NON-MINIMUM-PHASE MICROWAVE FILTERS

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Summary. Ladder networks are particularly suitable for waveguide realizations. Conventional waveguide filters are designed by means of a transformation of lossless ladder networks terminated into resistances. This configuration, however, only allows the realization of a certain class of minimum-phase transfer functions. In this paper a type of microwave filter is analyzed, in which a reactive ladder is combined in a special way with a magic-T or 90-degree hybrid. This filter type can realize non-minimum-phase functions, which can offer very useful combinations of amplitude and phase responses. The insertion loss of the ladder network does not enter the transfer function in the same way as in ordinary waveguide filters. A synthesis procedure is given, and this has been applied in the construction of several microwave filters. The inclusion of non-reciprocity by means of ferrite devices to realize fixed or variable filters or microwave modulators has also been studied. A similar method is applied in the realization of driving point impedances at microwave frequencies, in order to realize such functions as cannot normally be realized by means of ladder structures.

Introduction. In non-minimum-phase networks, transmission zeros are permitted in the right half of the complex frequency plane, and on the imaginary axis, as well as in the left half-plane. These networks can offer far better combinations of amplitude and phase characteristics than minimum-phase networks with the same number of reactive elements (ref. 1). This is of prime importance in communication systems with frequency modulated signals, where a constant delay, in combination with a sharp amplitude cut-off, is required. Conventional waveguide filter structures do not, however,permit the realization of non-minimum-phase transfer functions. Waveguide filters are usually designed by means of a transformation of the lossless low-pass ladder with single branch elements (ref. 3). This configuration is particularly suitable for waveguide realizations, but allows only the realization of a restricted class of transfer functions.

Section I. Reciprocal networks. Non-minimum-phase functions can be realized by means of lattice networks. It is known that a "magic T" which is terminated in two of its ports by the impedances Z_A and Z_B is equivalent to a symmetrical lattice with the branch elements Z_A and Z_B. Such a realization is, however, non-canonical, as twice as many reactive elements as the degree of transfer function are usually required. In addition the resulting filter is very sensitive to variations in the reactive elements. If, however, a lossless ladder network is connected between two ports of a "magic-T" or 90-degree hybrid as in Fig. 1, these realizations can lead to canonical networks with reduced sensitivity, as the signals from the two ports are transmitted through the same reactive network. (These hybrid junctions have previously been used in microwave filters but in quite different ways, for exemple in connection with filters utilizing the cut-off effect, and in channel-dropping filters.)(ref.s 3 and 4)

The transfer function is formed by the reflected signals at the two ports, and the insertion loss of the reactive structure does not enter the transfer function in the same way as in ordinary filters. An analysis of the filter arrangement, with a "magic-T", and with and without a $\lambda/4$-transformer, leads to the transfer functions which are given in Fig. 2. The combination of a "magic-T" and $\lambda/4$-transformer is equivalent to a 90 degree hybrid, apart from an additional phase shift in the total transfer function. The transfer functions in Fig. 2 are expressed in terms of the elements of the z-matrix of the reactive network, and in terms of the even parts (M_1 and M_2), and odd parts (N_1 and N_2), of the input impedance $(M_1 + N_1)/(M_2 + N_2)$ of the resistively terminated reactive network as in the Darlington synthesis procedure. (ref. 5).
(Even and odd parts are interchanged in Fig.s 2 and 3 for networks corresponding to case B in the Darlington procedure).
The numerator of the transfer function is either an even or an odd function. This comprises an important class of non-minimum-phase transfer functions as is shown in ref. 1.

The synthesis procedure is uncomplicated, as neither the ensignant E = $(= M_1M_2 - N_1N_2)$ nor the reflection coefficient have to be evaluated, which is in contrast to the Darlington procedure. The terms of the odd parts N_1 and N_2 are determined, term by term, in a continued fraction expansion of the input impedance. The terms of the even parts, M_1 and M_2, are given by the desired transfer function H by identifying the even and odd parts $(H = m_1/(m_2 + n_2)$ or $H = n_1/(m_2 + n_2)$, where the m:s are even parts, and the n:s odd parts). Only the sum of the odd parts N_1 and N_2 is specified by H, and by making use of this degree of choice the terms of the odd parts are chosen, in the expansion, in such a way, that, the reactive network is a simple ladder structure. This can be achieved by ensuring that two terms drop out in every step up to the next to the last one in the continued fraction expansion. The number of elements in the resulting ladder network is determined by the degree of the denominator of the transfer function, and the transmission zeros do not require any additional reactive elements. In higher order filters certain restrictions are imposed on the numerators. Corresponding waveguide cavities, with irises or pins as reactive discontinuities, are then calculated from the ladder network using conventional methods (ref. 2).

This method has been applied in the construction of several filters. Two filters with centre frequencies below the microwave range (30 MHz and 7 MHz) were built mainly for the purpose of experimental verification of the method. The 30 MHz filter was constructed for use with a 90 degree hybrid (Merrimac QH-1-20), and the 7 MHz filter for use with a differential transformer which is equivalent to a "magic-T". Close correspondence between measured and calculated values was achieved, both in these filters, and in a number of X-band waveguide filters of lower order which have also been developed. An X-band filter of degree five is at present under construction. These filters can be made very compact by the use of folded "magic-T:s".

Section II. Non-reciprocal networks. Anti-reciprocal phase shifters can be realized at microwave frequencies by means of ferrite elements which are asymmetrically situated in the waveguide structure. As the ferrite elements are not situated in the middle of the waveguide where the electric field is strongest, the phase shifters can be made to handle high power. Furthermore the phase shift can be electronically controlled by varying the applied magnetic field. If an antireciprocal phase shifter is connected in cascade with the reactive network in the type of filters described in Section I, the transfer function is formed not only by the signals reflected by the network but also by the signals which are transmitted through the reactive network. Fig. 3 shows the transfer functions in this non-reciprocal case. The lower network can be used to realize transfer functions having numerators with both even and odd parts. In the upper network the transmission zeros do not appear in complex conjugate pairs with respect to the centre frequency. This type of function is further discussed in the following section.

If the total network only consists of a variable anti-reciprocal phase shifter and a hybrid, as in Fig. 4, a microwave modulator is obtained. Such a modulator can be made to handle high power, and this is in contrast to modulators utilizing the Faraday rotation effect.

Section III. Driving point impedances. In ladder networks the ensignant E has all its roots at infinity. This restriction can be avoided by making use of a ladder network connecting two ports of a microwave hybrid to realize driving point impedances. If the output port is connected to a matched load, the input impedance of the total network can be shown to be:

$Z = (2N_1+M_1+M_2+2\sqrt{E}\cos\varphi)/(2N_2+M_1+M_2-2\sqrt{E}\cos\varphi)$, or

$Z = (2N_1+N_2+2M_1+j2\sqrt{E}\cos\varphi)/(2N_1+N_2+2M_2-j2\sqrt{E}\cos\varphi)$, if a $\lambda/4$-transformer is included. (Even and odd parts may be exchanged corresponding to case B in the Darlington procedure). The first type of network can be used to realize driving point impedances in which the centre frequency is maintained, but where the bandwidth can be altered by varying the antireciprocal phase shift. In the second type of network, poles and zeros do not appear in complex conjugate pairs with respect to the centre frequency, (within the frequency range that a 90 degree hybrid, or a 90 degree phase shifter and the anti-reciprocal phase shift can be realized). In these networks the anti-reciprocal phase shift can be used to shift the centre frequency, and this opens the possibility of realizing filters in which the centre frequency can be altered by varying one non-reciprocal element, without detuning the cavities.

References.

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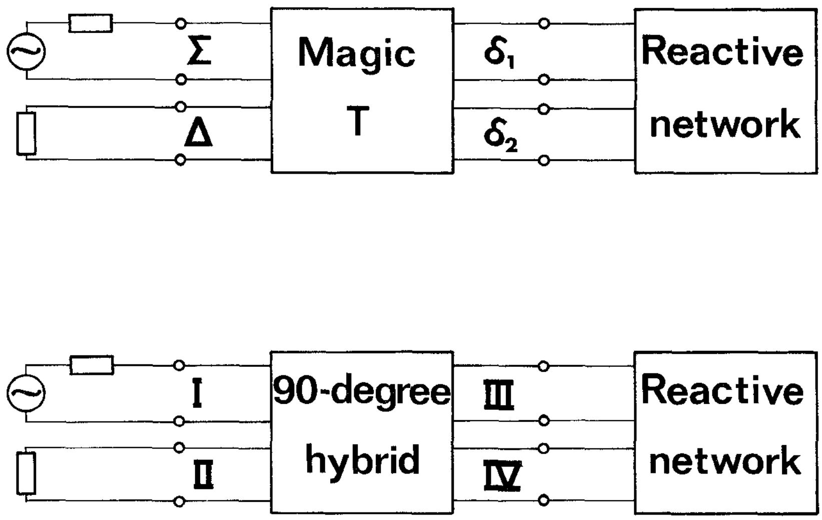
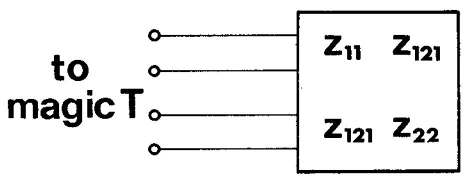
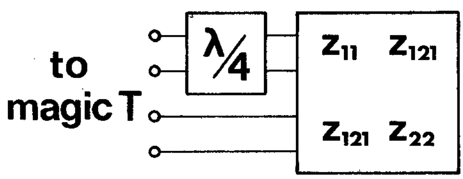


Fig. 1. Microwave filter arrangement.

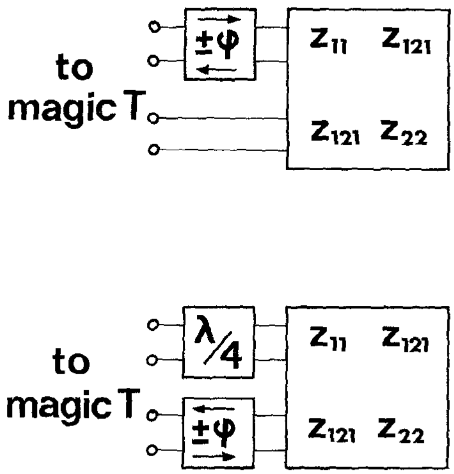


$$t=\frac{z_{22}-z_{11}}{\det z+1+z_{11}+z_{22}}=\frac{M_2-M_1}{M_1+M_2+N_1+N_2}$$



$$t=\frac{1-\det z}{\det z+1+z_{11}+z_{22}}=\frac{N_2-N_1}{M_1+M_2+N_1+N_2}$$

Fig. 2. Microwave filter transfer functions.



$$t=\frac{z_{22}-z_{11}+j2z_{121}\sin\varphi}{\det z+1+z_{11}+z_{22}}=\frac{M_2-M_1+j2\sqrt{E}\sin\varphi}{M_1+M_2+N_1+N_2}$$

$$t=\frac{1-\det z+2z_{121}\sin\varphi}{\det z+1+z_{11}+z_{22}}=\frac{N_2-N_1+2\sqrt{E}\sin\varphi}{M_1+M_2+N_1+N_2}$$

Fig. 3. Microwave filter transfer functions.

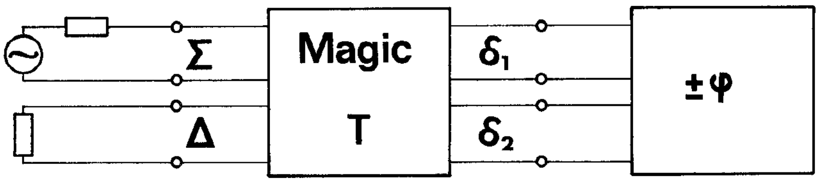


Fig. 4. Microwave modulator.