AN OPTIMAL LOW LOSS HF DIPLEXER USING HELICAL RESONATORS

Ralph Levy [1] and Konrad J. Andersen [2]

1. R. Levy Associates, La Jolla, CA 92037
2. RF Microsystems, Inc., San Diego, CA 92111

ABSTRACT

A diplexer comprising two narrow band closely spaced channels in the 164-175 MHz band having relatively low insertion loss, small size and light weight is described. The diplexer replaces a device with dimensions of 6 x 10 x 18 in. with one of dimensions 2.2 x 6 x 6 in. This order of magnitude reduction in size is achieved by the use of optimal asymmetric filters each having poles placed in the other passband giving high mutual isolation, combined with the use of helical resonators giving an unloaded Q of the order of 800, which is about twice as high as previous lumped element filters. Additional out-of band rejection is obtained by the unusual technique of introducing an additional attenuation pole in the output coupling.

INTRODUCTION

Filters and multiplexers used at HF (i.e. 30-300 MHz) are often very large due to low insertion loss requirements for closely spaced narrow band channels. Size reduction over cavity filters may be considered using helical resonators [1], but frequently these will not suffice if normal Chebyshev filters are used, since the degree of a filter is often such that the loss would be excessive. Finite attenuation poles must be included to give more optimum lower loss structures, but there appears to have been no work on helical resonator elliptic function type filters, i.e. lumped element elliptic filters where the resonator Q is much higher than the normal maximum of around 400.

A valid competitive technology which may be used for size reduction is by means of ceramic block combline filters [2]. The disadvantages here include high development costs, presently limited availability of manufacturing facilities, and comparatively large weight due to the heavy ceramic, which may also be expensive in material cost alone.

THEORY

The theory will be described by giving an example of a diplexer actually designed, constructed and tested. The specification was as follows:

Passbands: 166 $\pm$ 2 MHz and 173 $\pm$ 2 MHz

Passband insertion loss: 2.5 dB

Passband VSWR: 1.4:1 (band center)
2:1 (band edges)

Rejection: Mutual isolation > 45 dB
Rejection of either channel at 147 and 190 MHz > 30 dB

Chebyshev filters having poles all at zero and infinity of degree 6 would be required to meet this specification. The unloaded Q would need to be at least 1500, requiring distributed resonators with a ground plane spacing of 2.5 in. and electrical length of 45^{o}, giving a physical length of 8.9 in. Helical resonators having a minimum of 2 turns are unable to realize the desired Q. Single turn resonators might be used, and the resulting resonator would then be equivalent to a folded combline or hairpin resonator [3].

However the Q requirement may be reduced by a factor of 2 by using optimum generalized Chebyshev filter characteristics having poles at finite frequencies. The rejection is met using filters having 3 passband zeros and two finite frequency poles located within the opposite passband, giving the individual filter characteristics of Fig.1. These doubly-terminated filters may derived using FILSYN [4], or in our case using an independent synthesis program which derives generalized bandpass filters with arbitrary finite frequency poles.

The filters were synthesized initially without input and output couplings, as indicated in Fig 2. Admittance inverters were then incorporated at each end to

CH3141-9/92/0000-1187$01.00 © 1992 IEEE

1992 IEEE MTT-S Digest

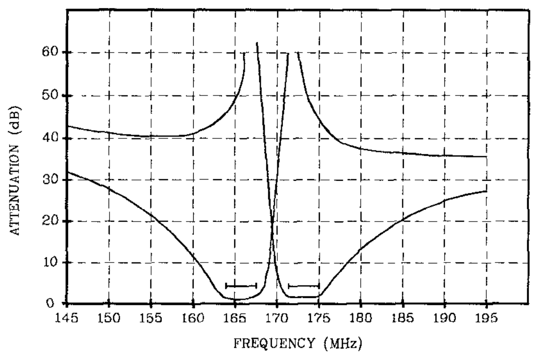


Fig. 1 Individual Filter Characteristics (not diplexed).

transform the internal admittance to one giving rather large inductors with minimal values for the shunt capacitors. The reasoning here is that it is desired to realize the inductors with helical resonators which are almost self-resonant, giving the maximum possible inductor unloaded Q.

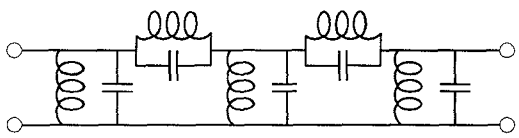


Fig. 2 Filter Topology before Addition of Coupling Inverters.

The equivalent circuit of a typical helical resonator is the Pi network shown in Fig. 3. In the case of a normal shunt helical resonator port 2 is grounded and a parallel LC resonant circuit is formed.

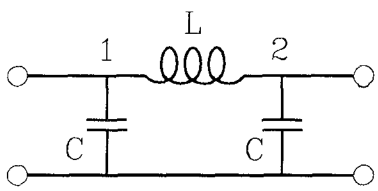


Fig. 3 Equivalent Circuit of a Helical Resonator.

When designing a filter of the type shown in Fig. 1 it is essential to include the "stray" capacitors of the helical resonators in the circuit, and these must be absorbed into the shunt capacitors.

The rejection of the high channel is < 30 dB at 190 MHz, necessitating the introducion of another pole of attenuation above this frequency. Since the filter already possesses the maximum number of such poles in the conventional sense, it was realized that the only convenient way to incorporate this pole was in the output coupling admittance inverter.

Approximate realizations of admittance inverters using simple elements having poles at zero or infinity, e.g. series C or series L with negative shunt elements, are well known [5]. It is also possible to use a more general admittance inverter where the attenuation pole is at a finite frequency rather than at zero or infinity [6]. The procedure for forming the coupling network with the pole is illustrated in Fig. 4. The required immittance is J, while the non-ideal realization is

$$J' = \omega C - 1/\omega L \qquad (1)$$

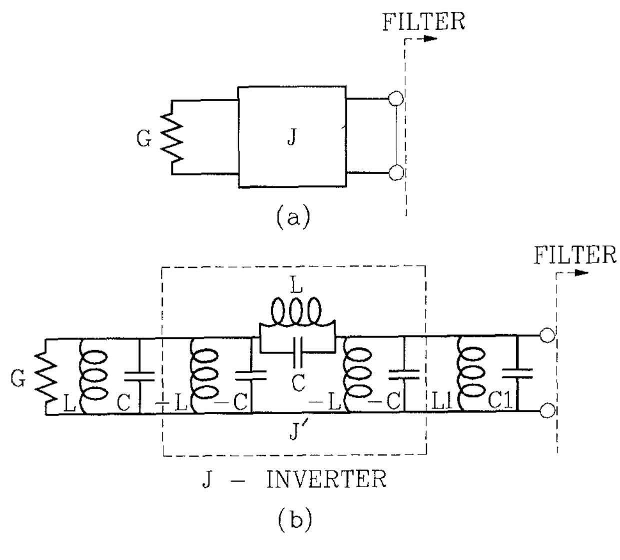


Fig. 4 Equivalence Between:
(a) Ideal Inverter Coupling, and
(b) Practical Pole-Forming Inverter Coupling.

which is frequency dependent. However for narrow band filters the variation across the filter bandwidth is acceptable.

The negative circuit elements of the inverter closer to the termination must be cancelled by introducing equal positive values next to the termination, as shown on the RHS of Fig. 4. The excess susceptance introduced is cancelled by adding the compensating susceptance J_1 to the side next to the filter. Equating admittances looking from the filter back towards the termination for the ideal and practical cases of Fig. 4 gives

$$J^2/G = J'^2/(G + jJ') + jJ_1 \tag{2}$$

Equating real and imaginary parts leads to the equations

$$J' = J / \sqrt{1 - J^2/G^2} \tag{3}$$

$$J_1 - J' = J \sqrt{1 - J^2/G^2} \tag{4}$$

The susceptance $J_1 - J'$ is to be formed by subtraction from the first shunt section of the filter.

The pole frequency is given by

$$\omega_p = 1 / \sqrt{LC} \tag{5}$$

which may be set to any desired frequency in the stopband, but not too close to the passband to avoid rapid change of the inverter impedance across the passband. In the present case the pole was set at 210 MHz with the passband edge at 176 MHz, which increased the rejection at 190 MHz by 7 dB.

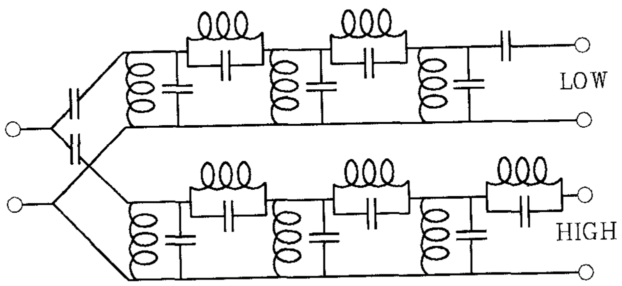


Fig. 5 Equivalent Circuit of the Diplexer

When the diplexer is formed by connecting the two filters in parallel, there is a degradation in input VSWR to > 2:1, but this may be rematched by adjusting the first few element values of each filter. This can be carried out using optimization, or as in the present case by computer tuning. The equivalent circuit of the diplexer is given in Fig 5.

The final theoretical characteristics of the diplexer are shown in Fig. 6, which also gives the measured performance.

PRACTICAL REALIZATION

The inductors, which have values in the range of 100 - 400 nH, were constructed of air coils using 18 gauge copper wire, which is sufficiently thick to prevent microphonics. The mean coil diameter was 0.75 in. with a 2.0 in. ground plane spacing to give an unloaded Q of 800 with self resonant frequencies at about 300 MHz, i.e. slightly higher than the operating frequency. This value of Q was derived from an accurate computer program using formulas derived by Macalpine and Schildknecht [7]. The approximate formula given in [1, p. 19], namely

$$Q = 44\, b \sqrt{f} \tag{6}$$

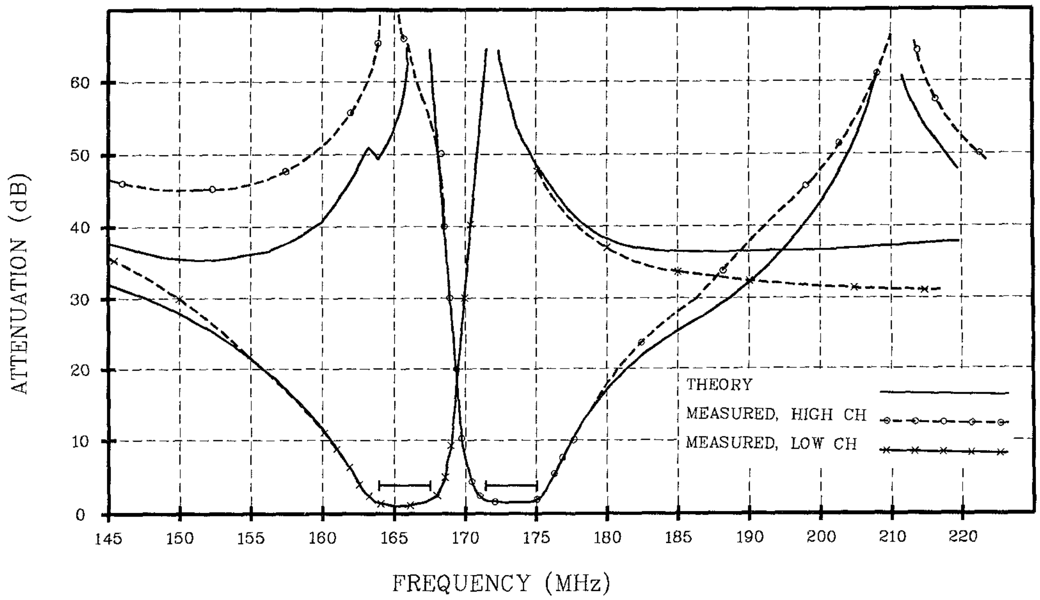
with b=2.54 cm (the shield spacing is 2 in.) and f=170 MHz gives Q = 1440 which is much too high. Actually (6) holds only for a ratio of shield spacing to inductor diameter of 2:1, wheras here the ratio is 2/0.75 = 2.67. Since the Q of an inductor is practically independent of the shield spacing unless it is very close to the coil (the field is confined mainly within the coil), it is more accurate to use an alternate formula

$$Q = 88\, a \sqrt{f} \tag{7}$$

where a is the mean coil radius in cm. anf f the frequency in MHz. Thus in the case of a 0.75 in. diameter coil at 170 MHz we have

$$Q = 88 \times 2.54 \times .75/2 \times 13.04 = 1093$$

This is still somewhat higher than the actual value of 800. The simple formula probably fails to take several practical effects into account, including lack of perfection in actual inductors.



In previous helical resonator filters inter-resonator coupling is usually electromagnetic by means of coupling apertures which are either predominately inductive or capacitive [1]. In the present case having elliptic-type filters the magnetic coupling is far too tight to be achievable by aperture coupling - the series inductors are of the same order of magnitude as the shunt inductors. Therefore the coupling inductances must be realized directly as actual helices. The coupling capacitors are in shunt across these to form the pole resonances.

The series and shunt capacitors may be constructed using any convenient dielectric material having low loss tangent. In our case the dielectric was Duroid 5880 (Rogers Corporation) which has a loss tangent quoted by the manufacturer as .0004 at 1 MHz, increasing to .0009 at 10 GHz.

The diplexer was constructed in brass and weighs 3 lbs, which would be much lower if aluminum were used.

The measured results in Fig. 6 agree reasonably well with theory, especially in the passbands. The stop band rejection differ somewhat from theory, explainably partly by the way the diplexer was tuned without regard to meeting predictions but rather to meeting the specifications.

CONCLUSIONS

A high frequency diplexer having closely spaced narrow bandwidth channels has been constructed having rather high (45dB) isolation between channels using optimal filters of degree 3 only. Novel features include the use of both series and shunt high Q helical resonators in an elliptic function type filter, giving exceptionally low loss in a reasonable volume, and the inclusion of an additional stop band pole incorporated in the output coupling network.

REFERENCES

[1] Peter Vizmuller, "Filters with helical and folded helical resonators," Artech House, Norwood, MA, 1987.

[2] F. B. McKee, "Microwave Filters," U.S. Patent No. 3,505,618, Filed May 4, 1967, Patented April 7 1970. (This is the first description of dielectric block filters, invented prior to the availability of temperature stable low loss ceramics).

[3] E. G. Cristal and S Frankel, "Hairpin-line and hybrid hairpin-line/half-wave parallel-coupled-line filters", IEEE Trans. on Micr. Theory and Tech., vol. MTT-20, pp. 719-728, November 1972.

[4] "Super Filsyn", DGS Associates, Santa Clara, CA.

[5] S. B. Cohn, "Direct-coupled resonator filters," Proc. IRE, vol.45, pp. 187-196, February 1957.

[6] R. Levy, "Mixed lumped and distributed linear phase filters," 1974 European Conf. on Circuit Theory and Design, IEE(London) Conf. publ. No. 116, pp.32-37, July 1974.

[7] W. W. Macalpine and R. O. Schildknecht "Coaxial Resonators with Helical Inner Conductor," Proc. IRE, vol. 47, pp. 2099-2105, December 1959.