

# HF Diplexer with Helical Resonators

*The authors introduce helical resonators in a 170 MHz radio diplexer. The lumped element approach is quite compact (2.2 x 6 x 6 in., less than 1/13 th the volume of the distributed model it replaces), has high isolation (45 dB) between channels spaced only 3 MHz apart, and relatively low loss (2.5 dB). An unusual technique of introducing an additional attenuation pole in the output coupling contributes to the isolation.*

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**F**ilters and multiplexers used in the high frequency band (HF) as well as those at the low microwave frequencies 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 Tchebyshev 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, that is, lumped element elliptic filters in which the resonator Q is much higher than the customary maximum Q of about 400.

This paper introduces the concept of helical filters and multiplexers having quasi elliptic function characteristics, and also demonstrates the advantage of using extra rejection poles realized in the input and output coupling networks. The techniques are illustrated by describing a diplexer consisting of two narrow band closely spaced channels in the 164-175 MHz band having relatively low insertion loss, small size and light weight. The diplexer measures 2.2 x 6 x 6 in., and replaced a

comblin diplexer of dimensions 6 x 10 x 18 in., more than 13 times the size of this helical diplexer.

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 850, about twice that customarily employed in previous elliptic function lumped element filters.

Extra out of band rejection is contributed through the unusual technique of introducing an additional attenuation pole in the output coupling. The technique has been applied to filters and diplexers in the 30-1600 MHz range, and is possibly extendable to higher frequencies.

A valid competitive technology which may be used for size reduction embodies ceramic block comblin filters [2]. However, their disadvantages include high development costs, presently limited availability of manufacturing facilities, and comparatively large weight due to the heavy ceramic, whose high cost might also mitigate against the approach.

### Theory

The theory is described through example of the actual diplexer designed, constructed and tested. The specifications are described in Table I.

**Table I. Specifications of the HF diplexer**

Passbands:	164-168 MHz and 171-175 Mhz
Max. Passband insertion loss:	2.5 dB
Max. Passband VSWR:	1.4 (band center) 2.0 (band edges)
Rejection: Min Mutual isolation:	45 dB
Min. Rejection of each channel at 147 and 190 MHz:	30 dB

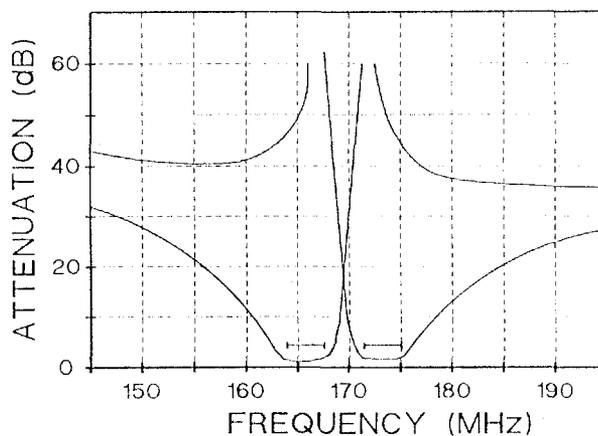
These specifications are typical for telemetry applications. The insertion loss may seem quite high, yet if the diplexer were realized using Tchebyshev filters of degree 6 the specifications could just be achieved by having poles all at zero and infinity. The minimum unloaded Q would be at least 1500, requiring distributed resonators with a ground plane spacing of 2.5 in. and electrical length of 45 deg., giving a physical length for each resonator of 8.9 in.

A ceramic block diplexer using a dielectric constant of 36 and employing optimal evanescent-mode filters could be constructed in dimensions of 3 x 3.5 x 10 in., or 2 x 2.8 x 8 in. using a dielectric constant of 80. These figures are for Tchebyshev

filters, and might be reduced somewhat if finite frequency poles were introduced by cross-coupling between non-adjacent resonators or other means. Such technology might indeed be useful, but the ceramic blocks are probably unavailable in such large sizes, and would be very expensive in material costs alone, as well as having considerable weight. Development is a specialized process of limited availability, with high initial development costs.

Helical resonators having a minimum of 2 turns cannot realize the desired  $Q > 1500$ . Single turn resonators might be used, and the resulting resonator would then be equivalent to a folded comblin or hairpin resonator [3], which could indeed be considered as a possible attractive alternative technique.

However the Q requirement may be reduced by a factor of 2 by using optimum generalized Tchebyshev 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 Figure 1. These doubly-terminated filters may be derived using FILSYN [4], or, as in our case, using an independent synthesis program which derives generalized bandpass filters with arbitrary finite frequency poles.



**Figure 1. Individual filter characteristics (not diplexed).**

The filters were synthesized initially without input and output couplings, as indicated in Figure 2. Admittance inverters were then incorporated at each end to transform the internal admittance to one giving rather large inductors with minimal values for the shunt capacitors. The reasoning was that it is desirable to realize the inductors with helical resonators which are almost self-resonant, giving the maximum possible inductor unloaded Q.

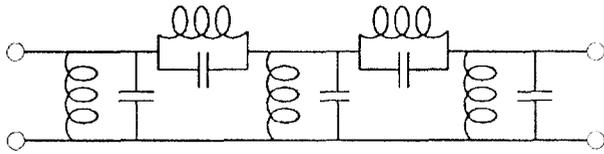


Figure 2. Filter topology before addition of coupling inverters.

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

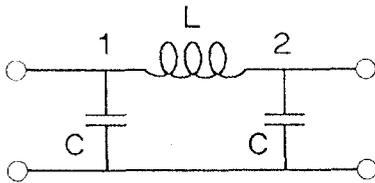


Figure 3. Equivalent circuit of a helical resonator.

When designing a filter of the type shown in Figure 1 it is essential to include the "stray" capacitors of the helical resonators in the circuit. These must be absorbed into the shunt capacitors. This implies that the helical resonators must be self-resonant at a frequency somewhat higher than the passbands, else there would be little or no shunt capacitance available for this; and the resonators would tune too low.

The rejection of the high channel is < 30 dB at 190 MHz, necessitating the introduction 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. The fact that this possibility exists contradicts the impression that elliptic-type filters of degree  $n$  may have only  $(n-1)$  finite frequency poles. This is true for direct-coupled filters, but if loose coupling by means of reactive coupling sections is employed, extra poles may be incorporated within these couplings.

Approximate realizations of admittance inverters using simple elements having poles at zero or infinity, for example, series C or series L with negative shunt elements, are well known [5]. It is also possible to use a more general admittance inverter for which 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 Figure 4. The required immittance is  $J$ , while the non-ideal realization is

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

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

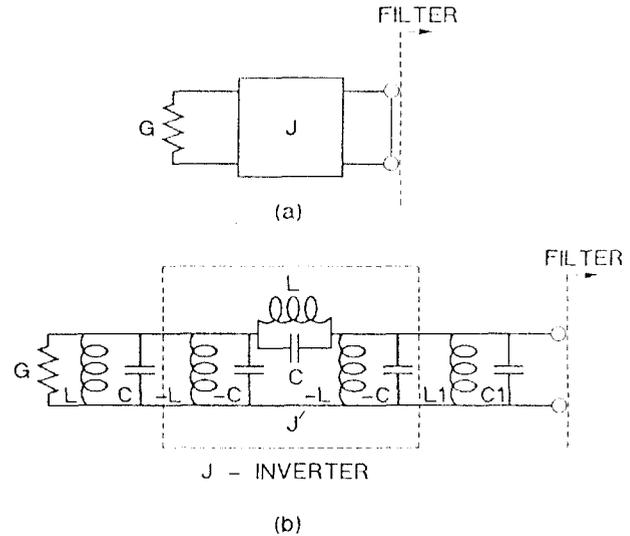


Figure 4. Equivalence between (a) Ideal inverter coupling, and (b) Practical pole-forming inverter coupling.

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

$$(2) \quad J^2/G = J'^2/(G + jJ') + jJ_1$$

Equating real and imaginary parts leads to the equations

$$(3) \quad J' = J / \sqrt{1 - J^2/G^2}$$

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

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

The pole frequency is given by

$$(5) \quad \omega_p = 1 / \sqrt{LC}$$

which may be set to any desired frequency in the stopband, but not so close to the passband as to cause a rapid change of the inverter impedance across the passband. The resulting mismatch may be compensated by re-optimization of the filter response. In the present case the pole was set at 210 MHz with the passband edge at 176 MHz, increasing the rejection at 190 MHz by 7 dB.

When the diplexer is formed by connecting the two filters in parallel, there is a degradation in input VSWR to  $> 2:1$ , because this corresponds to the parallel combination of two 50 ohm circuits. However, this may be rematched by adjusting the first few element values of each filter. This can be carried out using optimization, or as in this case by computer tuning. The resulting equivalent circuit of the diplexer is given in Figure 5.

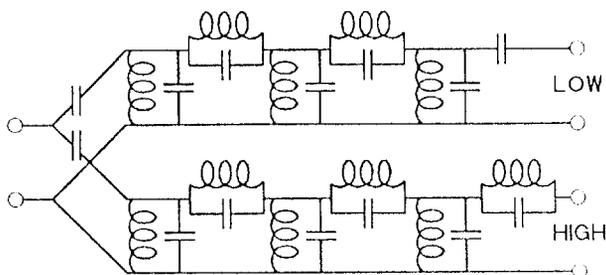


Figure 5. Equivalent circuit of the diplexer.

The final theoretical characteristics of the diplexer are shown in Figure 7, 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 rugged to prevent microphonics for most applications. In more severe vibration environments it would be possible to use low-loss formers, or to use a dielectric foam, the method most commonly employed for lumped-element filters. The mean coil diameter was 0.75 in. with a 2.0 in. ground plane spacing to

give an unloaded  $Q$  of over 1000 and self resonant frequencies at about 300 MHz, somewhat higher than the operating frequency. This value of  $Q$  was derived from an accurate computer program using formulas derived by Macalpine and Schildknecht [7]. An approximate formula frequently quoted in the literature, e.g. [1, p. 19], namely

$$(6) \quad Q = 44 b \sqrt{f}$$

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

$$(7) \quad Q = 88 a \sqrt{f}$$

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

$$(8) \quad Q = 88 \times 2.54 \times .75/2 \times \sqrt{13.04} = 1093$$

which is actually quite close to the value obtained from the more complex theory [7].

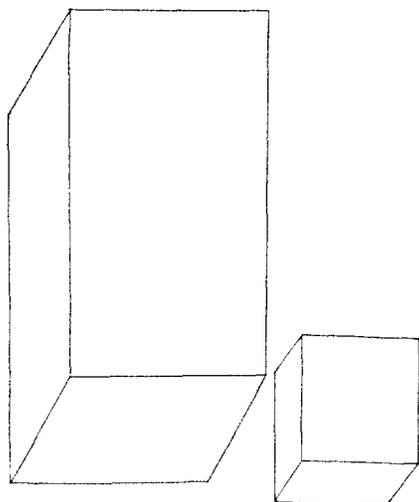
The resonator  $Q$  is lowered by that of the capacitors, typically about 2000-5000. In the present example computer analysis of the filters indicated that the effective  $Q$  reduced to approximately 850.

In previous helical resonator filters inter-resonator coupling is usually electromagnetic, obtained 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

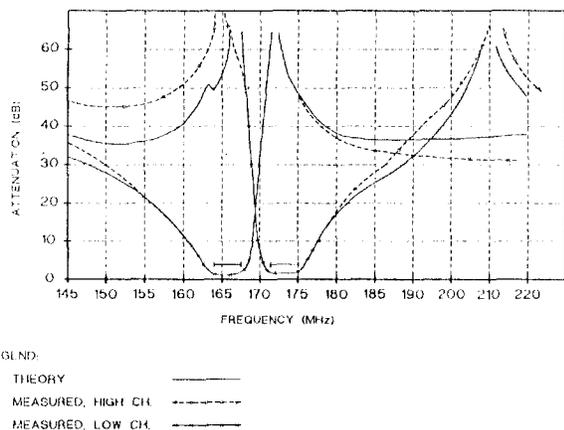
having low loss tangent. In our case the dielectric was Duroid 5880 (TM, Rogers Corporation) which has a manufacturer rated loss tangent of 0.0004 at 1 MHz, increasing to 0.0009 at 10 GHz.

The diplexer constructed in brass weighs 3 lbs; a much lower weight would be obtained using aluminum. Figure 6 illustrates the proportionate volume reduction, at no loss in performance, obtained by replacing the original combline diplexer with this helical resonator design.



**Figure 6.** Illustrating the relative size reduction realized by the present diplexer compared with the combline version.

The measured results in Figure 7 agree reasonably well with theory, especially in the passbands. The stop band rejection differ somewhat from theory, explainable partly by the way the diplexer was tuned. The tuning was performed in order to meet



**Figure 7.** Actual measured diplexer performance.

the specification objectives rather than to achieve the closest fit between theoretical and experimental performance. For example, the theoretical passband return loss would have been higher than that obtained with the experimental tuning. This would have given extra stopband rejection.

### Other Applications

This diplexer, with 2.5 dB passband insertion loss, may be considered too lossy for many applications, but in applications with somewhat wider passbands and generally less severe adjacent frequency isolation requirements, insertion losses as low as 0.2 dB may be obtained, in which case the diplexers would be capable of quite high power transmission.

These types of helical resonator elliptic-function filters and multiplexers may be used at higher frequencies. The usefulness of the technique is limited by the fact that as frequency increases the maximum Q of a helical resonator decreases. This may be seen from (7) which expresses the well known fact that the Q is proportional both to the coil diameter and to the square root of frequency.

As frequency increases the necessary coil diameter decreases linearly with frequency in order that the slightly-above-resonance condition be maintained, so that overall the realizable Q of an optimal helical resonator decreases proportionally to the square root of frequency. Of course, for a given coil the Q increases with frequency, but eventually it becomes self resonant.

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*The requirement of this design that the coils have self resonance above the passband dictates smaller sized coils with increasing frequency, and correspondingly lower Q.*

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For example a similar diplexer built for the GPS bands of 1227 and 1575 MHz was constructed with helical resonators having Q values of around 400, which is about 1/3 of that achievable at 170 MHz, as expected from the above considerations. However the size was very small, and when considering passband insertion loss, the elliptic function characteristics used have the same effect as doubling the Q when compared to regular Tchebyshev filters, so that in some circumstances the technique becomes competitive with cavity filters having the customary Tchebyshev response.

It is worth noting that quasi-elliptic-function lumped element filters are widely used for microwave filters up to quite high frequencies (as much as 10 GHz) with resonator Q values of the order of 150 - 250, in which cases the inductors may or may not be designed to operate close to resonance.

## References

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Ralph Levy received his B.A. degree in Physics from Cambridge University (St. Catharine's College) England in 1953, the M.A. degree in 1957, and a PhD from London University in 1966.

From 1953 to 1959, he was a member of the Scientific Staff at GEC, Stanmore (now Marconi Space and Defence Systems), England, where he was engaged in guided missile, radar, and countermeasures systems, and on waveguide components. In 1959 he joined Mullard Research Laboratories, Redhill, Surrey England, and continued his work on microwave components and systems, utilizing coaxial and stripline media. He developed a widely used technique for accurate instantaneous frequency and/or bearing measurement using several microwave discriminators in parallel (Digital IFM). This FCM work included the development of very broad-band components, such as decade hand-width directional couplers and broadband matching theory applied to amplifiers.

From 1964 to 1967 he was a member of the faculty at Leeds University and carried out research in microwave network synthesis, including realizations of distributed elliptic function filters, and exact synthesis techniques for branch guide and multi-aperture directional couplers. During this period he consulted for GEC, Decca Radar, and Weinschel Engineering.

From 1967 until 1984 he was with Microwave Development Laboratories, Natick, Mass., as Vice President of Research. His research work there resulted in practical techniques for designing very broad band mixed lumped and distributed circuits, and synthesis of a variety of microwave passive components, including development of multi-octave multiplexers in suspended substrate stripline.



From 1984 to 1988 he was with KW Microwave, San Diego, CA., as Vice President of Engineering, working mainly on design implementations and improvements in their filter based products.

During August 1988 to July 1989 he was with Remec Inc. San Diego, CA., as Vice President, and continued with advances in suspended substrate stripline components, exact and user-friendly synthesis of filters with arbitrary finite frequency poles, and microstrip filters.

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Dr. Levy is the author of more than 50 papers, 2 books, and 12 patents. He has been active in professional activities with the IEEE, and during the 1986-88 period was Editor of the IEEE Transactions on Microwave Theory and Techniques, and a member of the MTT-S ADCOM. He is currently an officer of the San Diego MTT Chapter, and is a member of the Steering Committee for the 1994 MTT-S Symposium.

Konrad Andersen, received his BSEE degree at California State Polytechnic University, San Luis Obispo, 1965, and took graduate courses in advanced mathematics and engineering at San Diego State University, during 1966-1968.

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Mr. Andersen has over 20 years professional experience in electronic and systems engineering, and has expertise in the design and fabrication of RF communication equipment, RF switches, and filter subassemblies for military applications.

Much of his experience has been with the development of tactical and strategic communication and navigation systems. He has contributed to numerous military programs including the Navy EHF SATCOM Program (NESP), MILSTAR, the Global Positioning System (GPS), the Joint Tactical Information Distribution Systems (JTIDS), Circuit MAYFLOWER, the High Frequency Improvement Program (HFIP), the Automatic Guard Receiving Terminal (AGRT), and the LINK-11 Improvement Program. More recently, he designed and fabricated a filter subsystem for the U.S. Air Force Special Communications Integrated Package (SCIP) and the Sensitive Compartmental Information Facility (SCIF).

Mr. Andersen had design responsibility for a Navy GPS User-Terminal. Modules developed included the antenna, preamplifier, pseudo-noise (PN) correlator, time-of-week generator, low-noise frequency reference, data processor, code and carrier tracking circuitry, and the range and velocity measurement functions.

He is providing RF system engineering support for the design and development of the Miniaturized GPS Receiver being developed under the sponsorship of DARPA and the application of the resulting gallium-arsenide chip set to the Marine Corps Small Unit Navigation Systems (SUNS).

While at the Naval Ocean Systems Center (NOSC), during 1971-1979, Mr. Andersen was program manager and responsible RF systems engineer for various Navy communications systems. His design experience includes spread-spectrum modulation techniques, software development, digital timing and control circuitry, receivers, and transmitters.

