CONTIGUOUS BROADBAND MATCHING OF MULTIPLE RESONANT LOADS

J.D. Rhodes and M.J. Thornton*

ABSTRACT

The design theory for combining a number of loaded resonators to yield a prescribed return loss characteristic is presented with explicit formulae for the maximally flat case. Prototype element values are given for the equiripple case and the design of a broadband multi-channel frequency discriminator is used to demonstrate an application. Other applications include antenna arrays and broadband microwave amplifiers.

INTRODUCTION

The design of broadband microwave systems for applications such as ESM is often hampered by the fundamental constraint that the gain-bandwidth product of many microwave components and devices are not compatible with the required system performance. An example of such a component is the detector diode. Here a typical device operating through Ku band would exhibit an RC product of 0.15 ns giving a gain-bandwidth product of only 4 GHz. In practice for broadband detectors this can lead to a loss in sensitivity of as much as 10 dB compared with narrowband detectors, including matching circuit losses.

This paper describes the design theory for combining a number of relatively narrowband resonant load sections to yield a broadband response, whilst maintaining a prescribed return loss characteristic. The general form required for the reflection coefficient is given and the prototype element values for the maximally-flat case are derived. Results for the equiripple case are illustrated both for the lumped prototype circuit and a realissation including distributed components.

The general concept of optimally matching a number of loaded resonators is directly applicable to a range of systems including multi-element antenna arrays and broadband microwave amplifiers in addition to a new type of integrated frequency discriminator detector system.

THEORY

The prototype circuit as shown in Fig. 1 is assumed, where the frequency invariant reactances jB_r tune out the capacitances C_r at frequencies Ω_r. We wish to derive the impedance function Z(p) of the form

$$Z(p) = \sum_{r=1}^{n} \frac{1}{C_r p + jB_r + G_r} \qquad (1)$$

such that the reflection coefficient $S_{11}(p)$ approximates zero in some prescribed manner over a region on the imaginary axis, $p = j\omega$.

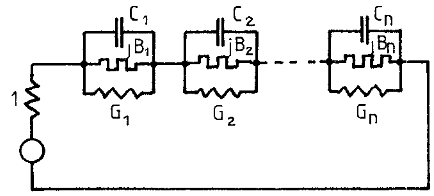


Fig. 1 Prototype circuit

Consider $S_{11}(p)$ of the form

$$S_{11}(p) = \frac{N(p)}{N(p+x)} \qquad (2)$$

where x is real and N(p) is an even or odd polynomial with all j-axis zeros, thus N(p+x) is Hurwitz for x>0. The corresponding impedance function is

$$Z(z) = \frac{N(z + x/2) + N(z - x/2)}{N(z + x/2) - N(z - x/2)}$$

where $z = p + x/2$

$$\frac{N(z + x/2) + (-1)^n N(-z + x/2)}{N(z + x/2) - (-1)^n N(-z + x/2)} \qquad (3)$$

Now, since N(z + x/2) is Hurwitz Z(z) is a reactance function

$$Z(z) = \sum_{r=1}^{n} \frac{A_r}{z - jZ_r} \qquad (4)$$

where A_r are positive and real. Hence

$$Z(p) = \sum_{r=1}^{n} \frac{A_r}{p + x/2 - jZ_r} \qquad (5)$$

MAXIMALLY FLAT CASE

Here $S_{11}(p) = \dfrac{p^n}{(1+p)^n}$ (6)

where the value of x has been normalised to

Professor J.D. Rhodes and Dr. M.J. Thornton are with Filtronic Components Ltd., Kildare Terrace, Leeds LS12 1DB, England.

unity. Now

$$Z(p) = \frac{(1+p)^n + p^n}{(1+p)^n - p^n} \qquad (7)$$

Hence the poles and residues may be found and the prototype element values may be written

$$C_r = 2n \sin^2[(2r-1)\pi/2n]$$

$$G_r = C_r/2 \qquad (8)$$

$$B_r = \frac{n}{2} \sin[(2r-1)\pi/n]$$

EQUIRIPPLE CASE

For this case the magnitude of the reflection coefficient takes the form

$$|S_{11}|^2 = \prod_{r=1}^{n} \frac{(\omega - \omega_r)^2}{1 + (\omega - \omega_r)^2} \qquad (9)$$

and the zeros may be found numerically such that $|S_{11}|$ is equiripple. The prototype element values may then be derived as in the maximally flat case by forming the input impedance, factorizing and extracting complex poles to give the residues, A_r, as in eqtn. (5). The element values G_r, C_r and jB_r are tabulated in Table 1 for the case $n = 4$, with max return loss 10, 15 and 20 dB. The prototype circuit bandwidth for these results is from $\omega = -1$ to $+1$.

	r = 1	2	3	4
RETURN LOSS 10 dB				
jB_r	-4.37	-2.03	2.03	4.37
C_r	4.58	6.47	6.47	4.58
G_r	.854	1.21	1.21	.854
RETURN LOSS 15 dB				
jB	-3.17	-1.76	1.76	3.17
C	2.97	5.20	5.20	2.97
G	.786	1.38	1.38	.786
RETURN LOSS 20 dB				
jB	-2.53	-1.65	1.65	2.53
C	2.08	4.51	4.51	2.08
G	.731	1.58	1.58	.731

TABLE 1 Element values to give equiripple return loss for n = 4.

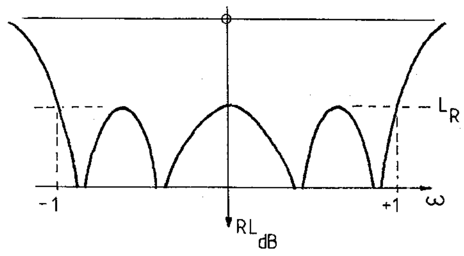


Fig. 2 Normalised, Prototype return loss for equiripple designs

CIRCUIT REALISATIONS

A physical realisation of the basic prototype is shown in Fig. 3, and is obtained from eqtn. (3), by extracting alternate phase shift sections and complex poles, (finishing with a short or open circuit) and then scaling each section to the desired load impedance. Cascading the parallel coupled line sections approximates the required series resonators, separated by phase shift sections for bandwidths in the region of an octave or less.

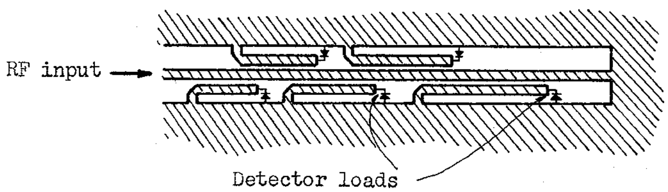


Fig. 3 Parallel coupled line realisation

An alternative realisation useful for wider bandwidths is shown in Fig. 4. Here the dual of eqtn. (3) is used and the shunt resonators are formed by short lines capacitively coupled at a common feed, and resonating the load capacitance. The circuit values of each resonant section are obtained by equating the conductance and slope parameters to those of the prototype circuit at the resonant frequency of each section.

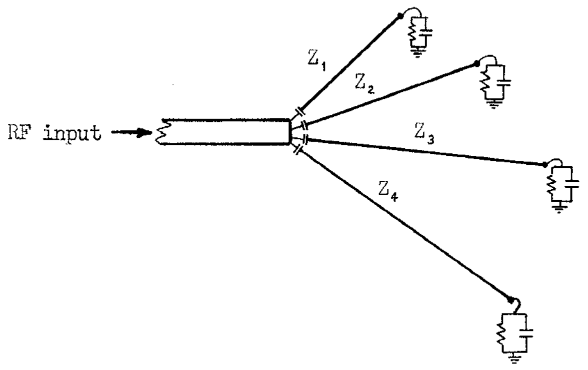


Fig. 4 Broadband circuit realisation

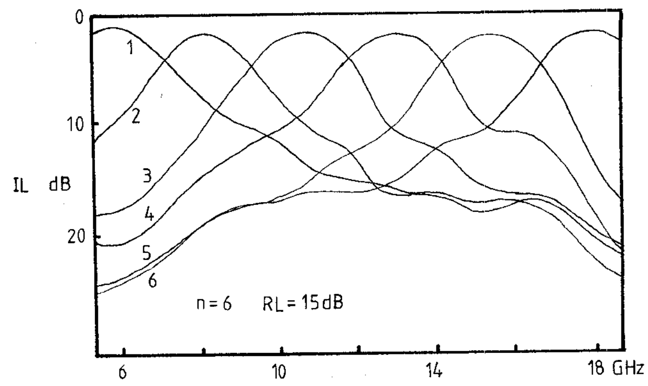
Fig. 5 shows the computed response of such a circuit for a 6-18 GHz 6 channel design.

APPLICATIONS

This technique is applicable to any microwave device where a reactive component may be tuned out to produce a loaded resonator, or where a resonant structure already exists, such as detectors and dipole antennae respectively. In the case of the detectors the theory may be used to match a number of detector diodes over a broadband, generating contiguous frequency discriminator characteristics as shown in Fig. 6. This technique is being used to develop a low cost IFM unit which has several advantages over conventional delay-line discriminators - high sensitivity, wide band (both RF and video), multiple signal capability and system flexibility.

The design theory is directly applicable to multi-element broadband antennae. Here, each element, which may be a dipole or resonant stub, is a natural resonant section with the radiation forming the resistive component. For this application, mutual coupling between elements may be accounted for by adjustment of the phase separators between the elements.

Another application of this work is in the design of broadband microwave amplifiers. The gain-bandwidth product limitation of GaAs FET devices may be avoided, not by cascading many lower gain stages but by contiguously combining a number of narrowband high gain stages at their input and output ports.



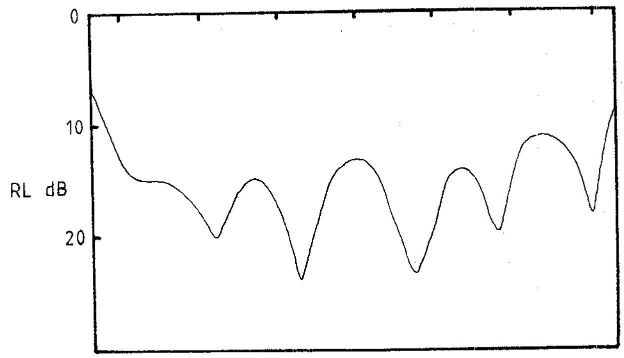


Fig. 5 Insertion Loss and Return Loss of circuit shown in fig. 4 for n = 6, RL = 15 dB

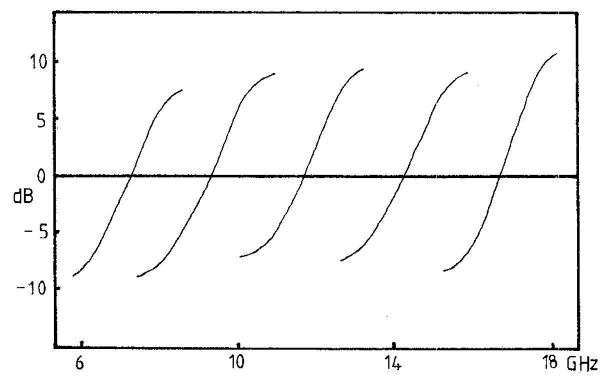


Fig. 6 Discriminator Characteristic formed by Insertion Loss Characteristic of fig. 5.