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THE SIMPLIFIED REAL FREQUENCY METHOD APPLIED TO THE ACTIVE FILTERS SYNTHESIS

E. EL HENDAOUI, A. PERENNEC, P. JARRY

FACULTE DES SCIENCES ET TECHNIQUES - LABORATOIRE D'ELECTRONIQUE
ET SYSTEMES DES TELECOMMUNICATIONS - LEST - URA CNRS n° 1329 -
6, avenue Le Gorgeu - 29287 BREST CEDEX - FRANCE

ABSTRACT

In this paper, we present a direct method for the design of active filters using the simplified real frequency technique. This method bypasses the analytic theory and yields the maximum flat transducer power gain (T.P.G) possible. An example for low-pass filters is given at the end.

INTRODUCTION

Microwave filters have been widely utilised in Telecommunications systems, radars and signal processing. These filters were often built in passive networks (waveguide, transmissionlines). The idea of active filters is not new, since it was introduced in the thirties. However, early configurations were very cumbersome, since vaccum tubes were used as the gain elements. Bulky size, cost, high power consumption have seriously limited their use. But with the invention of the cheap transistors, and the development of integrated circuits technology, all these reasons have promoted new interest in Active Filters. Recently, many studies have been carried out on this subject. In this paper, we present a computer-aided design procedure (C.A.D), which is a simplified real frequency technique. We show how the simple formalism of this method allows us, without complex calculations, to optimise many characteristies of microwave active filters.

SIMPLIFIED REAL FREQUENCY METHOD

The simplified real frequency method was introduced by H. J. CARLIN and B. S. YARMAN [1], to design Broad-band Multistage Microwave Amplifiers. It was also applied for double matching problems. Its efficacity resides in its simplicity of use. No equivalent schema of transistors is necessary, it utilizes only the measured scattering parameters [S], of the FET devices, and the nonunilateral behaviour of the transistors is taken into account. The optimisation is based on a modified MARQUARD routine for least squares whitch is descrebed by J. J. MORE [2]. It has been shown [3] that the scattering parameters of an equaliser, E, can be completely determined from the numerator h(p) of the input reflection $e_{11}(p)$. E is assumed to be a ladder network and lossless, the other parameters are given as :

$$e_{11} = h(p)/g(p) \;;\; e_{21}=f(p)/g(p) \;;\; e_{22}= \pm h(-p)/g(p)$$

the sign of e_{22} depends on the parity of f(p), and is positive if f(p) is even.

The equalisers E up to the present have all been assumed to be a minimum phase structure with transmission zero only at $\omega = \infty$, $\omega = 0$ (i . e, $f(p) = p^k$) [4]. In our case we consider the general form of the f(p) given below :

$$f(p) = p^k \prod_{i=1}^{i=m} (p^2 + \omega_i^2)$$

$$g(p) = g_0 + g_1 p + g_2 p^2 + \dots + g_n p^n$$

$$h(p) = h_0 + h_1 p + h_2 p^2 + \dots + h_n p^n$$

n : presents the total number of reactive elements
k : presents the number of the attenuation poles at origin
m : presents the finite attenuation poles

The filter will be :

Low-pass if k = 0 ($h_0 = 0$)

Band-pass if $1 < k + 2m < n$

High-pass if $k + 2m = n$

The coefficients of the polynomial h(p) are chosen as unknown and the polynomial g(p) is generated from the Hurwitz factorization of :

$$g(p)\, g(-p) = h(p)\, h(-p) + f(p)\, f(-p)$$

since the matching network is lossless.

The optimisation is performed simultaneously on the transducer power gain and the VSWR.

Refering to (fig. 1) the transducer power gain (T.P.G) is given by :

$$T_K(\omega) = T_{(K-1)} \frac{|e_{21K}|^2 |S_{21K}|^2}{|1-e_{11K} S_{GK}|^2 |1-\hat{e}_{22K} S_{11K}|^2}$$

i.e $T_k(\omega) = T_{(k-1)} . E(\omega)$

where $T_{(k-1)}$: is the TPG of the first (k-1) stages with normalised resistive terminations

$(e_{ij})_k$: scattering parameters of the kth equaliser E_k

$(S_{ij})_k$: scattering parameters of the kth FET

$\hat{e}_{22k}$: reflection coefficient measured at the output of E_k to the left

S_{Gk} : reflection coefficient measured at the output of F_{k-1} to the left

then :

$$\hat{e}_{22k} = e_{22k} + \frac{e_{21k}^2 S_{GK}}{1-e_{11k} S_{GK}}$$

$$S_{GK} = S_{22(k-1)} + \frac{S_{21(k-1)} S_{12(k-1)} \hat{e}_{22(k-1)}}{1 - S_{11(k-1)} \hat{e}_{22(k-1)}}$$

The overall transducer power gain $T(\omega)$ is defined after the back end equaliser E_{k+1} has been added :

$$T(\omega) = (T_1 . T_2 \dots T_k)\, E_{(k+1)}(\omega)$$

and the VSWR is given by :

$$R_{ink} = \frac{1 + |\hat{e}_{11\,1}|}{1 - |\hat{e}_{11\,1}|}$$

R_{ink} is computed recursively from the load by :

$$S_{lk} = S_{11k}$$

$$\hat{e}_{11k} = e_{11k} + \frac{e^2_{21k} S_{lk}}{1 - e_{22k} S_{lk}}$$

$$S_{l(k-1)} = S_{11(k-1)} + \frac{S_{12(k-1)} S_{21(k-1)} \hat{e}_{11k}}{1 - S_{22(k-1)} \hat{e}_{11k}}$$

Where

$\hat{e}_{11k}$: is the reflection coefficient measured at the input of E_k to the right

S_{lk} : is the reflection coefficient measured at the input of F_k to the right

Finally with the back-end equaliser E_{k+1} we get the overall input VSWR (R_{in}) :

$$R_{in} = R_{in(k+1)} = \frac{1 + |\hat{e}_{11\ 1}|}{1 - |\hat{e}_{11\ 1}|}$$

Due to the technique optimization the output VSWR is only defined after the back-end equaliser E_{k+1} has been added

$$R_{out} = \frac{1 + |\hat{e}_{22(k+1)}|}{1 - |\hat{e}_{22(k+1)}|}$$

The objective is to obtain the flattest and highest gain possible, the minimum VSWR. This objective is achieved by minimizing the objective function, with a least LEVENBERG MARQUARD algorithm. This can be written as :

$$E^2 = \sum_{j=1}^{N} w1 \left(\frac{T(\omega_j)}{T0} - 1\right)^2 + w2 \left(\frac{R_{in}}{R0} - 1\right)^2$$

Where T0, R0 are the desired gain and input VSWR respectively. N is the number of sampling frequencies over the bandpass, W1, W2 are weighting functions. At each iteration, coefficients ($h_i + \Delta h_i$) are corrected using a least LEVENBERG MARQUARD algorithm based on the procedure of J. J. MORE [2] : Vector Δh is given as :

$$\Delta h = - [J^T J + \alpha D^T D]^{-1} J^T e_0$$

where :

- e_o : is the error vector
- J : The Jacobien matrix of e (i.e elements are $\partial e_j / \partial h_i$; j = 1 ... N ; i = 1...n)
- D : a diagonal matrix with takes into account the scaling of the problem
- α : The Levenberg-marquard parameter.

J. MORE introduced a relationship between J, D, α which permit rapid convergence, and this without initial guess of the vector h.

RESULTS :

The illustration of an example of an active filter over the band 1-5 GHz, with a transmission zero at 5.6 GHz is given (fig. 2), the gain is about 5.8 db and the output and input VSWR varies from 1.1 to 2.2.

The low-pass L-C elements are presented (fig. 3). The synthesis is done with the procedure proposed by NEIRYNCK, and BASTELAER [5]. The poles at origin and at infinity, are extracted, as usual, by a single capacitance or inductance, which can either in shunt or series according to the sign of h_n/g_n if k = 0, or h_o/g_o if not. Where as the finite attenuation poles are realised by a resonant (antiresonant) circuit located in the shunt (series) arms. The extraction of these arms is made possible by a shifting capacitance in the series (shunt) arm, chosen in such a way that the input impedance Z_e, after extraction of this capacitance exhibits a zero (pole) at the frequency of the attenuation pole considered. Z_e is given by :

$$Z_e = \frac{g(p) + h(p)}{g(p) - h(p)}$$

The FET is embedded in network including a parallel feedback loop, a drain series industance and a gate shunt resistance, for to reduce the imput reflection coefficient and stabilise the transistors. The measured scattering parameters of the transistor was given in table I.

N = 5 for E1 and n = 7 for E2. The values of the elements are as follows :

E1	E2
c1 = 0.1833 pF	c1 = 0.0594 pF
c2 = 0.9373 pF	c2 = 0.6895 pF
c3 = 0.5721 pF	c3 = 0.5236 pF
c4 = 0.1328 pF	c4 = 0.7907 pF
L1 = 0.8617 nH	c5 = 0.4726 pF
L2 = 4.8112 nH	L1 = 1.1713 nH
	L2 = 4.7896 nH
	L3 = 3.3377 nH

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TABLE I

Measured Scattering Parameters (S) of the transistor:

Freq	S11		S21		S12		S22	
GHz	MAG	DEG	MAG	DEG	MAG	DEG	MAG	DEG
1	0.9984	-9.97	2.602	172.00	0.0096	83.87	0.853	-3.33
2	0.9938	-19.78	2.567	164.11	0.0190	77.85	0 849	-6.61
3	0.9867	-29.29	2.511	156.41	0.0278	72.01	0.843	-9.79
4	0.9776	-38.39	2.438	148.98	0.0360	66.45	0.836	-12.85
5	0 9673	-46.98	2.354	141.87	0.0434	61.20	0.828	-15.76
6	0.9563	-55.03	2.261	135 11	0 0500	56 29	0 819	-18.51
7	0 9452	-62.51	2.165	128.70	0.0558	51.74	0.811	-21.13
8	0 9345	-69.41	2.068	122.64	0.0608	47.53	0.802	-23.6

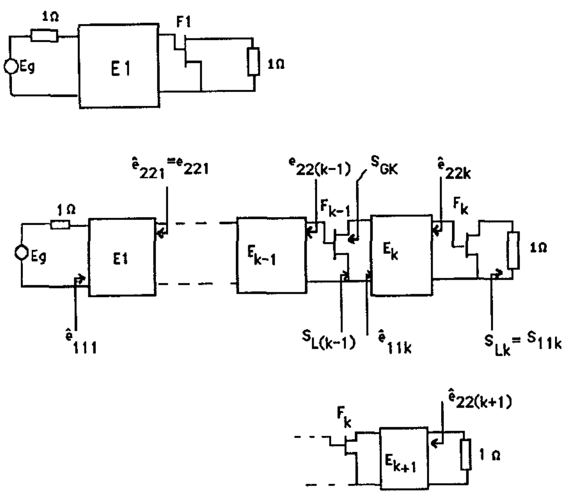

Fig.1

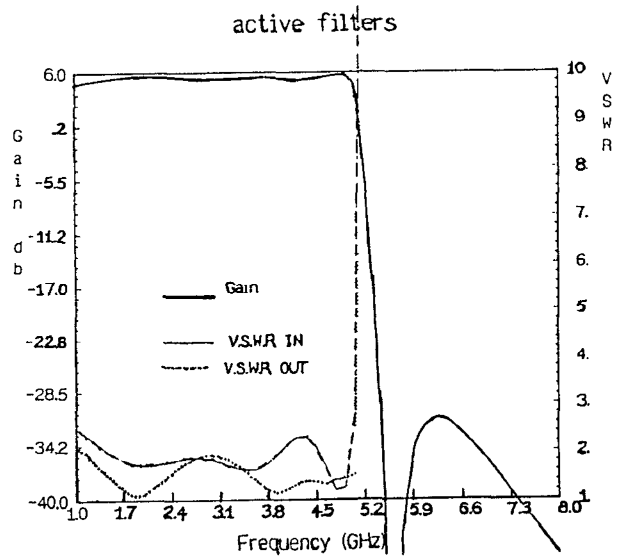

Fig.2

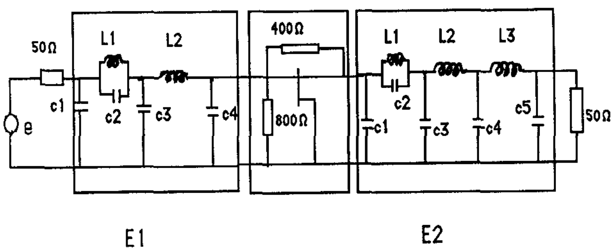

Fig.3