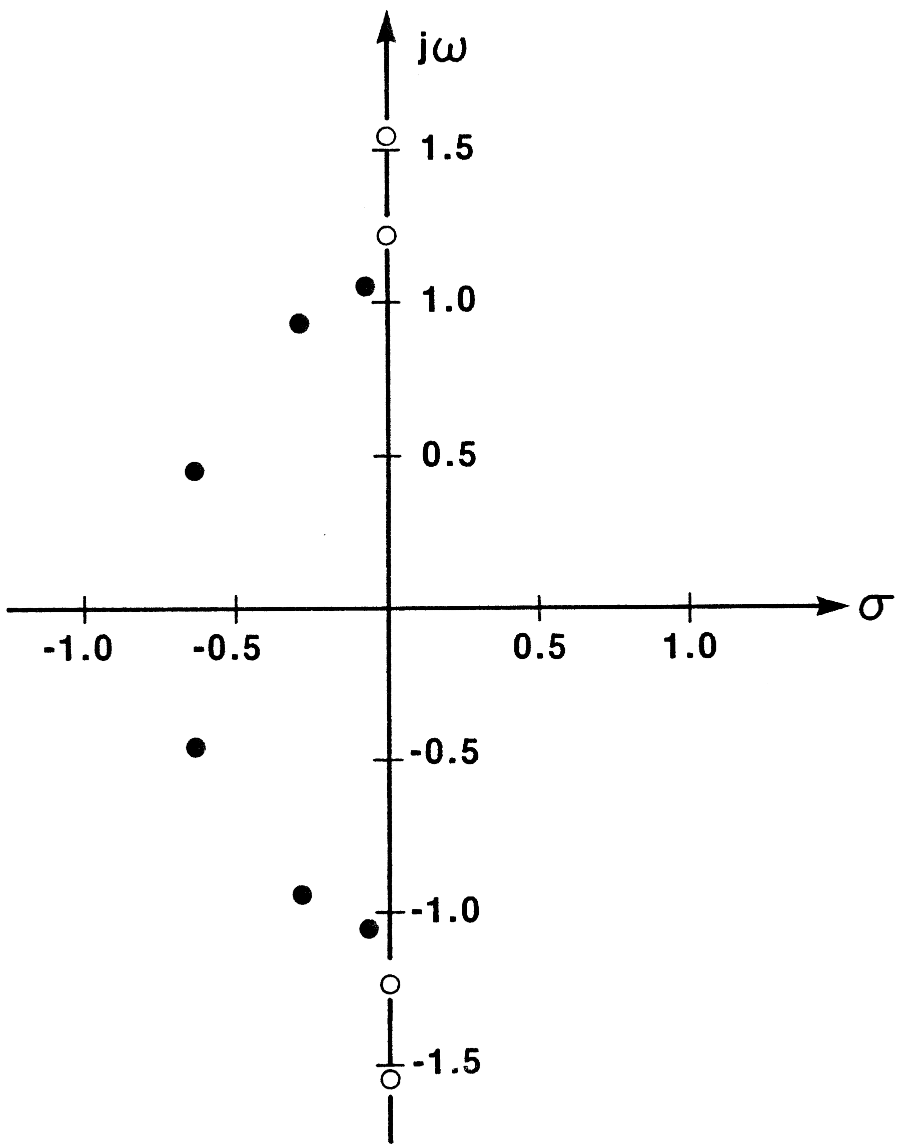

Fig. 5 Pole-zero pattern of the transfer function of Design 1. ●— poles, o — zeros.

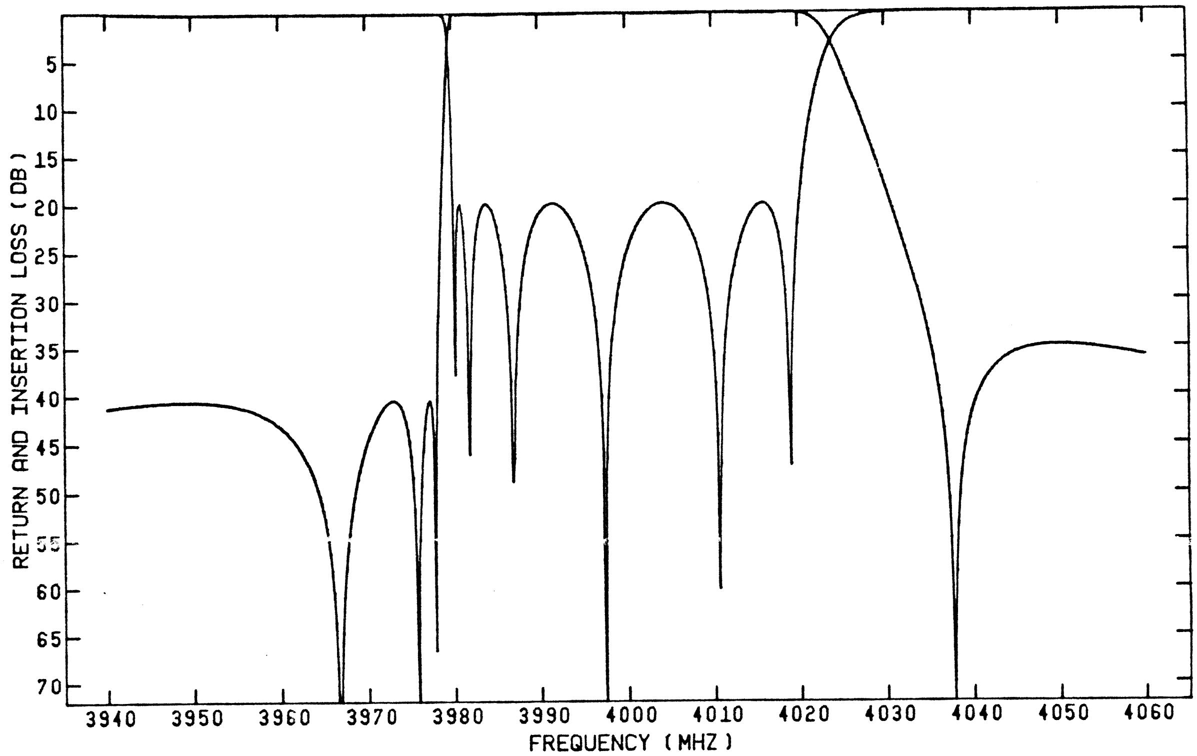

Fig. 6 Return loss and insertion loss of Design 2.

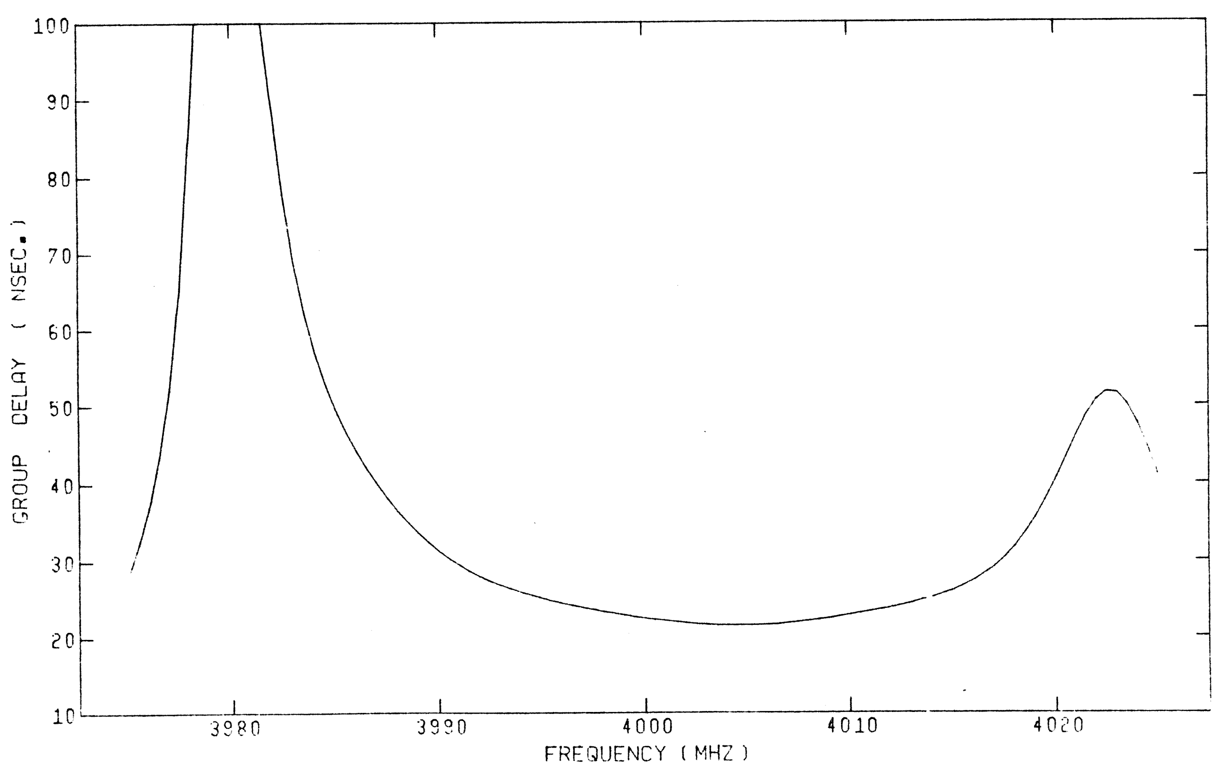

Fig. 7 Group delay of Design 2.

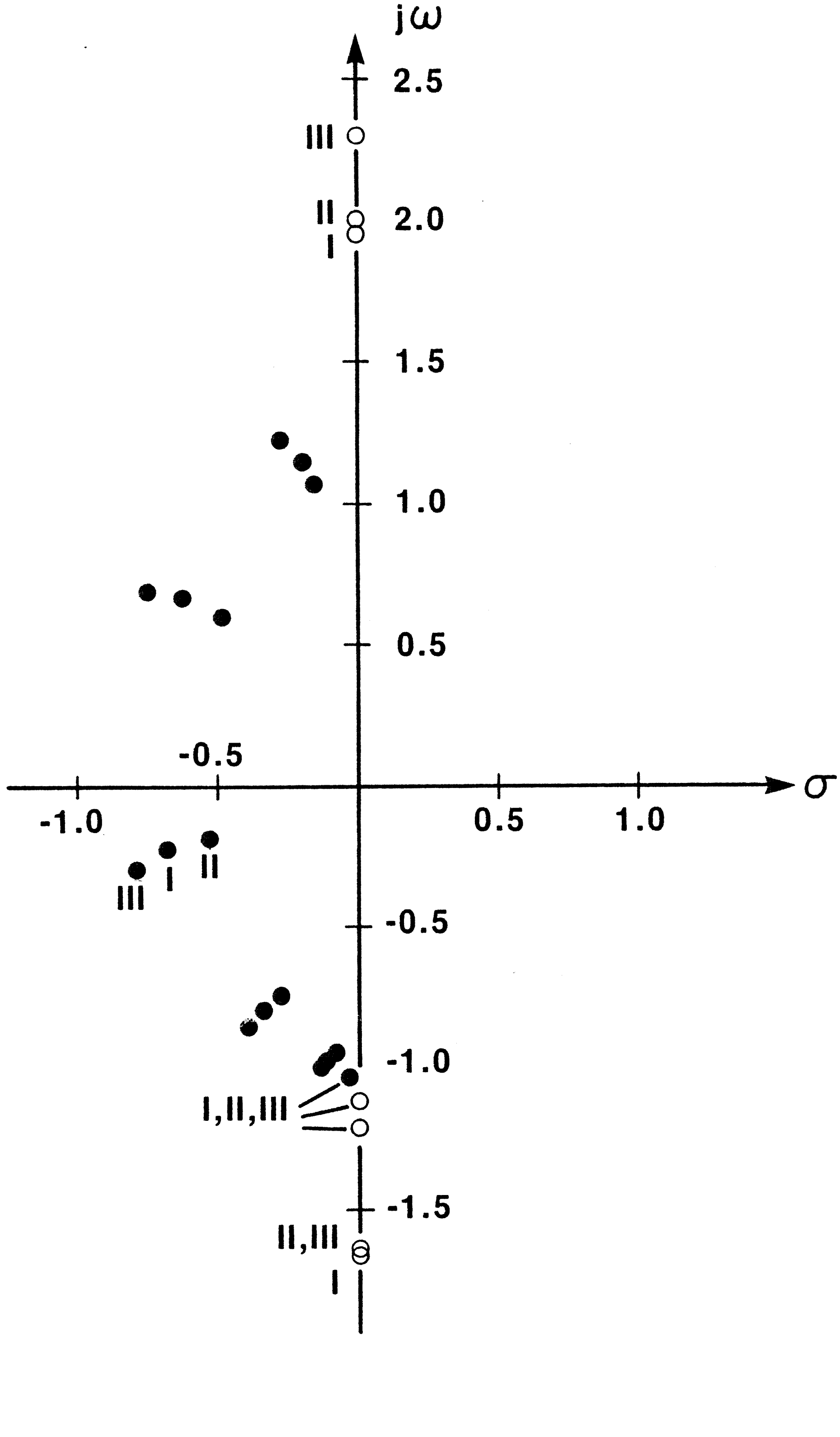

Fig. 8 Pole-zero pattern of the transfer functions of Designs 2, 3 and 4.
●– poles, o – zeros
I – Design 2, II - Design 3, III - Design 4.

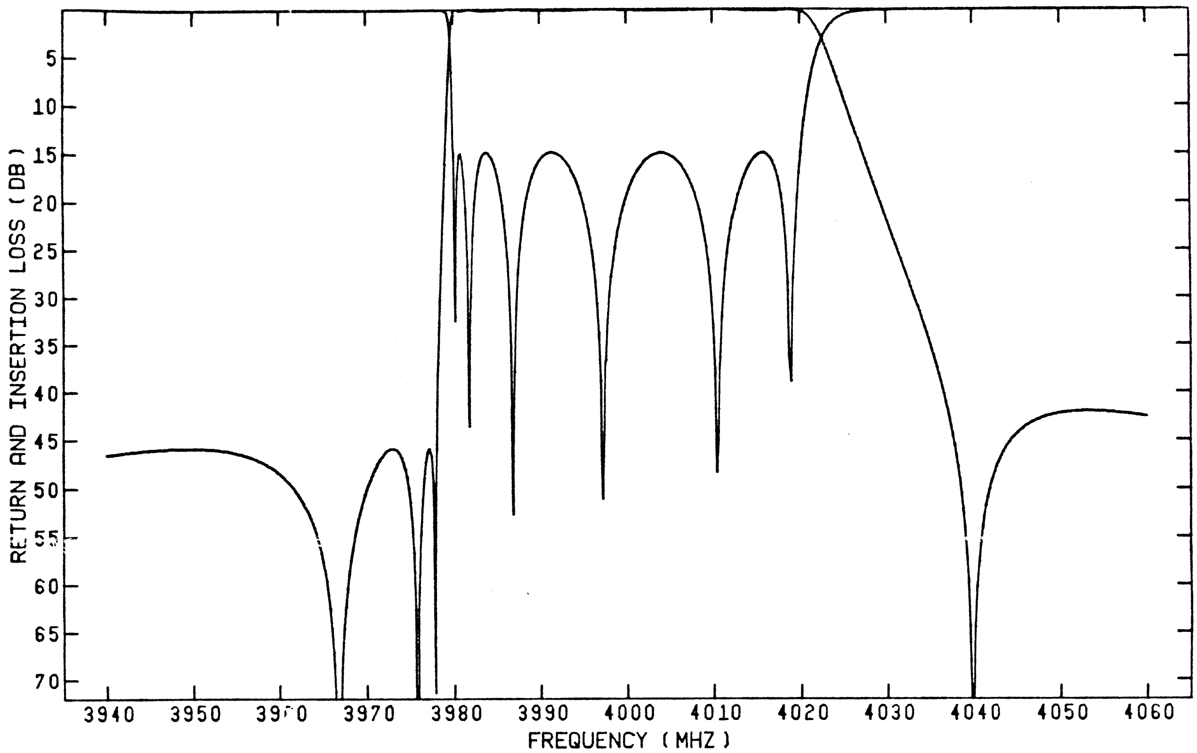

Fig. 9 Return loss and insertion loss of Design 3.

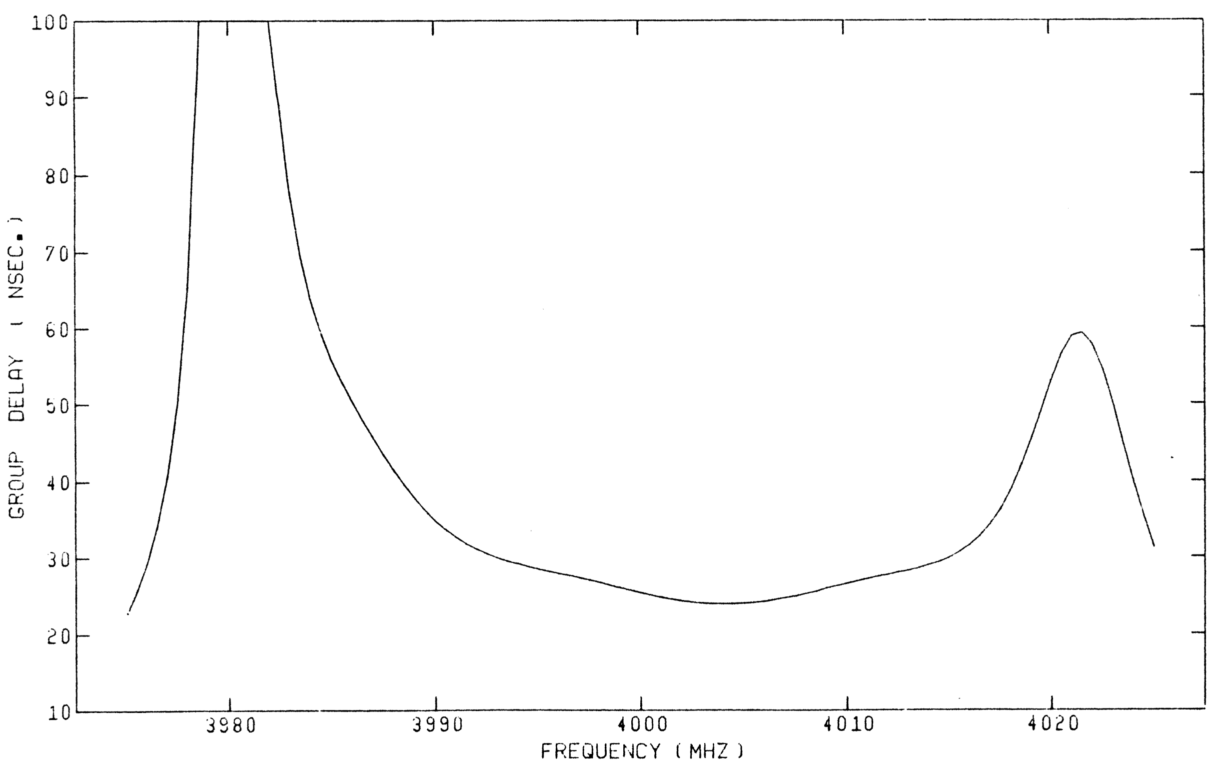

Fig. 10 Group delay of Design 3.

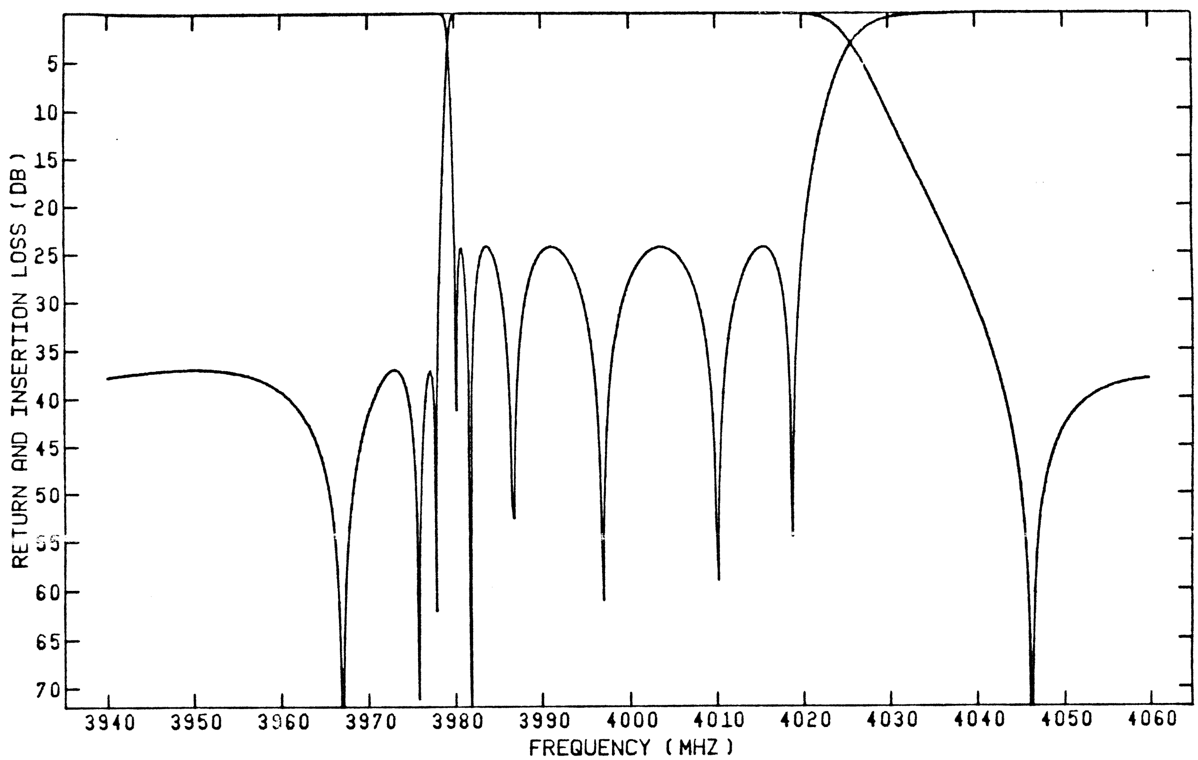

Fig. 11 Return loss and insertion loss of Design 4.

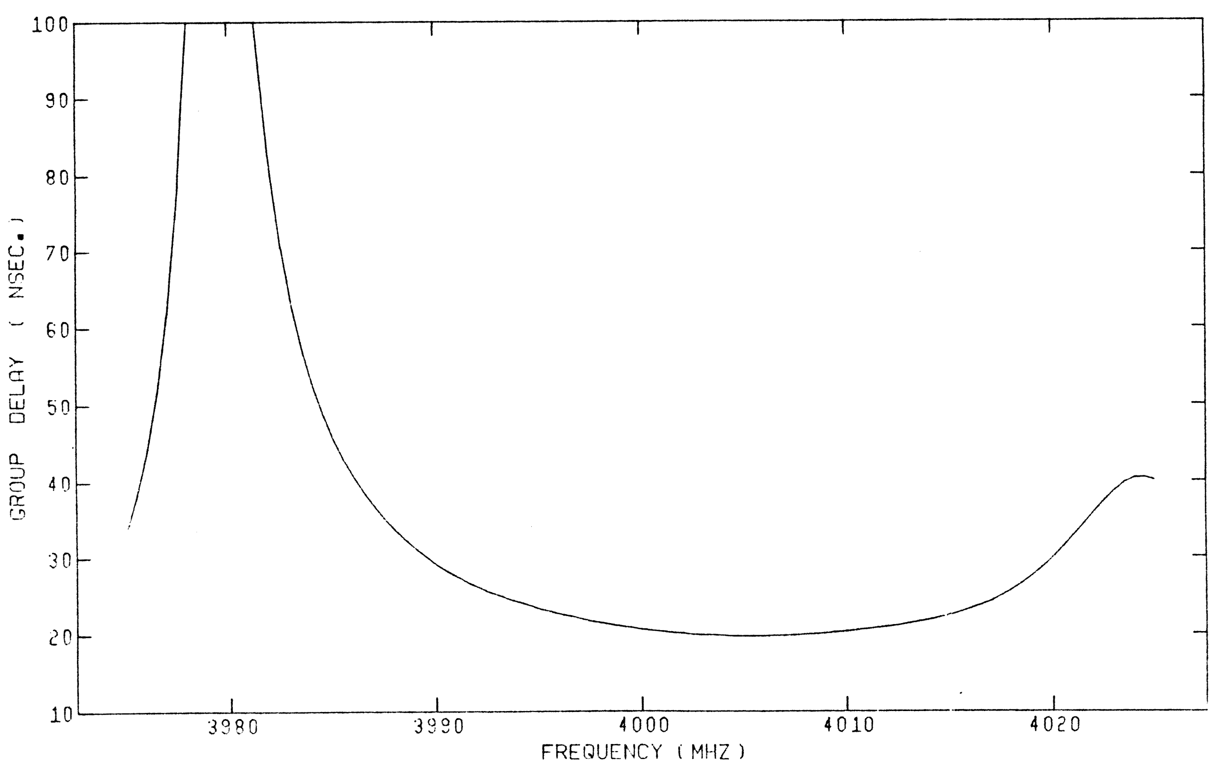

Fig. 12 Group delay of Design 4.

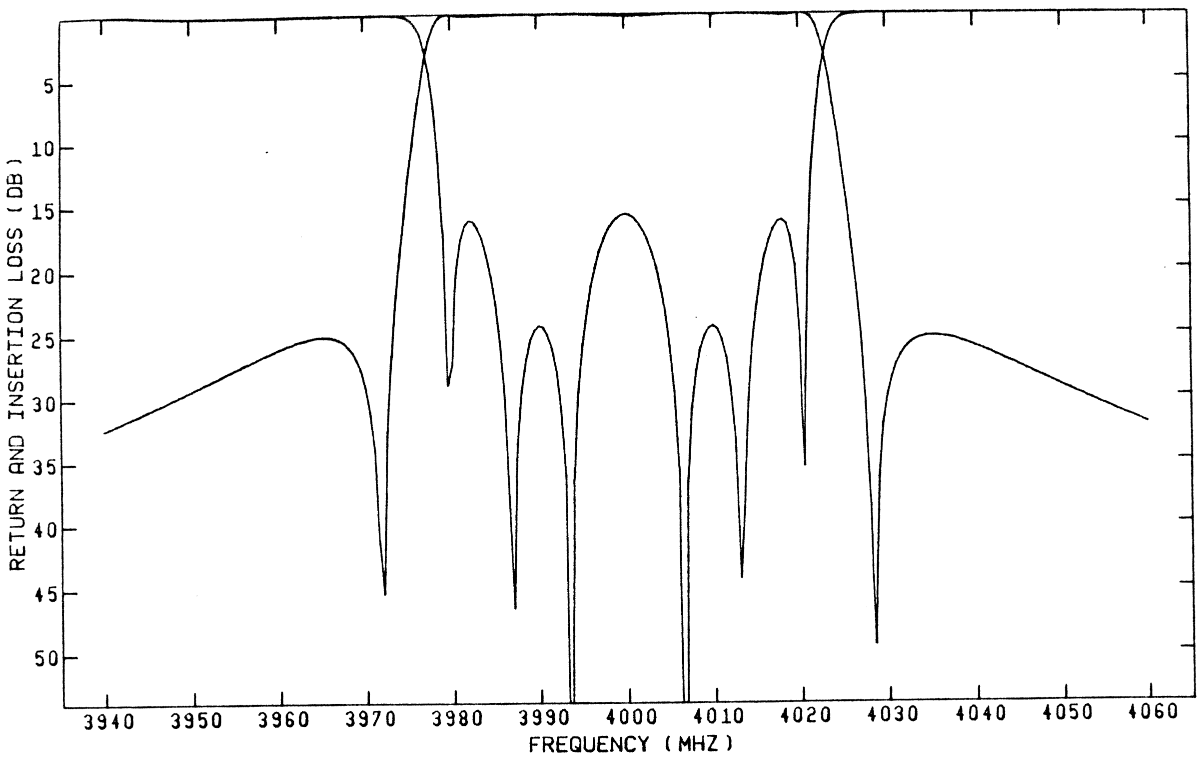

Fig. 13 Return loss and insertion loss of Design 5.

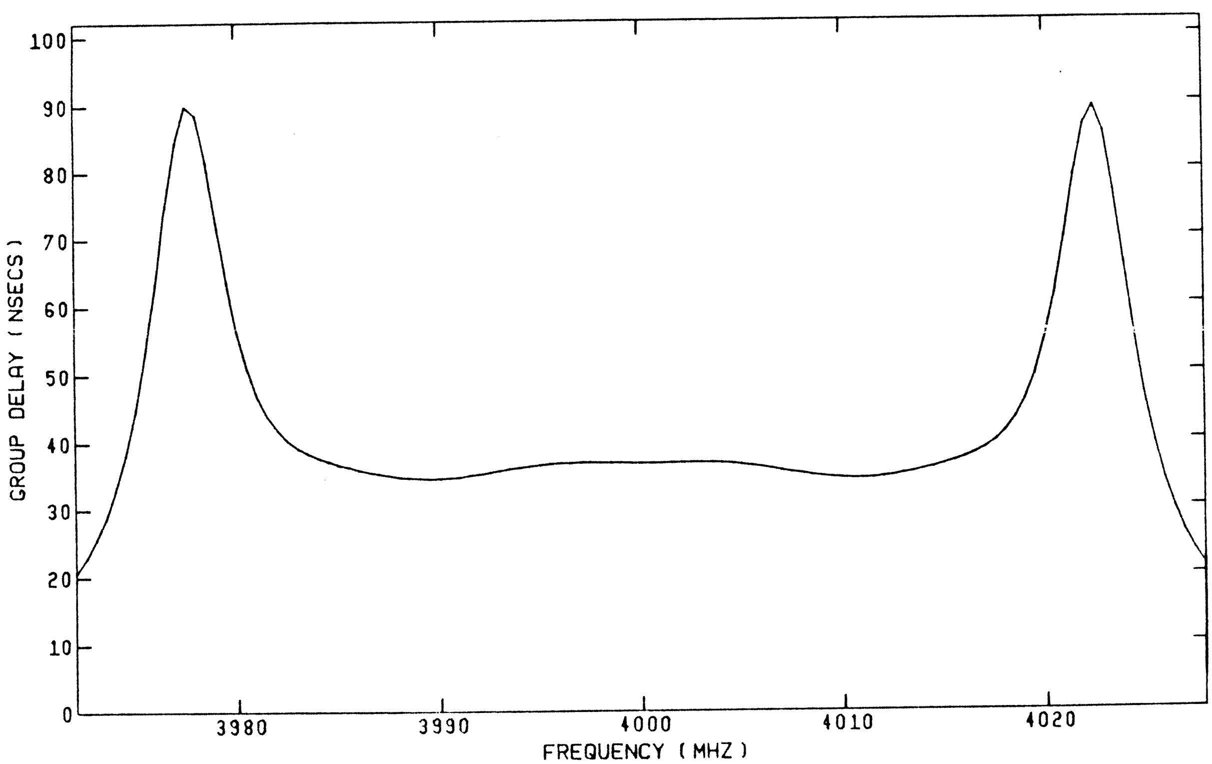

Fig. 14 Group delay of Design 5.

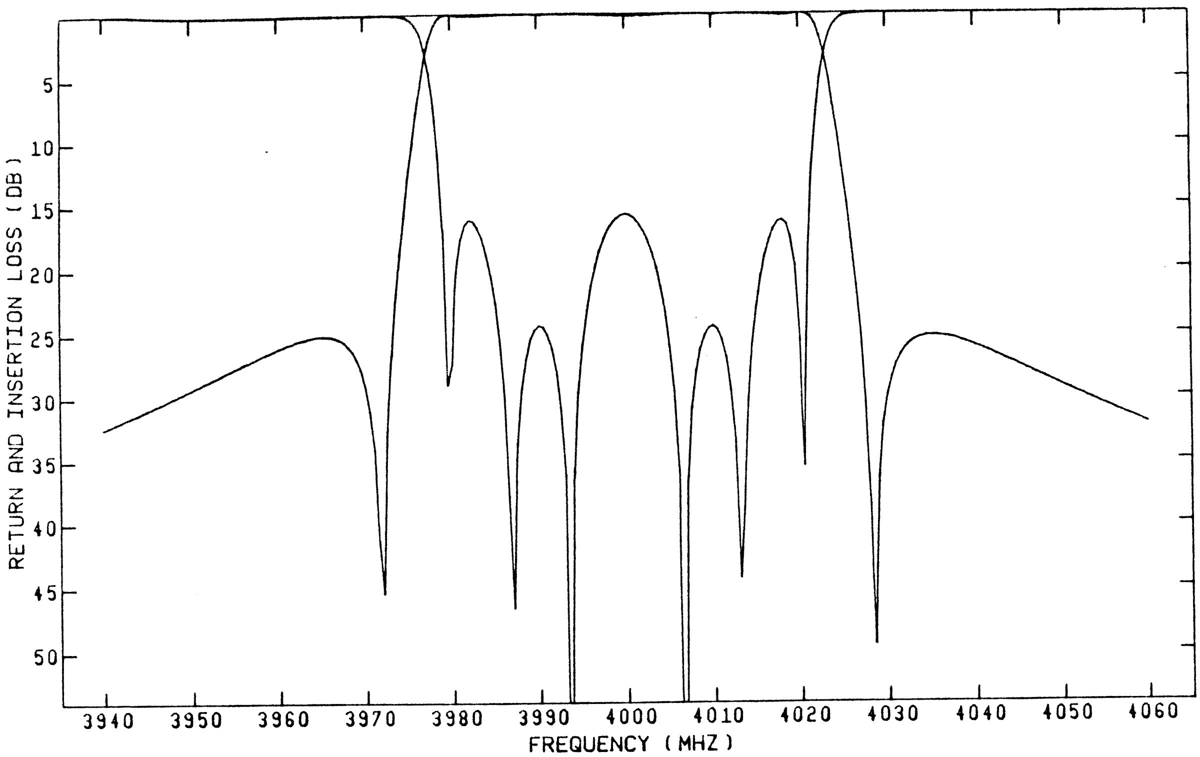

Fig. 13 Return loss and insertion loss of Design 5.

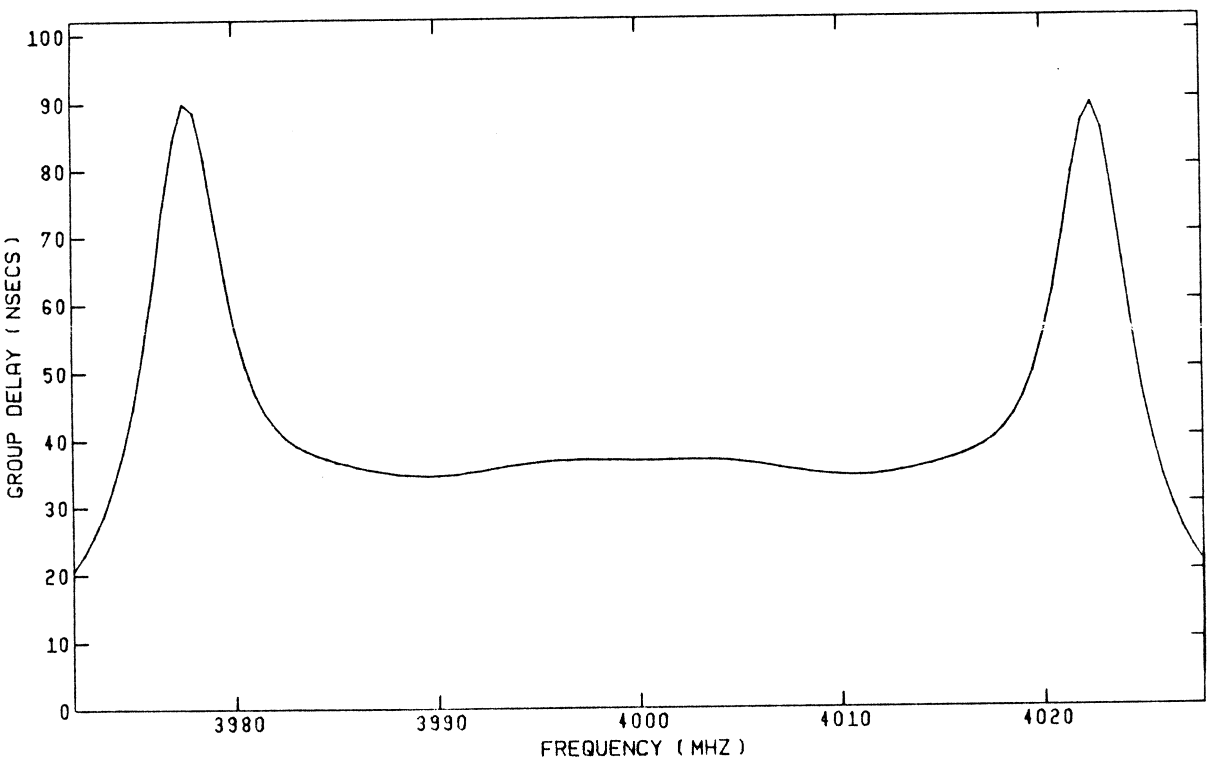

Fig. 14 Group delay of Design 5.

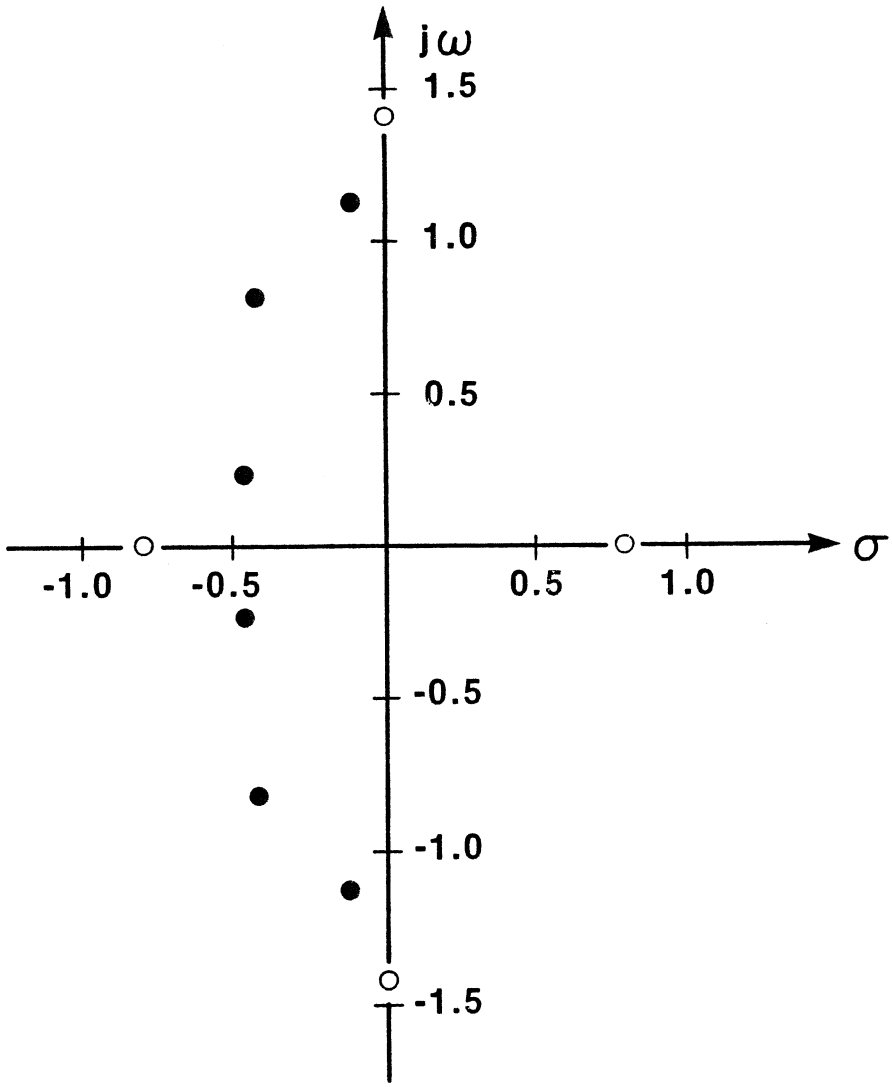

Fig. 15 Pole-zero pattern of the transfer function of Design 5.
● – poles, o – zeros.