Waveguide Filters You Can Build – *and* Tune

Part 1 – Waveguide Post Filters

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In all but the simplest microwave systems, filtering is needed, to eliminate both undesired radiation and unwanted interference. For most amateurs, this means copying some published filter. Simple filters often have inadequate performance, while more complex ones can be difficult – in understanding, fabrication, or tuning. Waveguide filters can be designed to provide good performance using available software and fabricated by a modestly equipped amateur.

Waveguides have very low loss because the energy is contained inside the guide, in air, rather than traveling in a conductor. A resonant length of waveguide, with very low loss, thus forms a high-**Q** resonator; for X-band waveguide, the theoretical **Q** approaches 10,000[1]. This high unloaded **Q** enables design of very sharp filters with low loss. Since only metal and air are involved, and the waveguide dimensions are tightly controlled, results are quite predictable.

If a review of filter terminology and design basics is needed, please see the Appendix, "Filter Tour," for a brief filter overview with minimal mathematics.

Figure 1

The waveguide filters to be described are direct-coupled resonator filters[2]. The **WGFIL** program[3] by Dennis Sweeney, WA4LPR, does an excellent job of designing either iris or post filters. An iris filter is more intuitive – each waveguide iris is a perpendicular wall, so that two irises create a resonant cavity in a section of waveguide. A small hole in each iris provides coupling out of the cavity, with coupling controlled by the size of the hole. Fabrication involves cutting very thin slots across the waveguide and soldering an iris in each slot.

Waveguide post filters are much easier to build – just drill a hole through the center of the wide dimension, insert a post all the way through, and solder both ends. Design is a bit more complicated: a post is a shunt inductance in the waveguide, which acts both as a cavity wall and as an impedance inverter coupling the adjacent cavities. A larger post blocks more of the guide, so coupling is reduced by larger posts. The most difficult calculation is the distance between posts for a desired resonant frequency, since the diameter of the post also affects the resonance. The **WGFIL** program does an excellent job here and I have made several filters that perform exactly as expected; a few of them are shown in Figure 1. These are tabulated in Table 1 for those who prefer to duplicate a proven design.

Since a waveguide filter is not hard to build and results are predictable, there is a temptation to design a really high performance filter, with multiple sections. The filter may be easy to build, but it is really difficult to tune – we must allow for some tuning to compensate for construction tolerances. A multi-section filter has extremely good stop-band rejection, and a mistuned filter has NO passband – if nothing detectable gets through, then it is nearly impossible to do any tuning. I have a nice 6-section filter for 24 GHz that I've never been able to tune properly, even with a fancy Vector Network Analyzer (VNA).

Some very good waveguide post filter designs have been published – I can recommend filters by N6GN for 10 GHz[4] and 5.76 GHz[5]. These are three section (four-post) filters with good performance which can be built and tuned by a reasonably well-equipped microwaver. Figure 2 shows the performance of two 10 GHz filters built and tuned in my basement – all the data shown is measured with a nice Rohde & Schwarz **ZVA** VNA (www.rohde-schwarz.com) set up by Greg, WA1VUG, at the Eastern VHF/UHF Conference in 2008.

The easy-to-build waveguide post filters use surplus waveguide and posts of brass or copper hobby tubing, available at some hardware stores and hobby shops, or online from www.smallparts.com. The hobby tubing comes in increments of 1/32 inch diameters, so only a few of the smallest sizes are suitable for the higher microwave bands, particularly 10 GHz and 24 GHz. This severely limits filter design to a very few bandwidths, particularly for multiple-section filters that require several different post diameters.

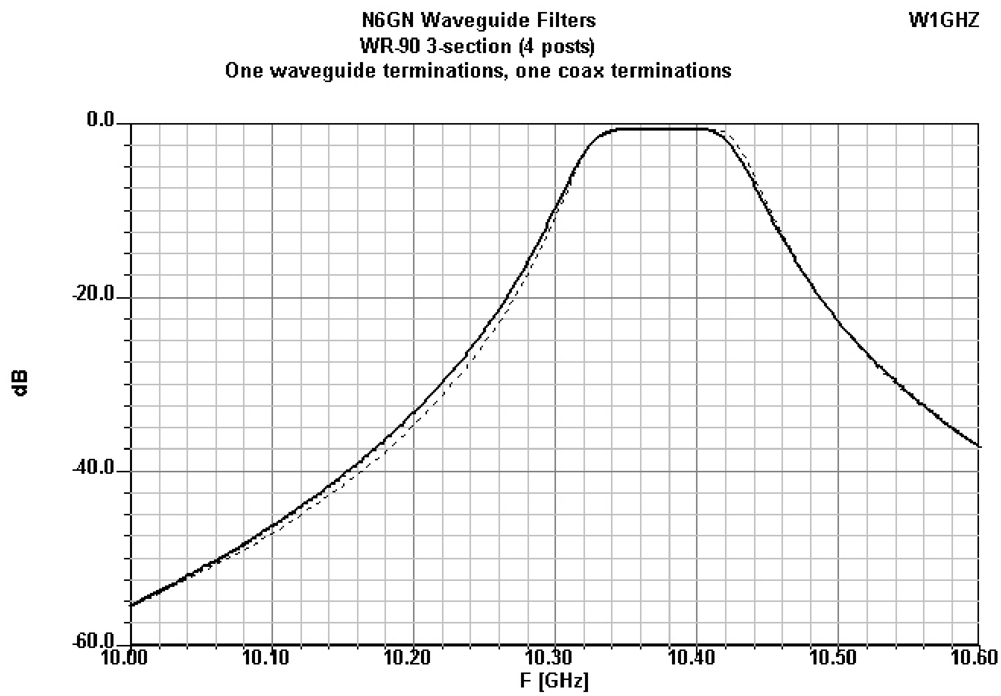

Figure 2

Simple Double-Tuned Filters

While I was playing with **WGFIL** trying to find a better filter, it occurred to me that hams don't need multiple-section filters with steep skirts – our main requirement is to reject LO leakage and mixer images. The most popular IF frequency is 144 MHz, so a usable filter needs perhaps 30 dB rejection at a frequency 144 MHz away from the operating frequency, like the ones in Figure 2. A balanced mixer provides some additional rejection, and the image frequency, twice as far removed, will be further down.

This requirement doesn't seem too difficult. Perhaps something as simple as a double-tuned filter would be adequate. It could be a narrow, high-**Q** filter, because waveguide has low loss and a double-tuned filter has only two tuning adjustments – and they should be identical. Some trial runs with **WGFIL** for a two-section (three post) filter suggested that several possibilities existed using the available hobby brass. Figure 3 is a sketch showing the simplicity of these filters – this sketch includes integral waveguide-to-coax transitions.

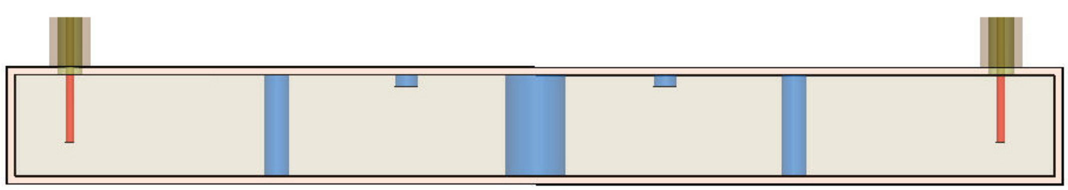

Figure 3 – Sketch of 3-post Waveguide Filter

I did a bit of simulation using Ansoft[6] **HFSS** electromagnetic software, then built a couple of filters for 5760 and 10368 MHz. These worked so well that I tried some other waveguide sizes and frequencies. Measured results are shown in Figure 4 for 10368 MHz, with filters in both WR-90 and WR-75 waveguide, with different bandwidths. All provide adequate LO rejection for a 144 MHz IF at 10368 MHz, with about 0.7 dB loss at the operating frequency. The narrower two, WR75-20, with 20 MHz bandwidth, and WR90-30, with 30 MHz bandwidth, appear over-coupled. The latter one is tuned to one of the peaks, the one providing best rejection, rather than centered. We shall see later how to adjust a filter design for flat coupling.

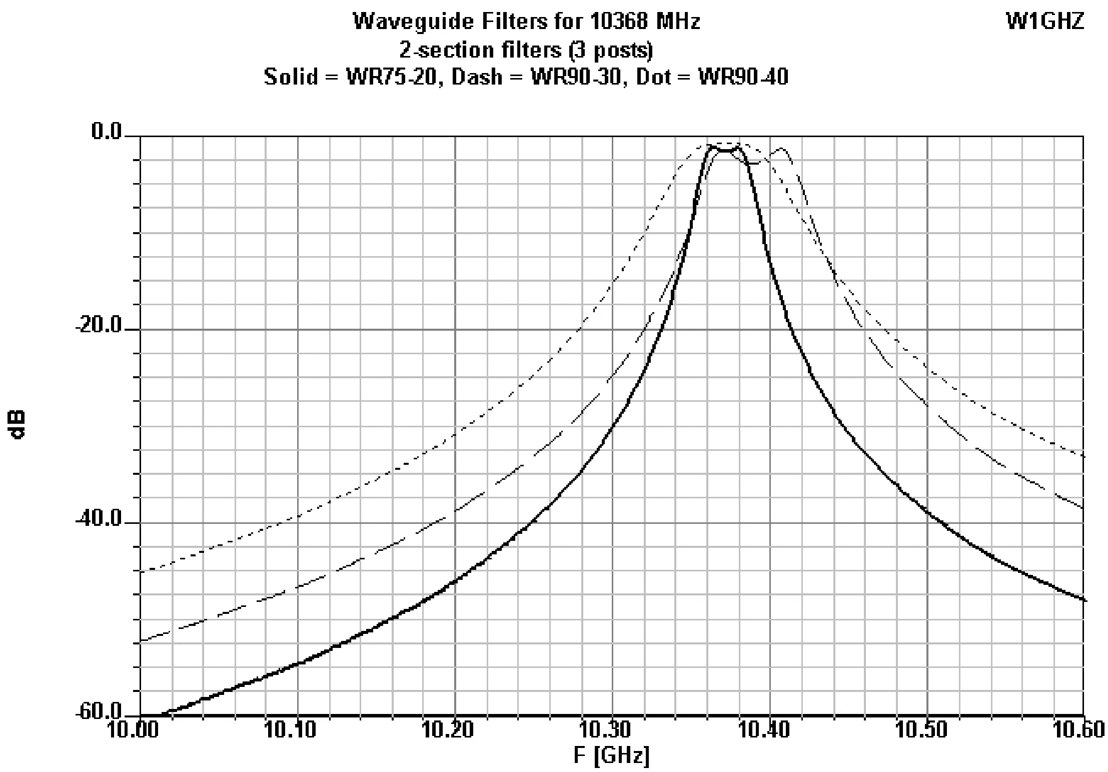

Figure 4

Figure 5 plots filters for 5760 MHz, in two different waveguides, WR-137 and WR-159, with different bandwidths. All of these provide at least 35 dB of LO rejection for a 144 MHz IF at 5760 MHz, with about 0.5 dB loss at the operating frequency. The version in WR-159 waveguide is slightly overcoupled and tuned to one of the peaks, rather than centered.

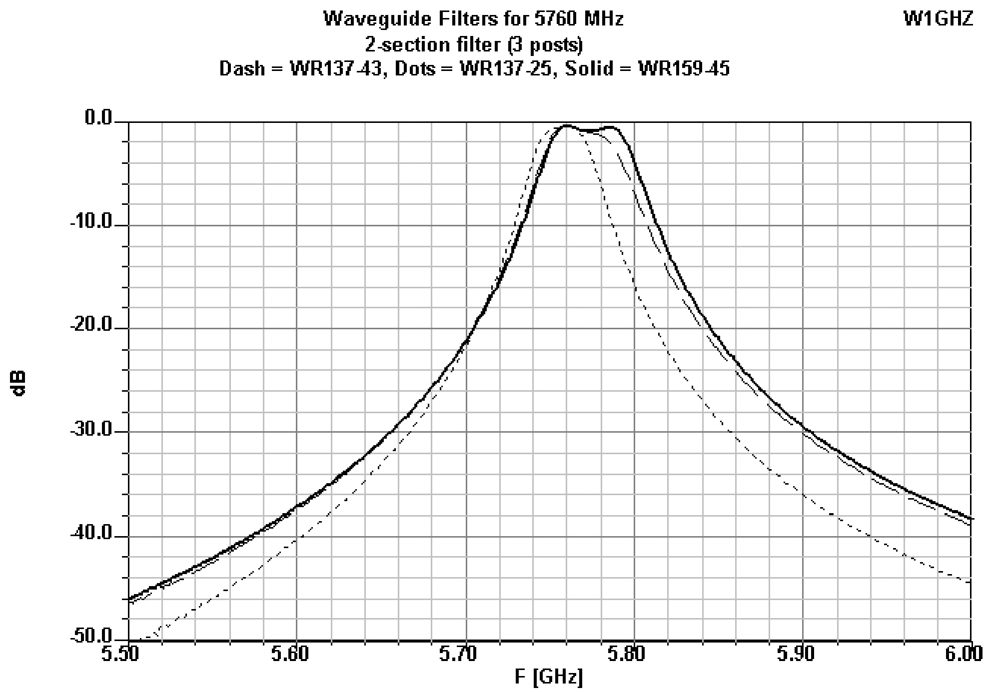

Figure 5

Waveguide filters for 3456 MHz are relatively large, but provide excellent performance, shown in Figure 6. Loss of these filters is less than ¼ dB, with a flat response, and LO rejection for a 144 MHz IF is more than 30 dB.

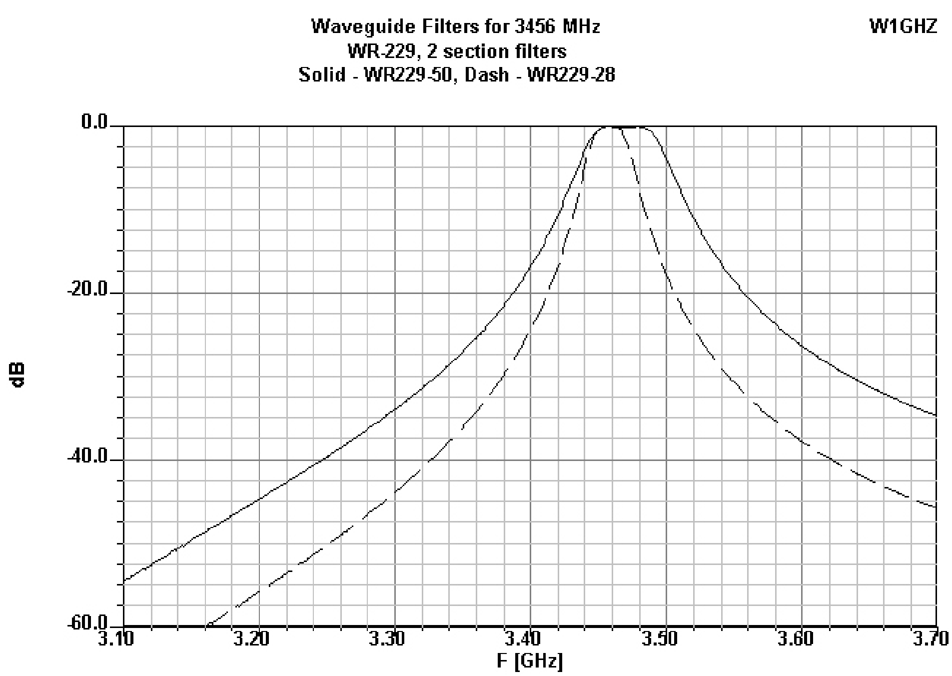

Figure 6

All of these two-section filters were easily tuned using basement test equipment, since there are only two tuning screws which should have identical settings. Also, the dimensions were chosen so that tuning would only require a small penetration by the tuning screws.

Over-coupled Filters

The response of most of these filters has a nice flat top, with reasonably steep skirt – an ideal Maximally-flat, or Butterworth, filter. However, a couple of them show an overcoupled response, with a dip in the middle. These are a little harder to tune, since it requires picking the hump that gives the best LO rejection and adjusting accordingly. The Return Loss is particularly sensitive to overcoupling; in Figure 7, the Return Loss is very good at the hump frequencies, but not as good in between. With ideal coupling, the Return Loss would be good over the whole passband. The **WGFIL** calculations are pretty good, but not perfect, especially for larger post diameters – that's why it gives a warning when the posts are bigger than ¼ of the waveguide width. As a result, we don't always get perfect coupling, especially when we round off the diameter to the nearest 1/32".

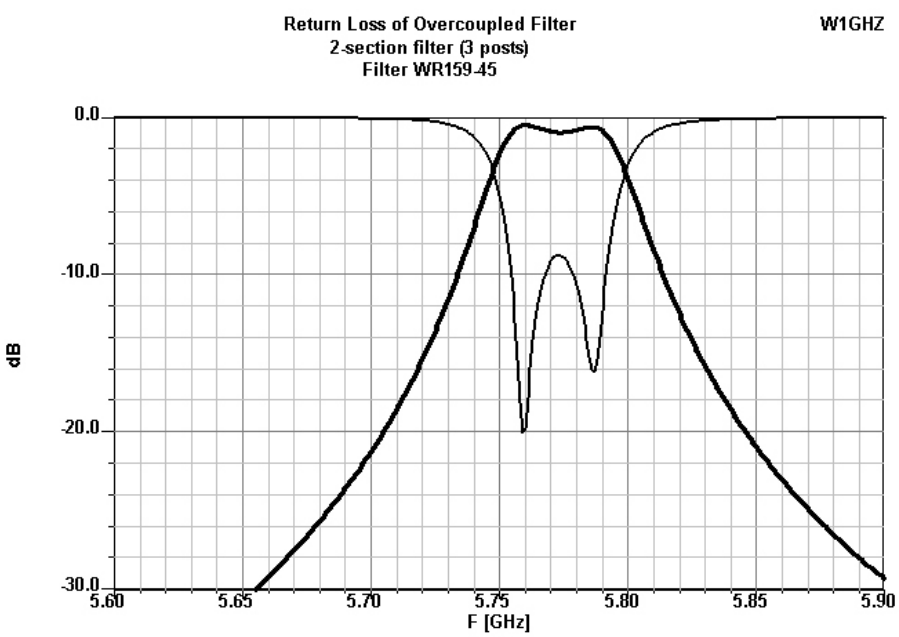

Figure 7

I really wanted to make a reproducible 24 GHz filter, since I don't know of any that have been published. Trial runs with **WGFIL** weren't as promising, since only two or three sizes of hobby tubing are small enough, and I only found one promising combination.   I considered using AWG wire sizes – copper wire is readily available – but the diameters weren't right. Some commercial filters use multiple small posts rather than a large one, but a bit of research didn't find any simple answers for using them.

I recalled that some commercial filters have an extra screw next to each post; does it vary the coupling? I went back to **HFSS** to find out. What I found was that the extra screw increases the coupling, in effect making the post *smaller*. This was the answer. I also found that the coupling screw must be inserted a long way, nearly half the waveguide height, to have a significant effect, so that small adjustments should be easy. I designed and built two more 10 GHz filters in WR-75 waveguide, with slightly oversize center posts, to try out the coupling screw. The coupling screw should decrease the effective size of the center post to adjust the coupling to the desired response. It worked perfectly, as shown in Figure 8. Both filters are adjusted for a flat response and centered on the operating frequency. The wider one, with 42 MHz bandwidth, has lower loss, about 0.75 dB. The narrow one, with 20 MHz bandwidth, has higher loss, about 2 dB, but is sharp enough to provide about 20 dB of LO rejection for a 30 MHz IF at 10368 MHz. It is not surprising that the sharper filter has more loss, since a higher loaded **Q** is needed for the narrower bandwidth.

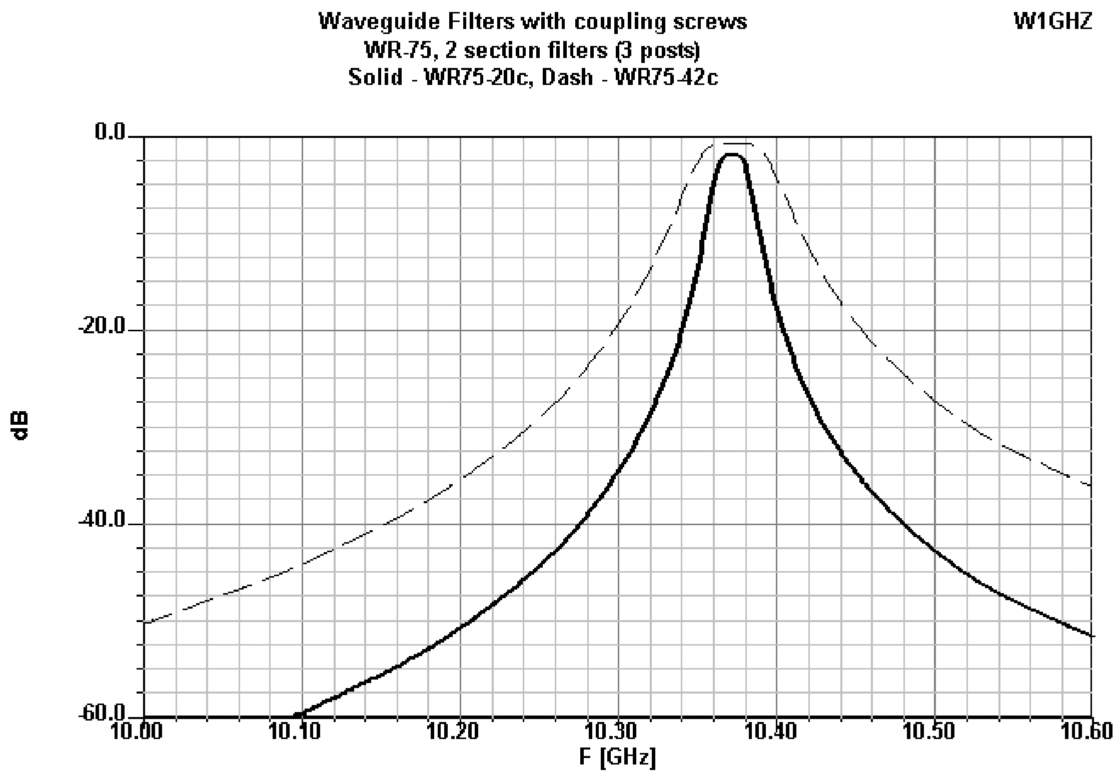

Figure 8

The coupling screw is next to the center post, halfway between the post and the side wall of the waveguide. While it would be possible to put a coupling screw next to the other posts and make them adjustable also, tuning would no longer be straightforward. For the simple 3-post filters, it is unnecessary, and each screw adds a small additional loss.

Then I made two filters for 24 GHz, in WR-42 waveguide. One was the best combination of available post diameters I could find, while the other has an oversize center post and a coupling screw. The results are shown in Figure 9 – both filters are sharp enough to use for a 144 MHz IF at 24 GHz. The one with the coupling screw has a nice flat response, while the other is slightly over-coupled and has a bandwidth slightly wider than expected as a result. Each has about 2.5 dB loss – not bad for a sharp filter at 24 GHz. These filters were tuned up at a single frequency, 24.192 GHz, since I don't have a sweeper for 24 GHz. The plotted data was measured with the VNA without any retuning. Figure 10 is a photo of the 24 GHz filters.

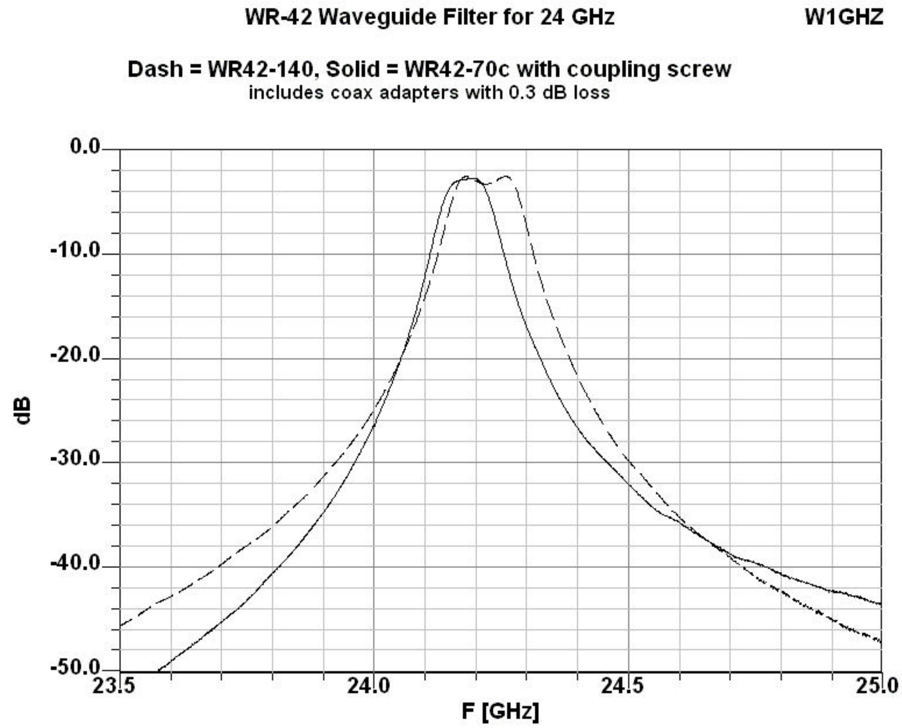

Figure 9

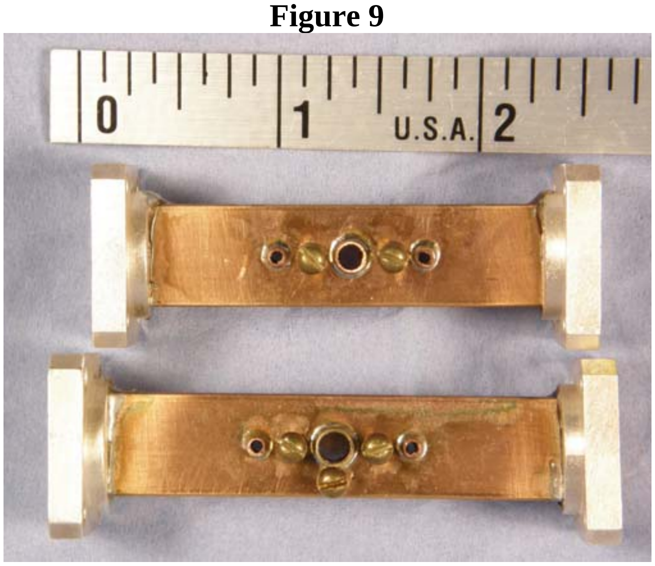

Figure 10 – WR-42 waveguide filters for 24 GHz

Tuning these filters now becomes very easy. Starting with all screws all the way out, the two tuning screws are slowly inserted simultaneously (turn one, then the other the same amount) until some output is found. Then peak the output. Since the response is undercoupled without the coupling screw inserted, there will only be a single peak. Next, insert the center coupling screw; the output will slowly increase, then start to decrease as the response becomes overcoupled with a dip in the middle. Backing the screw out to the peak yields the desired flat response. A final trim probably won't make much difference.

The tuning progression is illustrated in Figure 11, a simulation of the WR75-42c filter. The curve on the right shows the response before tuning, with no screws present – the filter is tuned to some higher frequency. The tuning screws alone move the response down to the desired frequency, yielding the curve labeled "NO coupling screw." Then the coupling screw is inserted; at 0.100" deep, the response flattens and the loss is reduced. Inserting the screw further produces an over-coupled response, first with a slight dip at the operating frequency, and then a huge one if insertion is continued. Most of us would back up when the output started to dip.

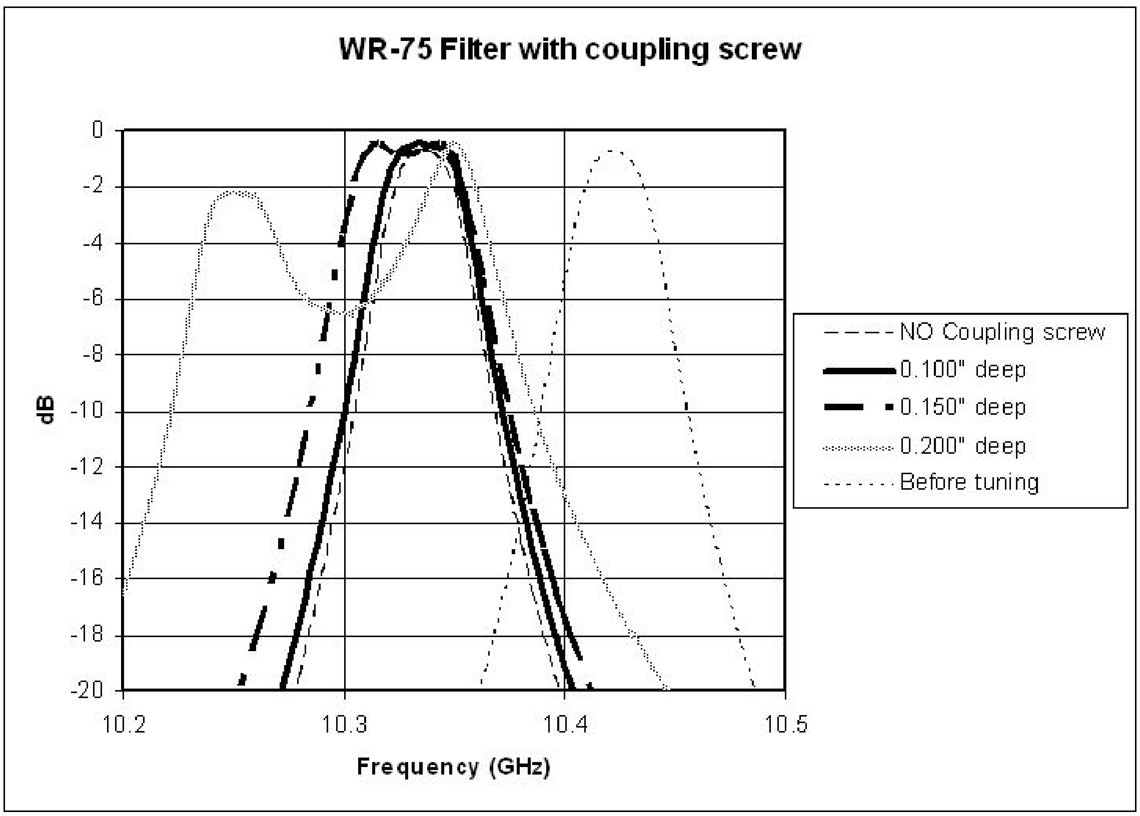

Figure 11

Some of the filters in Figures 4 and 5, as well as one of the 24 GHz filters, show an over-coupled response. These designs could be improved by making the center post one size (1/32") larger and adding a coupling screw next to the center post. Then they could be adjusted for a flat response.

End Termination

In a waveguide system, these filters only need waveguide flanges to connect. However, most systems for 10 GHz and down use semi-rigid coax for interconnections, so a coax – to-waveguide transition is needed. The most compact and convenient transitions are integral to the filters, one at each end. I use the dimensions that I published in ***QEX***[7], spacing the transition probe at least one waveguide width from the end post. A matching screw is neither needed nor desired – if the dimensions are correct, the Return Loss will be very good. Of course, a badly mismatched component following a filter can upset the filter response, but the place to correct this is not in the filter.

A good comparison is the filters with performance shown in Figure 2. One has integral coax transitions, while the other has waveguide flanges and was tested with external transitions. Any slight difference in performance is probably due to construction tolerances and tuning difference.

Construction

These filters are physically simple – the posts, tuning screws, and coax connectors are all on the centerline of the broad dimension of the waveguide. Important points are that the posts be accurately centered on the centerline and that the holes for the posts are snug, so that a minimal amount of solder is needed to make a good connection.

The highest frequency for each resonator is set by the distance between the posts – a tuning screw can only lower the frequency. The distances calculated by **WGFIL** are for no tuning screw, so they should be reduced slightly to raise the resonant frequency and allow a small amount of tuning. I estimate that I can locate a hole within 10 mils (0.25 mm), so I reduce the distance by 10 to 15 mils – adjust according to your tolerances. With only a small reduction, a very few turns penetration of the tuning screw is needed. A larger reduction in spacing will require more penetration, increasing losses and making the tuning more critical.

I measure and mark the centerline and hole positions with a cheap caliper, either dial or digital, using the points as a scribe (this would be criminal abuse with a quality tool). Then the holes are marked with a centerpunch and started with a small center drill. A drill press is essential for drilling the holes. For accurate, round holes that fit the posts snugly, I find that DeWalt "Pilot Point" drills work well; Black & Decker "Bullet" drills are nearly as good. For larger holes, Unibit step drills work very well. A small pilot hole drilled through both sides of the waveguide will allow making the larger holes from opposite sides.

Screw holes are tapped using the drill press to keep them square, turning by hand with the motor unplugged. Then the burrs inside the waveguide are cleaned up using a fine file. The outside of the guide and the posts is cleaned using a Scotchbrite pad – the coarser brown variety may be needed for old waveguide.

If the posts fit snugly, they will need to have one end chamfered slightly so that they may be pressed in – if they are loose enough to fall out, the filter will probably still work but it may have higher loss. Then flux is applied around the ends of each post where they projects from the waveguide. Finally, a single ring of thin solder is wrapped around each end of the posts and pressed into the flux to hold it in place.

Ends

If the filter includes an integral waveguide-to-coax transition, the ends must be closed with a short circuit. I use a plate of hobby brass a bit larger than the waveguide outside dimensions, so that the plate has a bit of overhang. I paint the ends of the guide, which have been filed square, with solder flux, then put the end plates on and clamp them in place. Scraps of firebrick or ceramic tile insulate the clamps from the end plates. Finally, I wind a ring of solder around the waveguide and press it into the flux to hold it in place.

Preparation for waveguide flanges is similar.

Soldering

For soldering with soft solder, I prefer a hot air gun to a torch. A hot air gun, the kind used for stripping paint, has no flame and doesn't get as hot, so the metal oxidizes less. I've had good results preheating the filter assembly on a hot plate to near soldering temperature, and then applying the hot air gun to each area being soldered. A few seconds after the hot air is applied to a spot, the ring of solder around the joint will melt and flow into the joint. As soon as the solder flows around the whole ring, move on to the next joint. When all the joints have been flowed, gently move the assembly onto a firebrick or other heat-tolerant surface to cool slowly.

Summary

The filters described here are intended to provide good performance with minimum complexity, so that they are easy to design and to tune.   These waveguide filters offer higher performance but do require some metalworking. Some proven designs are tabulated and the **WGFIL** software is sufficient to design custom filters.

All the filters described here are designed for "good enough" performance at a particular microwave ham band. Good enough means that commonly-used LO frequencies and mixer image frequencies are suppressed by at least 20 dB, and more than 30 dB in most cases. This should be adequate to radiate a clean signal and to suppress out-of-band interference.

3-Post Waveguide Filters

Table 1 W1GHZ 2008

Waveguide	Frequency MHz	Bandwidth MHz	End Post Diam - in	Mid Post Diam - in	Spacing in	Data
WR-75	10368	20	0.125	0.250	0.950	WR75-3-20
WR-90	10368	30	0.188	0.313	0.860	WR90-3-30
WR-90	10368	40	0.156	0.313	0.830	WR90-3-40
WR-137	5760	25	0.188	0.406	1.620	WR137-3-25
		43	0.156	0.375	1.600	WR137-3-43
WR-159	5760	45	0.250	0.500	1.480	WR159-3-45
WR-187	3456	7	0.250	0.438	3.000	(not built)
WR-229	3456	50	0.188	0.500	2.500	WR229-3-50
		28	0.250	0.625	2.540	WR229-3-28
WR-42	24192	140	0.094	0.156	0.360	WR42-3-140

With Coupling Screw

WR-75	10368	20	0.125	0.281	0.970	WR75-3c-20
WR-75	10368	42	0.094	0.250	0.940	WR75-3c-42
WR-42c	24192	70	0.094	0.188	0.375	WR42-3c-70c

References

1. G. F. Craven & C. K. Mok, “The Design of Evanescent Mode Waveguide Bandpass Filters for a Prescribed Insertion Loss Characteristic,” *IEEE Transactions on Microwave Theory and Techniques*, March 1971, pp. 295-308.

2. Ralph Levy, “Theory of Direct-Coupled-Cavity Filters,” *IEEE Transactions on Microwave Theory and Techniques*, June 1967, pp. 340-348.

3. Dennis G. Sweeney, WA4LPR, “Design and Construction of Waveguide Bandpass Filters,” *Proceedings of Microwave Update ’89*, ARRL, 1989, pp. 124-132. The WGFIL program may be downloaded from www.w1ghz.org/filter/WGFIL.COM

4. Glenn Elmore, N6GN, “A Simple and Effective Filter for the 10-GHz Band,” ***QEX***, July 1987, pp. 3-5.

5. Paul Wade, N1BWT, “A Dual Mixer for 5760 MHz with Filter and Amplifier,” ***QEX***, August 1995, pp. 9-13. available at www.w1ghz.org/10g/QEX_articles.htm

6. www.ansoft.com

7. Paul Wade, W1GHZ, “Rectangular Waveguide to Coax Transition Design,” ***QEX***, November 2006, pp. 10-17. available at www.w1ghz.org/10g/QEX_articles.htm

Part 2 - Evanescent Mode Waveguide Filters

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The only amateur publication of evanescent mode waveguide filters, to the best of my knowledge, is by Reed Fisher, W2CQH, in 1993[1]. I noticed the paper when it first appeared, but I recall thinking that they couldn't be very good filters, since evanescent modes have high loss.

Recently, I was looking for references on waveguide filters and came across several for evanescent mode filters. Apparently they actually work, and offer good performance in a compact package. What I hadn't realized is that resonant structures behave much differently than non-resonant waveguides. After perusing a few papers, I went back to Reed's paper to look for practical dimensions.

Starting with some of Reed's examples, I did some simulation with Ansoft **HFSS** electromagnetic software[2] and fiddled the dimensions until a reasonable filter response appeared. Then it was time to make one. Construction is really simple – two SMA connectors with a screw or two between them, as sketched in Figure 1, all on the centerline of the broad side of the waveguide. The critical dimensions are the distances between the screws (red), the *Mid Length*, and from the screws to the connectors (blue), the *Connector Length*. The SMA connector center pin connects to the far wall of the waveguide.

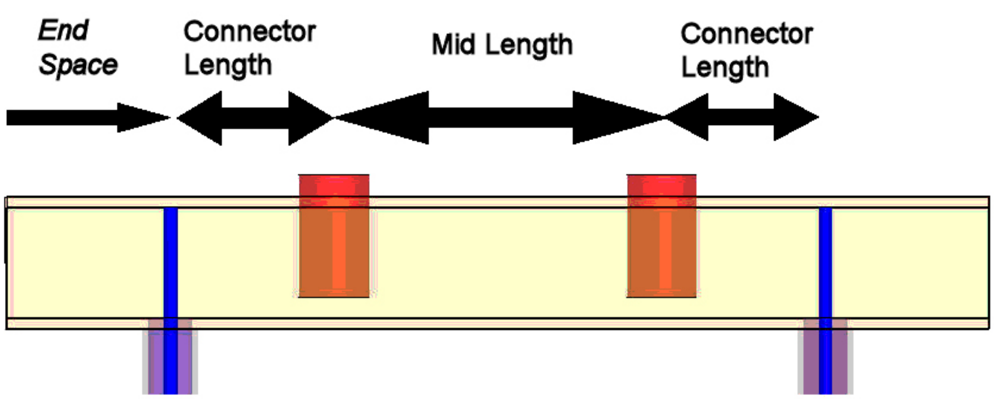


Figure 1 – Sketch of Evanescent Mode Waveguide Filter

The first filter I made was in WR-90 waveguide for 3456 MHz. The cutoff frequency for WR-90 is about 6.5 GHz, so this is way below cutoff – nothing should get through the waveguide at this frequency. However, the filter works pretty well – the reponse is plotted in Figure 2 – a pretty sharp filter, with less than 2 dB of loss.

Construction is exceedingly simple – drill holes for the connectors and screws, tap the screw holes, solder the connector pins to the far wall, and tune it up. Since energy doesn't travel in the evanescent guide without something resonant, the ends may be left open – I was amazed to stick my finger in the end with no effect! The length of the *End Space* in Figure 1 should be more than half of the waveguide width – I usually leave about 20 mm as a convenient length.

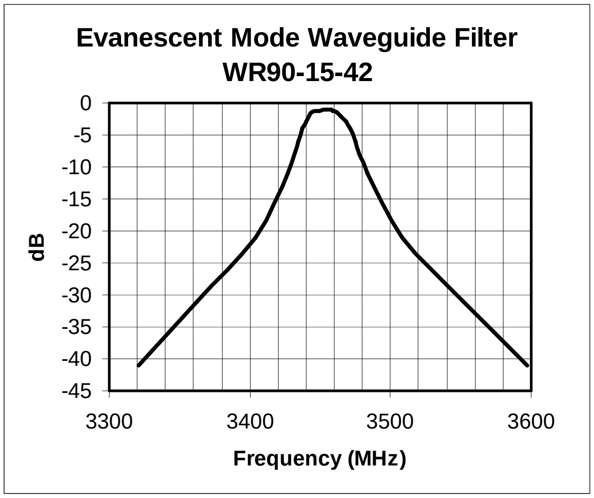


Figure 3 is a photo looking in the end of the finished filter, showing the two screws sticking in to the waveguide. These screws are tuning capacitors, so I used a couple of ¼-28 brass screws that I had in the junkbox with the ends faced flat on a lathe. The flat end might make the tuning smoother – it was still pretty touchy even with the fine-thread screw. I also learned that putting the screw head on the same side as the coax connectors is a bad idea – there isn't much room in which to work. Later versions have the screws on the opposite side, like Figure 1.

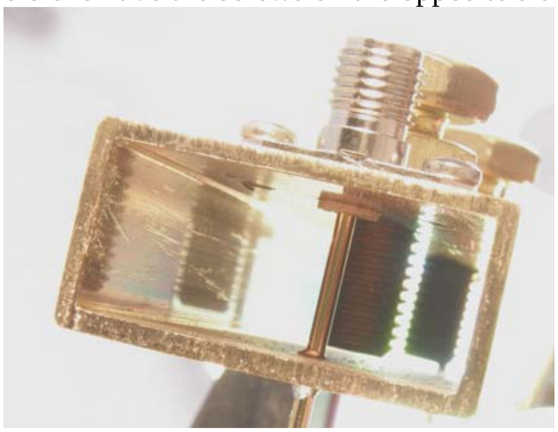


Figure 3 – Inside Evanescent Mode Filter

Evanescent mode Filter Theory

So how do these filters work? Basically, a section of evanescent mode waveguide, well below cutoff, acts as an inductor. We add a capacitor to make a resonant circuit.

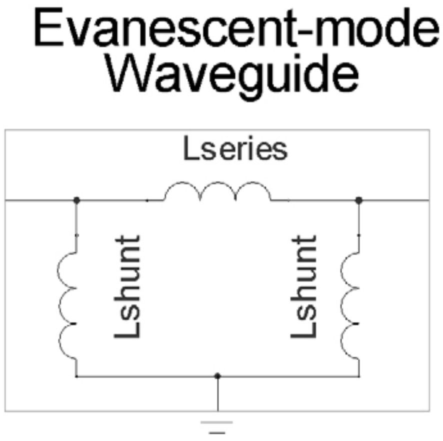


Figure 4

The simple equivalent circuit of a short length of evanescent mode waveguide is shown in

Figure 4, a series inductance with a shunt inductance at each end. If we put connections at each end and a screw in the middle, like Figure 5, it forms the single resonator shown schematically in Figure 6.

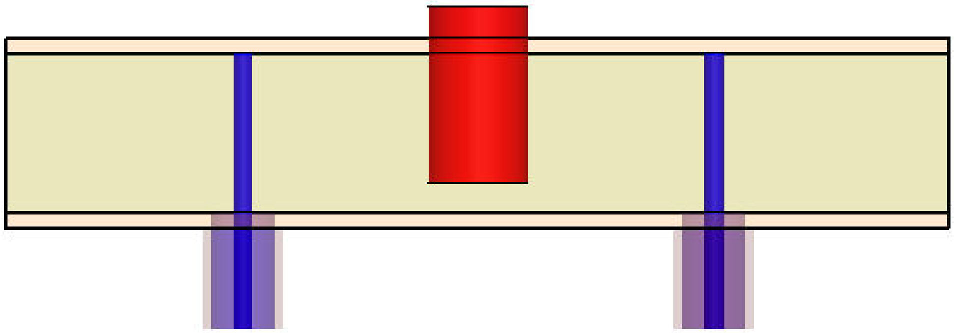


Figure 5

The tuning capacitor resonates with the two Lshunt, one on each side, and the Lseries couple to the SMA connectors, to set the loaded $\mathbf{Q_L}$ of the resonator – making Lseries larger, by increasing the spacing, reduces the coupling, and thus the loading from the 50 ohm source and load, so the $\mathbf{Q_L}$ will be higher.

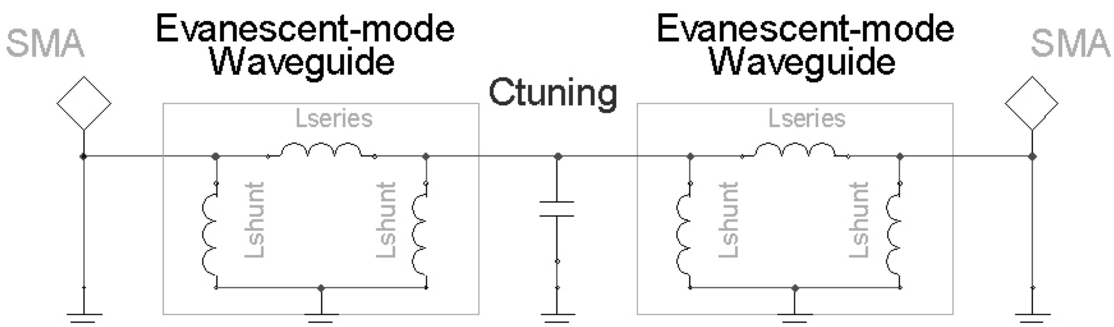


Figure 6

The inductances are calculated from the waveguide length, from the connector to the tuning capacitor (*Connector Length* in Figure 1, center-to-center), and the cutoff wavelength. First, the cutoff frequency of a rectangular waveguide is when the width of the guide is a half-wavelength. For WR-90, the width is 0.9" = 22.86mm, so the cutoff wavelength λ_c=45.72mm. Thus, the cutoff frequency is 300/45.72 = 6.56 GHz.

Craven & Mok[3] show a graph of unloaded **Q** for WR-90 waveguide. The theoretical **Q** is higher than 10,000 at 10 GHz, in a normally propagating $\mathbf{TE_{10}}$ mode, but slightly lower for the $\mathbf{TE_{10}}$ evanescent mode, perhaps 6,000 just below cutoff and falling to around 1,000 at 1 GHz. This is still high enough **Q** to use the lossless transmission line assumption, which simplifies calculations.

For a lossless TE_{10} evanescent mode, the characteristic impedance X_0 is calculated[2] from the cutoff wavelength and the waveguide width **a** and height **b**. λ is the free space wavelength:

$$0\,X_0 = \frac{\mathbf{b}}{\mathbf{a}} \times \frac{120\pi}{\sqrt{\left(\frac{\lambda}{\lambda_c}\right)^2 - 1}}$$

And the propagation constant γ:

$$\gamma = \frac{2\pi}{\lambda}\sqrt{\left(\frac{\lambda}{\lambda_c}\right)^2 - 1}$$

For a length waveguide ℓ, the reactances of the inductors are then calculated:

For Lseries, $X_L = jX_0 \sinh \gamma\ell \cong \frac{jX_0}{2} e^{\gamma\ell}$

And Lshunt, $X_L = jX_0 \coth \frac{\gamma\ell}{2} \cong \frac{jX_0}{2}$

Where **sinh** and **coth** are the hyperbolic sine and cotangent, respectively.

The approximations on the right side are from W2CQH. He uses them for a further approximation to estimate the loaded Q_L:

$Q_L \approx \frac{R_0}{X_0} e^{\gamma\ell}$, where R_0 = 50 ohms for a coax termination.

All these approximations have some error, and the errors add up, so that the estimated Q_L is lower by than the apparent values from simulation and measurement. The discrepancy is as much as a factor of two, which could lead to filters much sharper or lossier than expected. Also, the Johanson trimmer capacitors used to make the lower frequency filters are not nearly as high-**Q** as the waveguide, so the loss of the lower-frequency filters is higher.

Another factor is the inductance of the SMA connector pin – the inductance creates an additional impedance transformer, further reducing the loading and raising the loaded **Q**.

With waveguide-post filters, we found that a double-tuned filter was adequate for many applications. For a double-tuned filter, we need not only the evanescent mode waveguide length at each end, but also an additional length in the center with an Lseries calculated for the *Mid Length* in Figure 1 that provides the desired coupling between the two resonators tuned by the two capacitors. The double-tuned filter is shown schematically in Figure 7 and sketched in Figure 1. The two Connector Lengths should be identical, but the Mid Length is longer – increasing the length decreases the coupling.

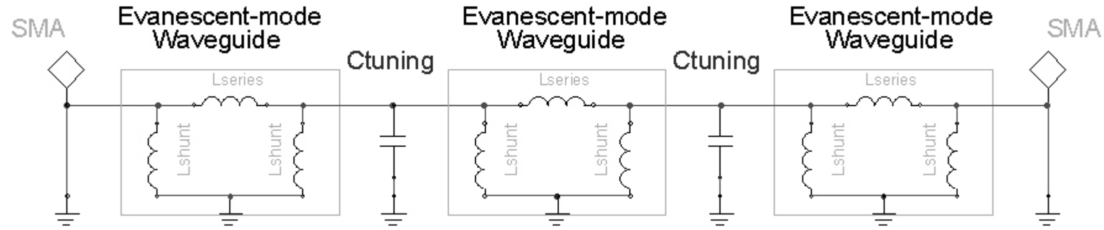


Figure 7

The result of all the approximations and errors is that we cannot calculate the parameters accurately enough using these equations to design a filter, even a simple double-tuned filter. Snyder[4] has more equations, but I have not had a chance to evaluate them. Instead, I have resorted to professional 3D electromagnetic software, Ansoft **HFSS**, to analyze various trial dimensions, and then to build some of the promising ones. Even then, some of the filters have a measured bandwidth slightly narrower than predicted.

More examples

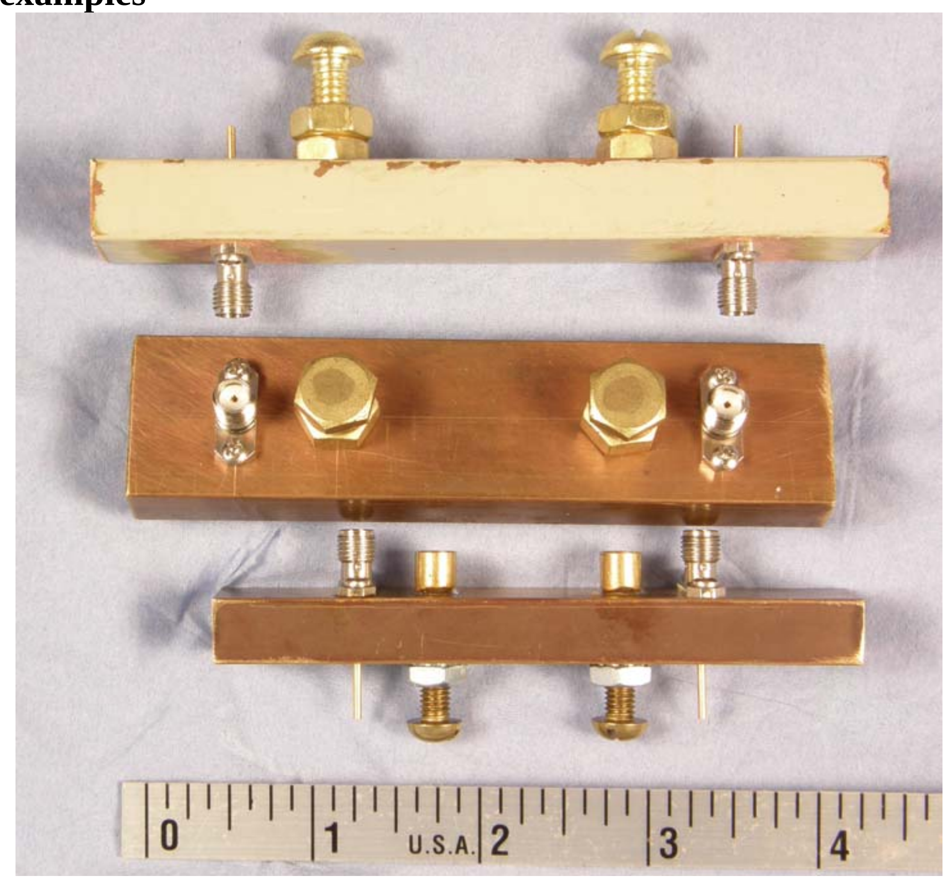


Figure 8

I have made a number of successful evanescent mode filters in all sizes of X-band waveguide, WR-90, WR-75, and WR-62, for frequencies from 5.76 GHz down to below 1 GHz. Figure 8 is a photo of several of the filters; the small physical size of these filters should be apparent. The limiting factor for low frequencies is the tuning capacitance – a simple screw provides only a fraction of a picofarad, and even the Johanson trimmer capacitors are limited to 10 pf or so. For 2.3 and 3.4 GHz, I increased the screw capacitance by soldering a ¼" diameter piece of hobby tuning to the opposite wall, so the #10 screw inside the tubing forms a concentric capacitor. This is sketched in Figure 9, and used in the bottom filter in Figure 8.

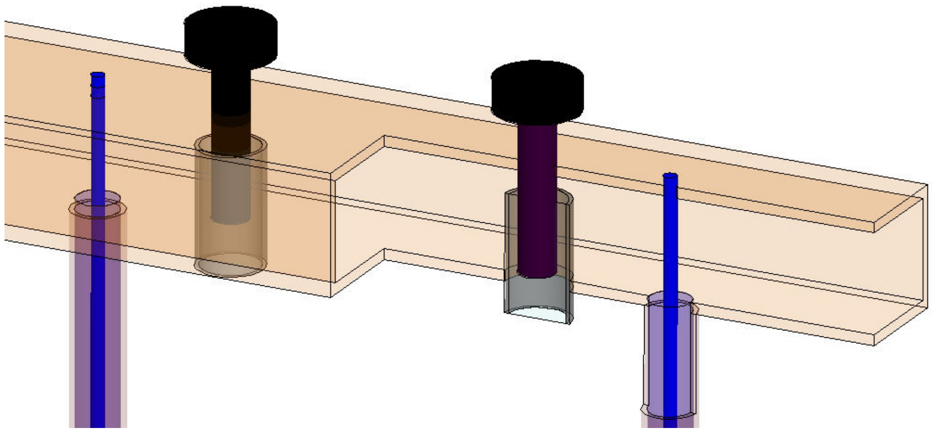
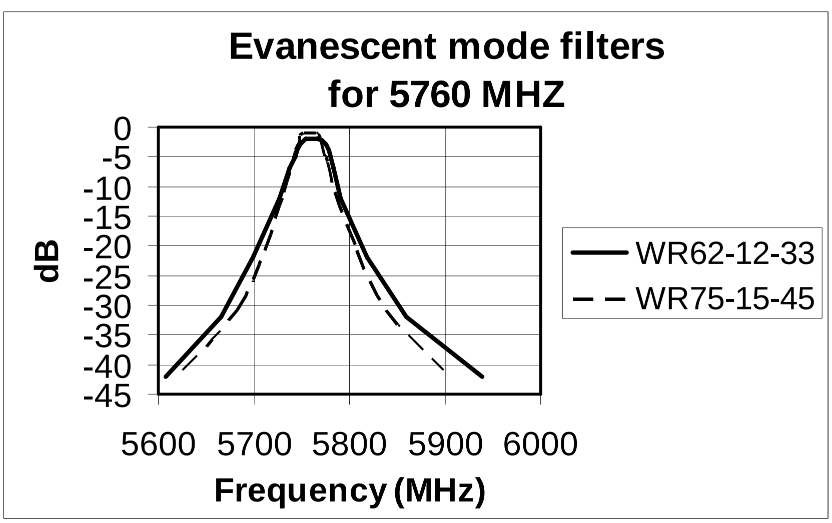


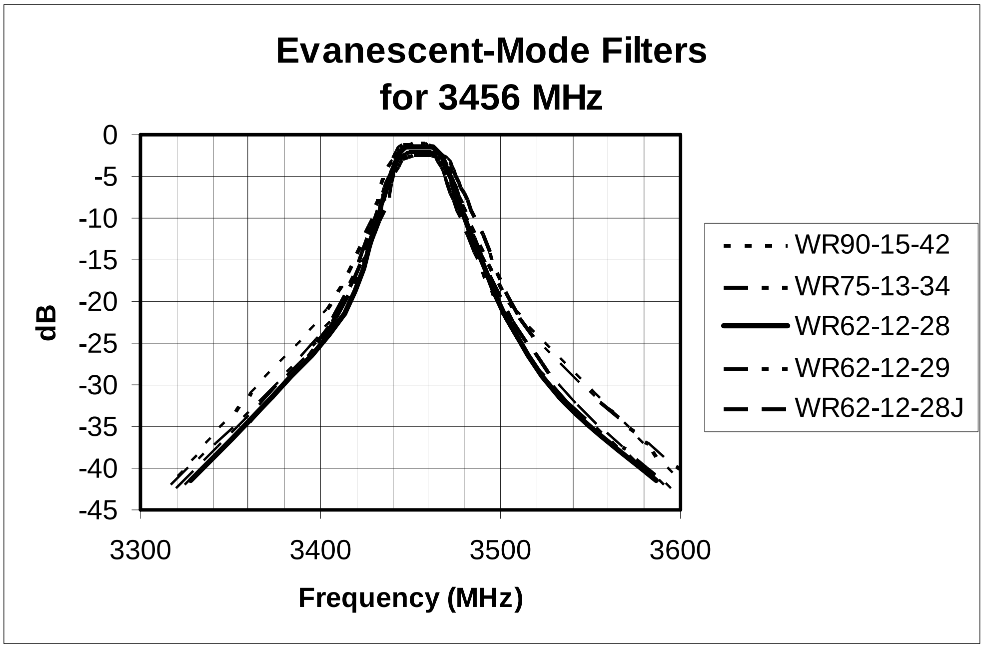
Figure 9 – Concentric screw-inside-tubing capacitor

The filters have a narrow passband with fairly low loss, and a wide stopband – there are no significant spurious responses below the cutoff frequency for the waveguide. Above the cutoff frequency, normal propagation can occur in the waveguide and the filter is less effective. Thus, a WR-90 filter for 5760 MHz could have an additional response starting at about 6.5 GHz, so it would not be very effective. On the other hand, a WR-62 filter would not have any significant spurious response below about 10 GHz, so it can be an effective harmonic filter as well as bandpass filter for 2304 or 3456 MHz. There is plenty of surplus WR-62 waveguide around, not very good for 10.368 GHz operation since it is very close to cutoff, but useful for these filters.

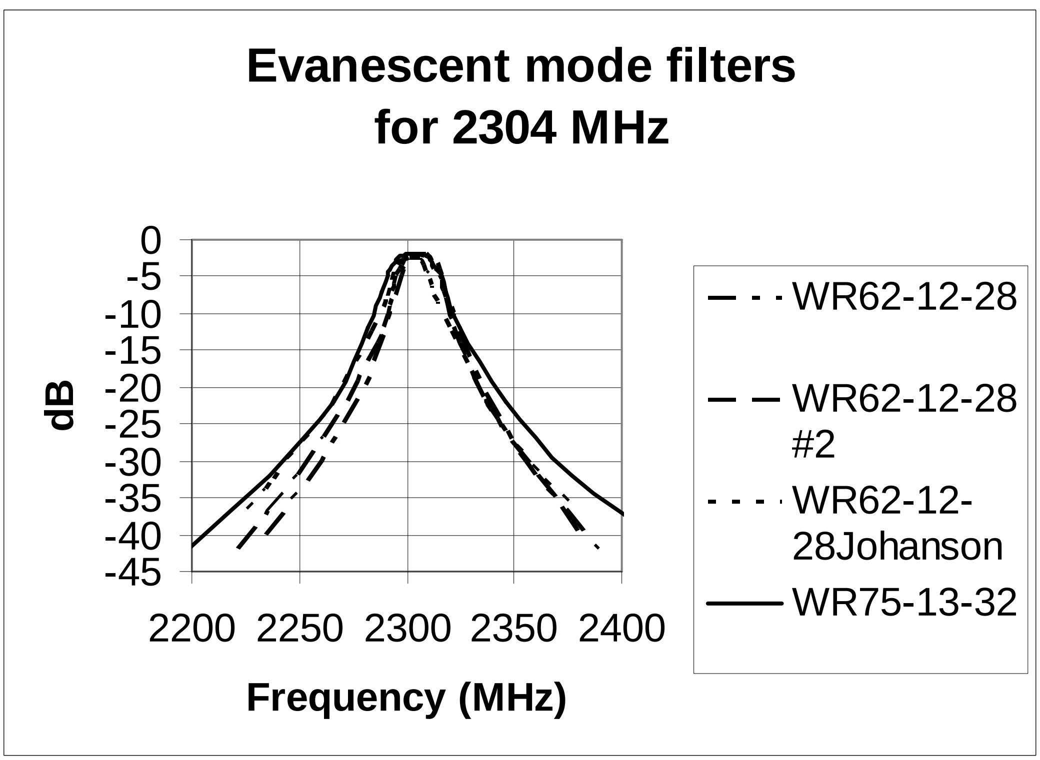


The performance of the 5760 MHz filters is shown in Figure 10, left – these are pretty sharp, yet the loss is under 2 dB.

For 3456 MHz, the performance is shown in Figure 11, also pretty sharp with loss under 2 dB.



The filters for 2304 MHz are small, the same size as 3456 MHz – in some cases, the same filter with the screws farther in – but they still have good performance, shown in Figure 12. The filters with concentric screws have loss under 2 dB, while the WR-62 filter with Johanson trimmers has about 2.5 dB loss at 2304 MHz.



With the higher capacitance of the Johanson trimmers, the lower end of the tuning range is extended. A WR-90 filter with large Johanson trimmers, model 5502, tunes from about 2 GHz down to 580 MHz. Figure 13 is a photo of the filter. WR-62 filters with the more common Johanson models 2954 and 5202 tune from 3456 MHz down to about 980 MHz. I tuned several of them to 1152 MHz for comparison, with results shown in Figure 14. The loss increases as we go down in frequency – I don't know whether this is due to lower **Q** of the evanescent mode waveguide, the trimmer capacitors, or both. Performance of this filter tuned for 902 MHz is shown in Figure 15.

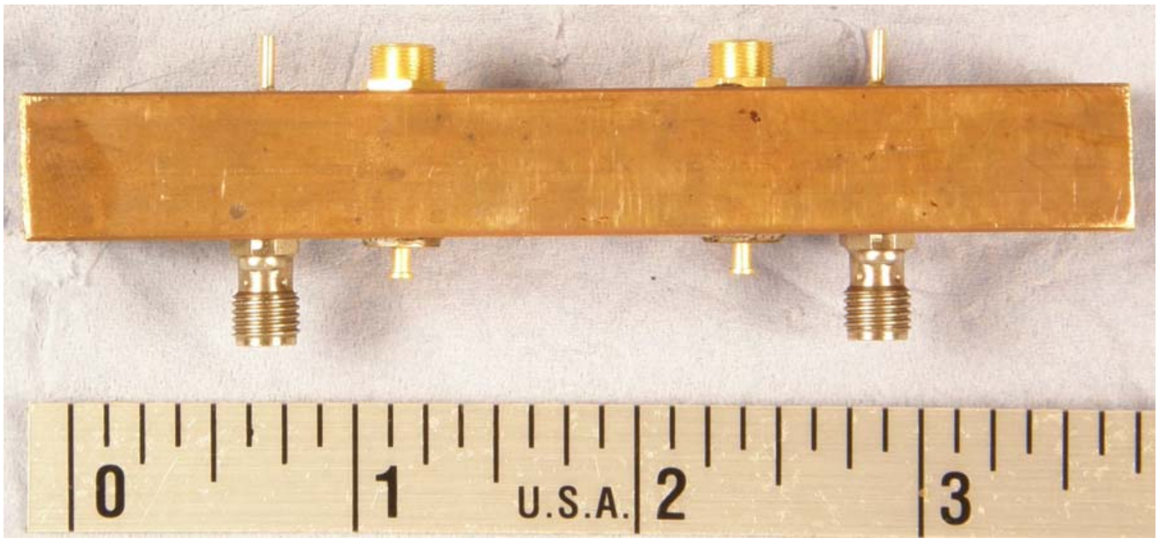


Figure 13 – Evanescent Mode Filter using Johanson capacitors

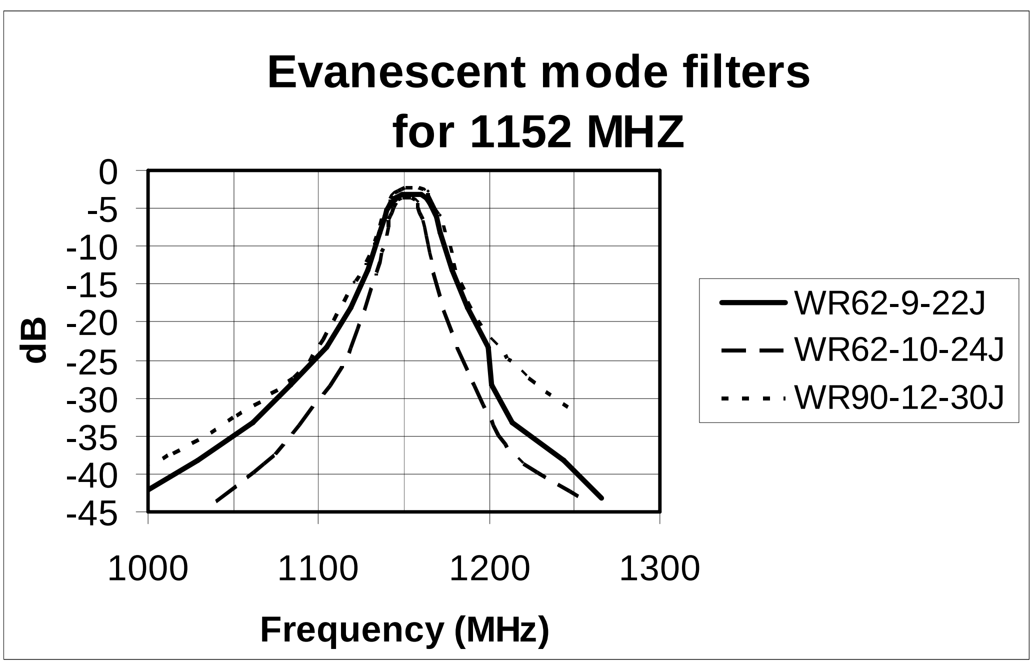


Figure 14

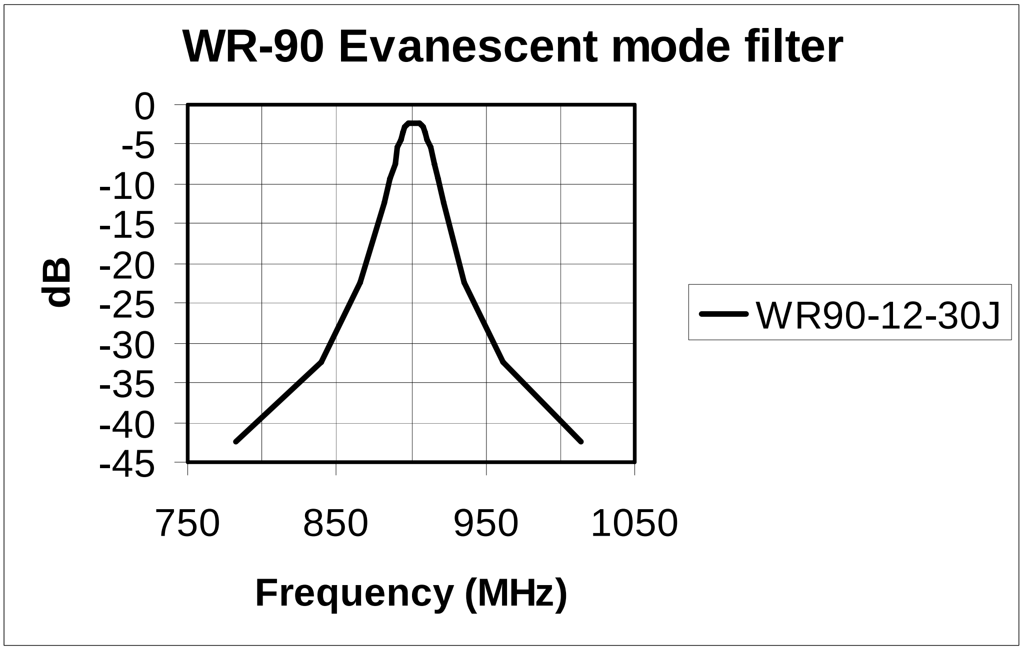


Figure 14

A summary of dimensions for all the successful filters is shown in Table 1.

Tuning the filters can be tricky, unless a swept-frequency test is available, since these filters tune over a wide range. If only fixed frequency testing is possible, then it is necessary to tune both screws together slowly until some output is noted. Then it is simply a matter of tuning for maximum output and minimum VSWR. Most of them tune with the screws inserted quite far into the guide, so it might be easier to start with the screws nearly shorting and back them out slowly.

The performance shown is tuned to ham bands, but several of the filters can be tuned to more than one band. Obviously, they can be tuned to any frequency in between, and more. Thus, the examples given in Table 1 should fulfill most requirements.

Summary

Evanescent mode waveguide filters offer very good performance in a compact package, and are easy to build for several of the lower microwave bands. While we have not worked out design formulas, a table of dimensions for a number of working filters is included. These examples utilize small lengths of any of the common X-band waveguides, including WR-62, which is of otherwise limited usefulness.

Evanescent Mode Waveguide Filters

W1GHZ 2008

Table 1

Designation	Band	Connector Length	Mid Length	Tuning Screw	Bandwidth (measured)
	MHz	mm	mm	USA	MHz
WR-62 Waveguide					
WR62-12-33	5760	12	33	#10	36,37
WR62-12-28	3456	12	28	conc #10	32
WR62-12-28	2304	12	28	conc #10	18.5
WR62-12-28J	2304	12	28	Johanson	18
WR62-12-28J	1296	12	28	Johanson	10
WR62-12-29	2304	12	29	conc #10	20
WR62-10-24	1152	10	24	Johanson	20
WR62-9-22	1152	9	22	Johanson	30
WR-75 Waveguide					
WR75-15-45	5760	15	45	"1/4-28"	29
WR75-13-34	3456	13	34	conc #10	37
WR75-13-32	2304	13	32	conc #10	25
WR-90 Waveguide					
WR90-12-30J	1200	12	30	Johanson	29
WR90-12-30J	1152	12	30	Johanson	28
WR90-12-30J	903	12	30	Johanson	22
WR90-12-30J	581	12	30	Johanson	15
WR90-14-35	2304	14	35	conc #10	(not built)
WR90-15-42	3456	15	42	"1/4-28"	33.5

Note: End Space typically 20 mm

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4. Richard Snyder, “New Application of Evanescent Mode Waveguide to Filter Design,” *IEEE Transactions on Microwave Theory and Techniques*, December 1977, pp. 1013-1021.

Appendix – Filter Tour

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Let's take a short tour of filters, skipping the deep math. But before we can talk about filters, we must start with resonators, the building blocks for filters. Common resonators include LC circuits, transmission line sections, and waveguide cavities, and quartz crystals. There are also mechanical resonators, which could be easier to visualize than the invisible workings of an electrical resonator.

Pluck a guitar string, or tap a suspended pot lid, and an audible tone will be produced for a few seconds. The mechanical resonator has been excited with mechanical energy; the energy is stored as a resonance and slowly released as sound. Good resonators will produce a pure tone for a longer time.

Another example of a mechanical resonator is a pendulum. A good pendulum will swing for a very long time with a constant period determined by its length. The amplitude of the swing will decay slowly, due to friction and air resistance, but the period does not change – the frequency is constant (frequency is the inverse of period). The stored energy is dissipated very slowly. In electrical terms, a pendulum is a high-**Q** resonator: **Q** is defined as the ratio of stored energy to energy dissipated.

To produce something useful, some energy must be extracted from a resonator. The guitar string produces sound to make music, while a pendulum may be coupled to a clock mechanism to tell time. When energy is extracted, the resonator decays faster – the **Q** has been reduced.

If we add energy to the resonator as fast as it is being extracted and dissipated, it can continue indefinitely. We could blow gently on the pendulum each time it starts a downward swing, but we must time it correctly – energy at the wrong frequency may be counterproductive. Of course, we could use a mechanical or electrical signal to time the addition of energy – I have a clock whose pendulum will run for eight days just by winding up a spring and allowing the mechanism to extract a tiny bit of energy from it for each tick.

The electrical equivalent of the clock mechanism is called feedback, adding energy to a resonator to make an oscillator.

Microwave Resonators

Typical microwave resonators are sections of transmission line: odd multiples of an electrical quarter-wavelength shorted at one end and open at the other, or multiples of an electrical half-wavelength either shorted at both ends or open at both ends. The transmission line may be coaxial, with an inner and outer conductor of various shapes, or waveguide formed by a conductor. These transmission line structures are often called cavities. Planar structures on dielectrics are also used. The shapes need not be regular or symmetrical, but odd shapes will complicate calculations.

Whatever the configuration, a single resonator is equivalent to a parallel **LC** circuit (often called a tank circuit for reasons lost in antiquity) like Figure 1. At the frequency where the reactance of the capacitor, $\mathbf{X_C}$, equals the reactance of the inductor, $\mathbf{X_L}$, the circuit is resonant, and it will ring at this frequency if excited by an impulse.

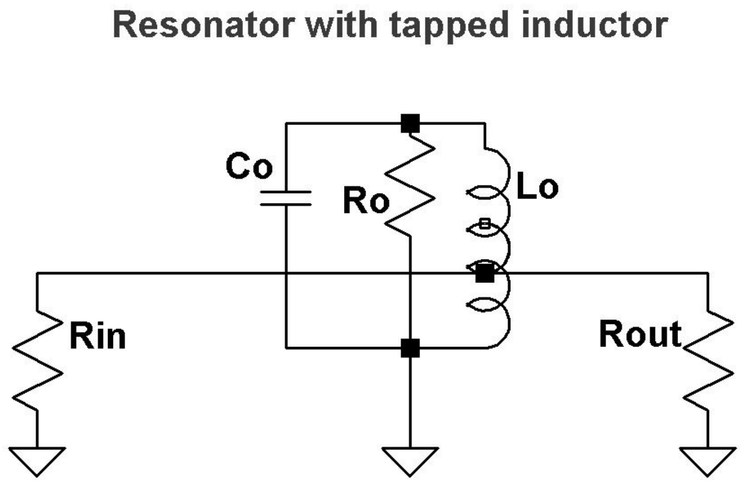


Figure 1

Any real resonator has some intrinsic loss, shown as the **Ro** in the circuit. This loss determines the intrinsic $\boldsymbol{Q}$, or unloaded $\boldsymbol{Q}$, of the resonator: $\boldsymbol{Q_U}$ = **Ro/X**. Since the reactances are equal, either **X** may be used.

When the resonator is connected to a circuit, the resistance added by the circuit appears in parallel with the intrinsic **Ro**, so the total **R** must be lower than **Ro**, reducing the $\boldsymbol{Q}$ to the loaded $\boldsymbol{Q}$, $\boldsymbol{Q_L}$. For example, suppose that we make the connection by tapping down on the inductor ¼ of the turns from the bottom. We might recall that the impedance ratio is the square of the inductor turns ratio; the turns ratio is 4, so the impedance ratio is 16. If we connect a 50-ohm circuit to the resonator, then the added resistance is 16 x 50 = 800 ohms. If the intrinsic **Ro** were 10,000 ohms, then the resultant **R** would be 740 ohms. For an arbitrarily chosen reactance **X** = 200 ohms, $\boldsymbol{Q_U}$ = 10000/200 = 500, while the loaded $\boldsymbol{Q_L}$ = 740/200 = 3.7, a significant reduction.

Selectivity

The selectivity of a resonator is determined by its loaded $\boldsymbol{Q_L}$. The 3 dB bandwidth, the difference between the frequencies where the response is reduced by 3dB, is simply $\mathrm{BW}_3 = \frac{\text{Frequency}}{Q_L}$. Figure 2 makes the effect clear – low $\boldsymbol{Q_L}$ resonators are not very selective, while high- $\boldsymbol{Q_L}$ resonators are quite sharp. The graph is centered at 1 GHz to make it easily scalable to any frequency – for example, the response at 0.8 times the desired frequency is exactly that shown at 0.8 GHz.

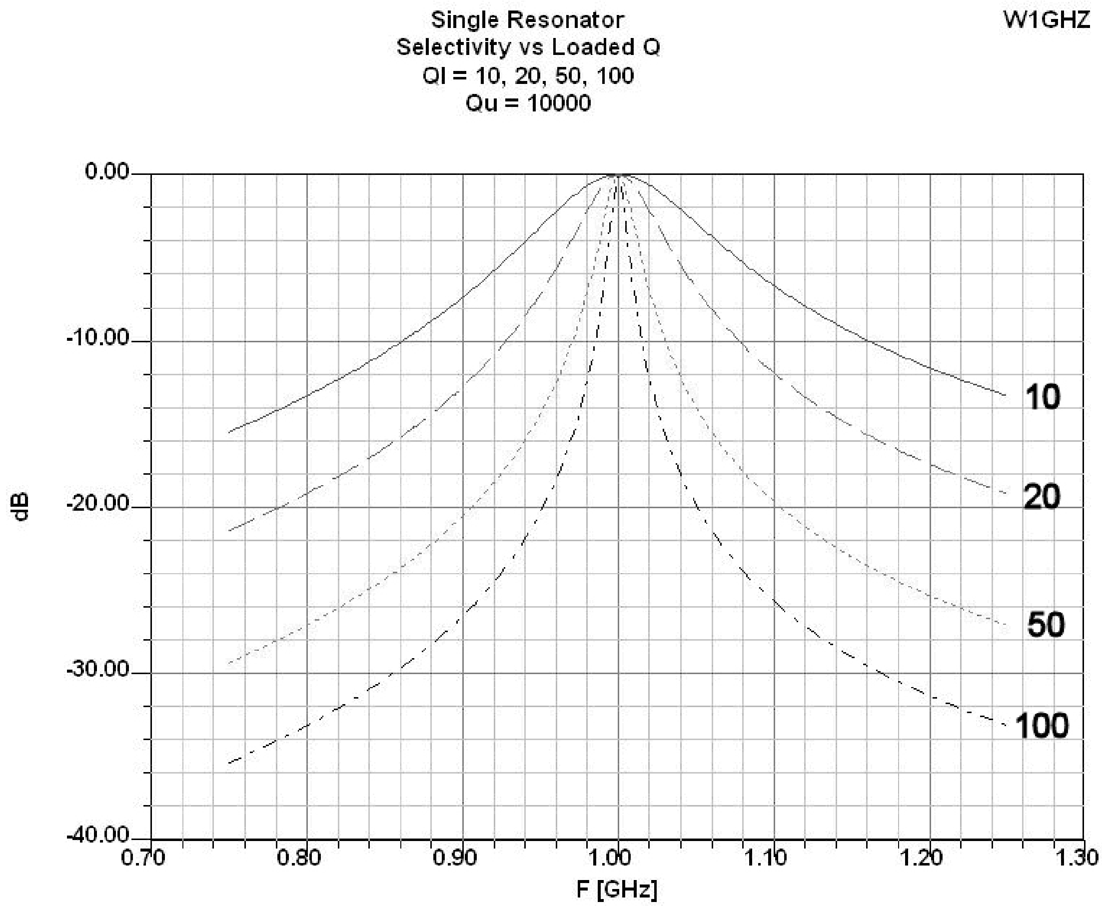


Figure 2

Why not just use high-$\boldsymbol{Q_L}$ resonators? Unless the unloaded $\boldsymbol{Q_U}$ is much higher than $\boldsymbol{Q_L}$, losses will be high, since Ro would be a significant part of the circuit. Some examples are shown in Figure 3 for a $\boldsymbol{Q_L}$ = 100, so that the bandwidth is only 1% of the operating frequency.

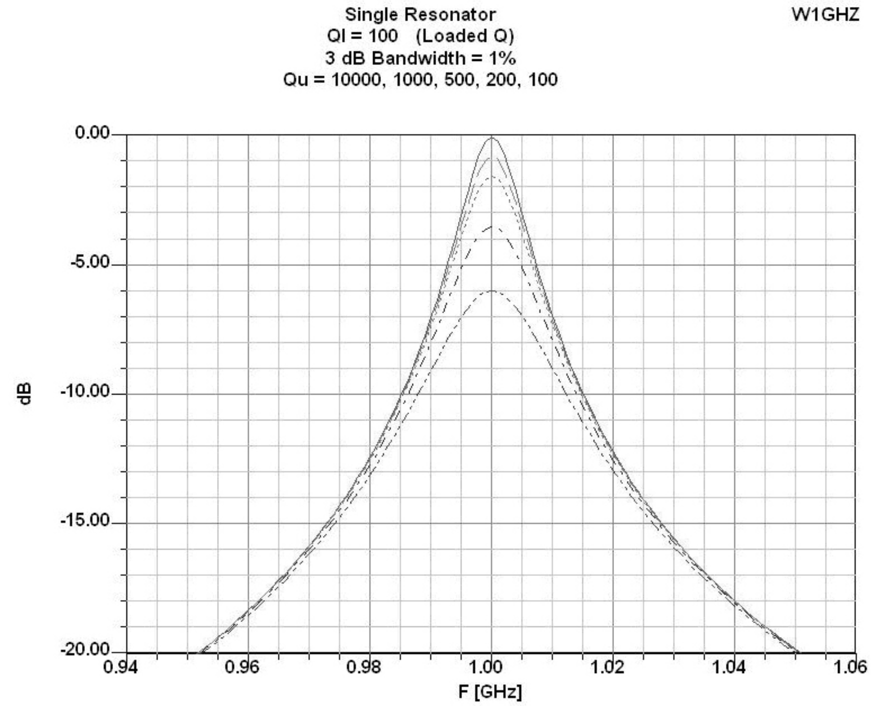
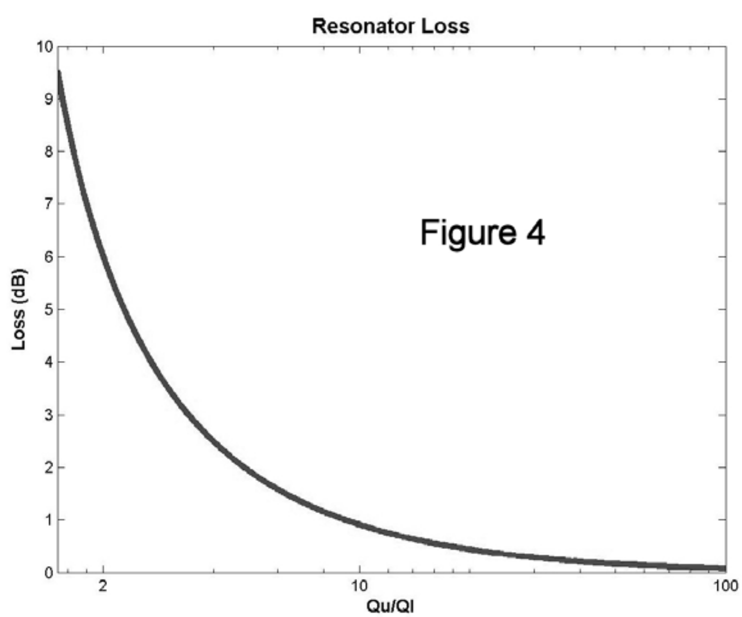


Figure 3

The loss increases rapidly as Q_U decreases. The loss may be calculated[1]:

$$\textbf{Insertion Loss} = \mathbf{20}\log\left(\frac{Q_U}{Q_U - Q_L}\right)\textbf{dB}$$

Figure 4 shows this relationship graphically: when the ratio of Q_U to Q_L is about 10, the loss is about 1 dB. The loss is lower with higher ratios, but loss increases rapidly with lower ratios: when $Q_L = Q_U$, the loss is 6 dB. Trying to make a sharp filter with low Q_U resonators will result in most of the power heating the filter.



At lower frequencies, unloaded Q_U may be improved by increasing physical size, like the large quarter-wave "cavities" used for repeater duplexers. However, at microwave frequencies, when large dimensions are a significant part of a wavelength, additional unwanted resonances will be created in the cavity.

Multiple resonators

A common rule of thumb is that a single capacitor or inductor in a circuit creates a 6dB/octave rolloff. A simple resonator, with one **C** and one **L**, should roll off at 12 dB/octave, where an octave is doubling or halving the frequency relative to the bandwidth. To get faster rolloff, for better out-of-band rejection, therefore requires additional resonators.

Simply connecting resonators together produces interactions which distort the response. A traditional technique, dating back to TRF receivers before the superheterodyne, is to separate resonators with amplifiers to limit interaction between the resonators. At the lower microwave frequencies, where MMIC amplifiers provide cheap gain, we often use two or three simple "pipe-cap" resonators separated by MMIC amplifiers. This combination can provide enough selectivity for good LO and image rejection. The resonators may be synchronously tuned, all at the same frequency, for narrowest bandwidth. Alternately, they may be stagger-tuned, to slightly different frequencies, for a wider passband while still providing fast rolloff.

Modern filter design techniques use multiple resonators, or sections, coupled together to control the interactions and achieve a desired response. By varying the coupling between resonators, the response may be controlled. A simple double-tuned circuit, Figure 5,

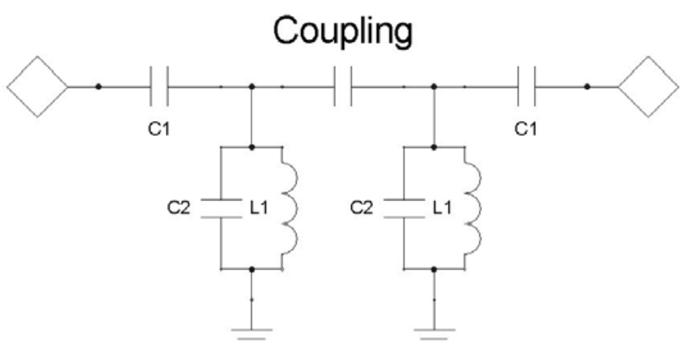


Figure 4

with two coupled resonators, is a good example. Figure 6 shows the effect of coupling: the optimum coupling achieves a flat response; overcoupling increases the bandwidth but creates some ripple in the passband, while undercoupling decreases the bandwidth at the cost of increased loss. Note that all the responses roll off at the same rate outside the passband – only additional resonators will improve the rolloff.

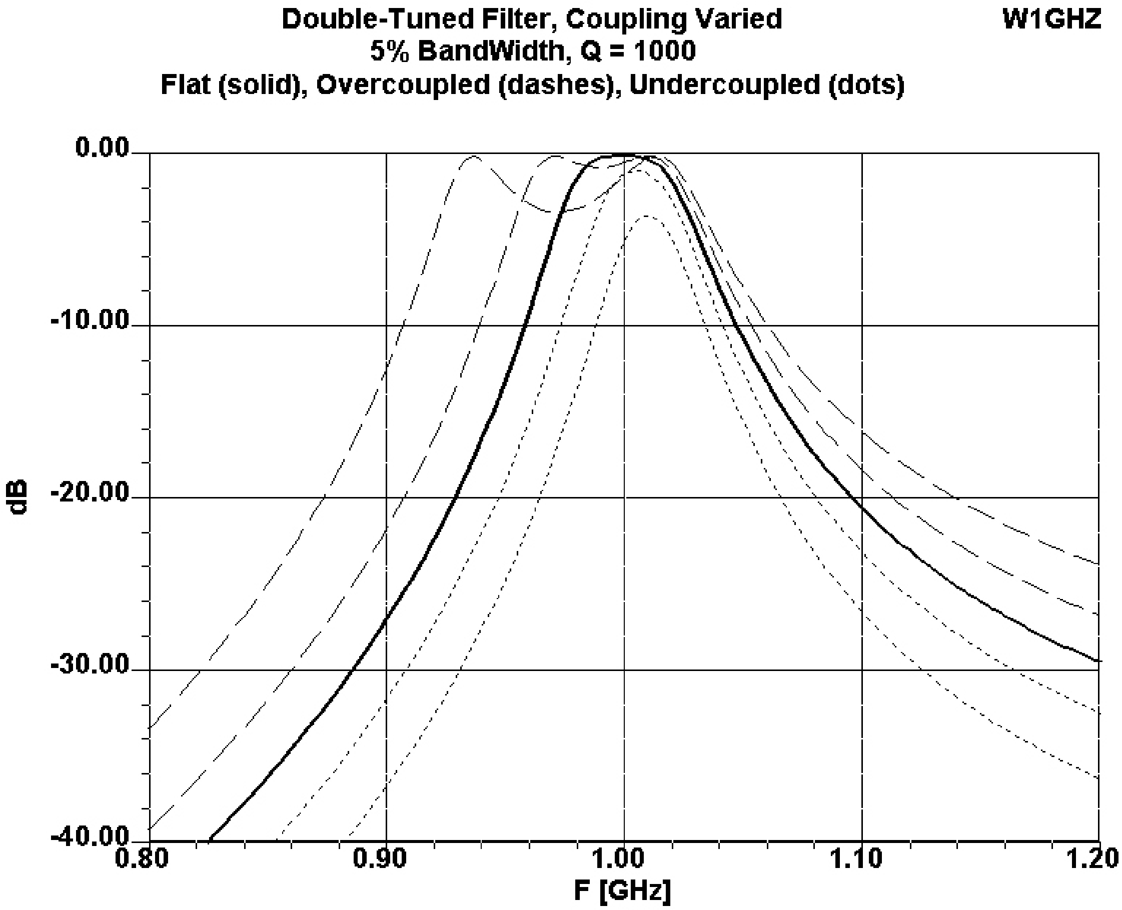


Figure 5

The coupling between resonators may be capacitive, as shown in Figure 5, or inductive, or magnetic, with no physical connection. The input and output connections may also be capacitive, as shown, inductive, either tapped down on the coils like Figure 1 or as a separate winding.

Adding additional resonator sections makes the filtering action much sharper, as shown in Figure 7. A filter may be designed for narrow or wide bandwidth with skirts as sharp as desired. However, the dimensions and tolerances become more critical, and tuning the filter can be much more difficult. I have a six-resonator filter that looks great in the computer design but has proven impossible to tune.

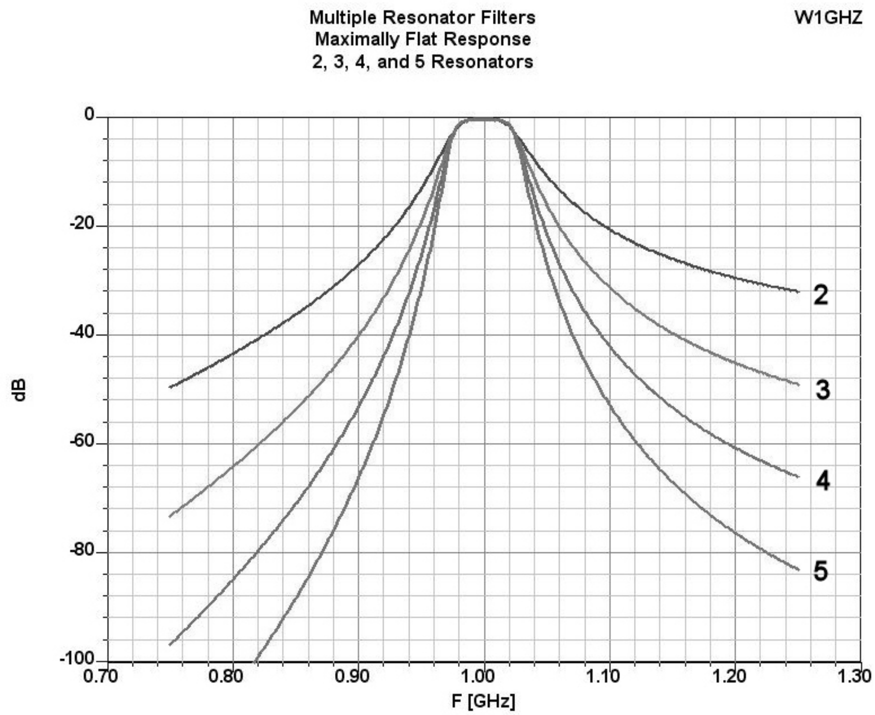


Figure 6

Many commercial applications have stringent filter requirements, to separate channels or to block adjacent bands. These may require a broad passband with low insertion loss, steep skirts that roll off quickly, and high stopband rejection. Figure 8 illustrates these terms. This filter has close to zero dB loss over a broad passband and more than 100 dB of stopband rejection – a good filter can be better than we can measure.

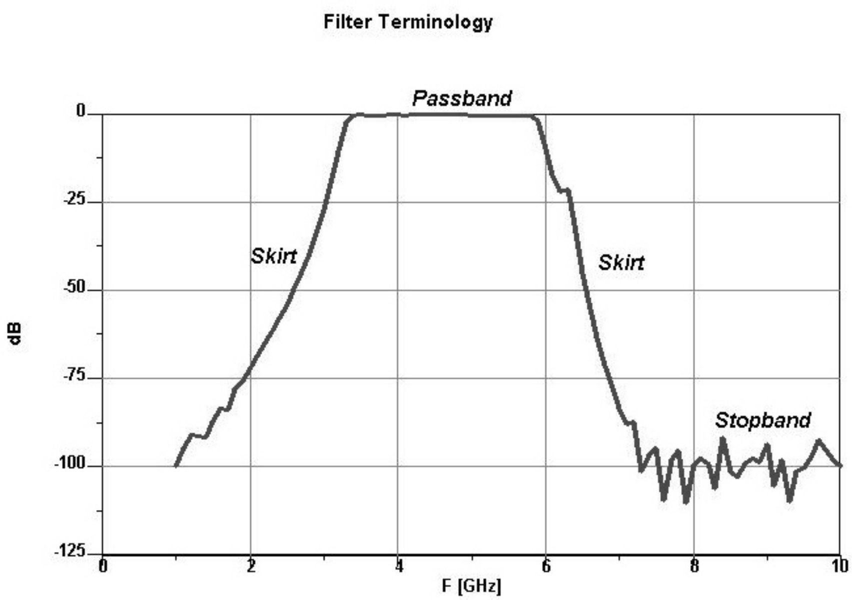


Figure 7

Advanced filter design techniques have been developed to meet these requirements. For instance, a Chebyshev (Чебышёв – also sometimes spelled Chebychev, Chebyshov, Tchebycheff or Tschebyscheff[2]) filter[3] has steeper skirts at the cost of some ripple in the passband loss; the allowable ripple is part of the design procedure. Figure 9 compares a 5-section Chebyshev filter to the Maximally-flat, or Butterworth, design from Figure 7. More advanced filter design techniques, like Cauer, elliptic-function, and cross-coupled filters, offer high performance at the expense of more complex design procedures and difficulty of tuning. Today, filter design software eases the task; traditionally, the design parameters were tabulated in books[4,5,6]. Either way, some engineering is still needed to design a practical filter than can be built.

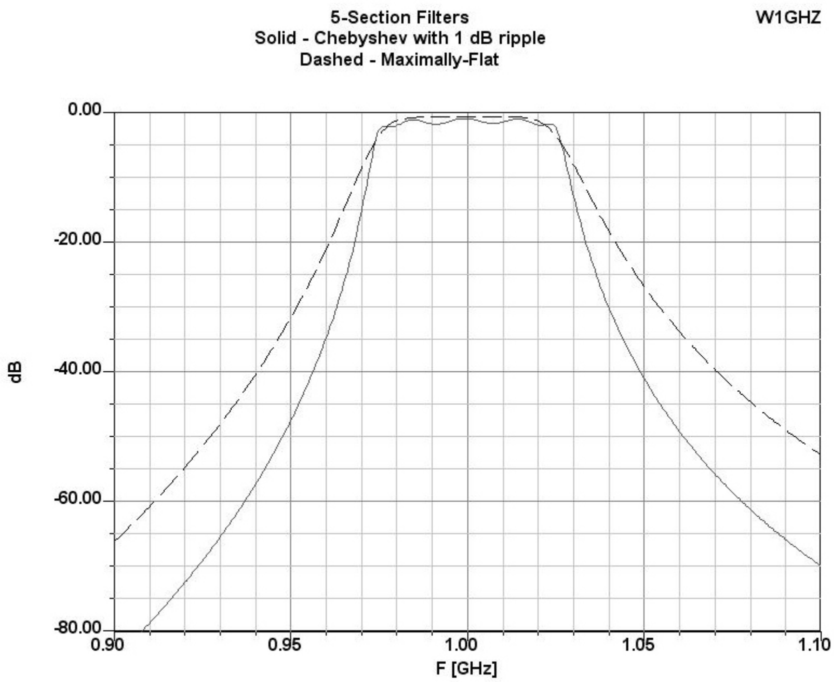


Figure 8

Amateurs rarely need such a fancy filter, except for the crystal filters in our transceivers. Most microwave operation is close to a standard calling frequency, so all that is required is a filter that passes the calling frequency and rejects the conversion image and any LO leakage from the mixer. For the common 144 MHz IF, the ratio of LO frequency to RF frequency is 0.89 at 1296 MHz and 0.937 at 2304 MHz. These ratios may be scaled to 0.89 GHz and 0.937 GHz on Figure 7. For at least 20 dB of LO rejection, a 2-section filter is adequate for 1296 MHz, while a 3-section filter may be needed for 2304 MHz. For higher bands, we need either a sharper filter or a higher IF frequency, like 432 MHz.

A sharp filter may be either very narrow or have more sections. How narrow a filter we can use is limited by several factors. The first is the unloaded $\boldsymbol{Q}$, $\boldsymbol{Q}_U$, of the resonators – we can't make a narrow, high-$\boldsymbol{Q}$ filter with low-$\boldsymbol{Q}$ resonators. A more practical limit whether we can tune a narrow filter *and* have it stay tuned over temperature and vibration, particularly for rover operation. The alternative, adding more sections, also has problems. Unlike the ideal filters in Figure 7, each section adds additional loss, so that the filter loss is proportional to the number of sections. Also, filters with more than 3 or 4 sections are very difficult to tune properly without sophisticated test equipment.

With high-***Q*** resonators, like those found in waveguide filters, a narrow double-tuned circuit, or two-section filter, should be satisfactory for many amateur applications. We can see from Figure 6 that the skirt selectivity is not affected by the coupling, so the Maximally-flat, or Butterworth, type is a good choice. The bandwidth of a two section Butterworth filter is $\sqrt{2}$, or 1.414, times the bandwidth of a each single resonator. Thus, to find the desired $\boldsymbol{Q_0}$, the loaded ***Q*** of each resonator, we simply calculate

$$Q_0 = \frac{\text{Frequency}}{\text{BW}_3} \times \sqrt{2}\ .$$

For circuits with discrete **L** and **C**, the coupling components are easily calculated[7]. However, for direct-coupled resonators like those in waveguide filters, we must rely on tables[6] or programs like **WGFIL**[8]. More important, we can estimate loss using Figure 4 if we have an idea of the unloaded ***Q***, $\boldsymbol{Q_U}$, of the resonators.

Different types of filter construction are available, each with advantages and disadvantages. Waveguide filters can have extremely high ***Q***, so that narrow filters are possible with very loss. However, at lower frequencies, they become very large. Printed circuit filters have low-***Q*** resonators, but are cheap and repeatable, requiring no tuning, so they may be preferred at lower frequencies where gain is cheap. Other possible choices include helical, interdigital[9], and combline filters, each offering different tradeoffs in loss, size, and difficulty in design and construction. It is a matter of choosing an adequate filter for each application.

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