Easy Microwave Filters Using Waveguides and Cavities

The authors show us how to design filters for a microwave station.

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Waveguide band-pass filters are frequency-selective devices that perform valuable functions in microwave equipment used in communications. They are most compatible with waveguide antenna feeds and are required for high-power applications and precision performance. They are primarily used at frequencies from 8 GHz to more than 100 GHz.

The main function of a waveguide signal filter is to provide adequate stop-band selectivity without introducing unacceptable insertion loss and frequency distortion. When constructed, the modeled prototype ideally achieves all the desired characteristics; but in practice, this does not automatically happen. Prior to the PC-simulation era, the recourse was to build and modify a prototype to achieve the desired filter characteristics. This is still a good way for amateurs to build many filters, as in the microwave field, when using small copper cavities realized using standard "pipe-cap" techniques.[1] See Figure 1.

Cavity Filters

The simplest filter for microwave is a single-cavity resonator. A cavity enclosed by metal walls has an infinite number of natural frequencies at which resonance will occur. Different resonance modes can be used with a cavity.[2] Most interesting for Amateur Radio use is the TM_{010} mode, where the resonant frequency depends only on the inner diameter (D) of the cavity. A practical formula for the resonant frequency is $f = 229.5 / D$, where D is in mm, and f is in GHz. Figure 2 shows the ***E*** (electric) and ***H*** (magnetic) fields in cylindrical cavities resonating in modes TM_{010} and TE_{011}.

We can see also from the Figure 3 graph that, with $D / L = 1$, high values of Q are theoretically possible; but that is not compatible with the necessity of avoiding spurious modes (as the TE_{111}) because the adjustment screw normally used (see example in Figure 1B) acts in a different manner for modes TM_{010} and TE_{111}.

As an example, with $L = 14$ mm (cylindrical cavity height) and $D = 20$ mm we have: $f_{TM010} = 11.48$ GHz and $f_{TE111} = 13.85$ GHz, with the spurious mode TE_{111} only 2.37 GHz away. Unfortunately, the highest Q of resonator cavities can't be reached when using a

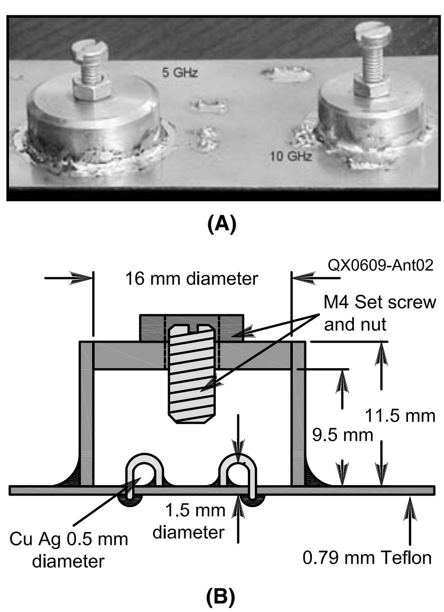

Figure 1 — Part A shows a typical pipe-cap resonator. Part B shows the dimensions for one example of a 10-GHz resonator.

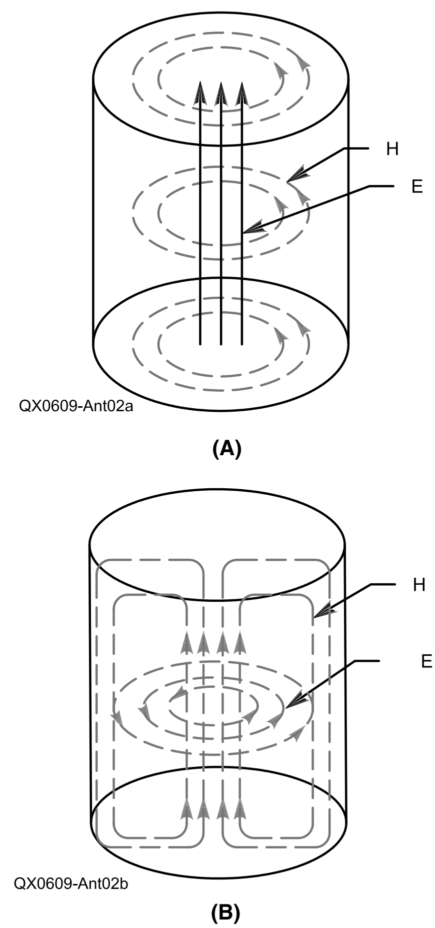

Figure 2 — Part A shows electric and magnetic fields in a TM_{010} cavity resonator. Part B shows electric and magnetic fields in a TE_{011} cavity resonator.

[1]Notes appear on page 42.

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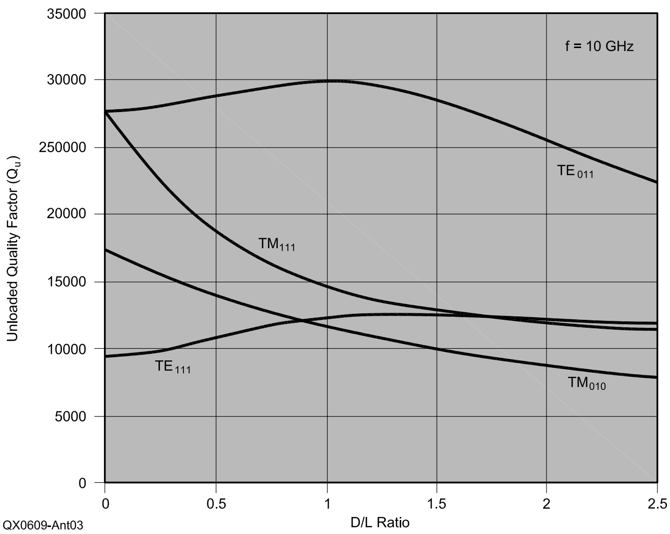

Figure 3 — Unloaded Q of microwave cavities.

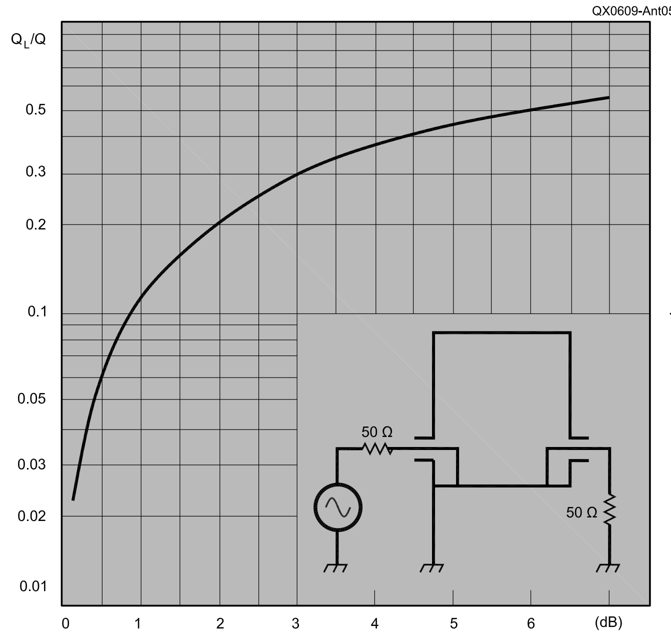

Figure 5 — Ratio of loaded to unloaded Q versus insertion loss in a cavity.

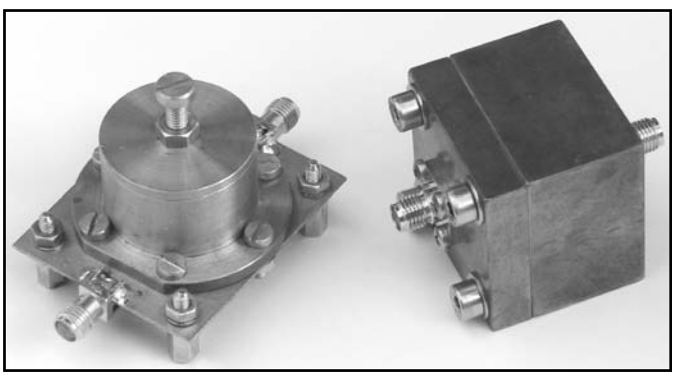

Figure 4 — Two TM_{010} laboratory cavities for "Q" tests.

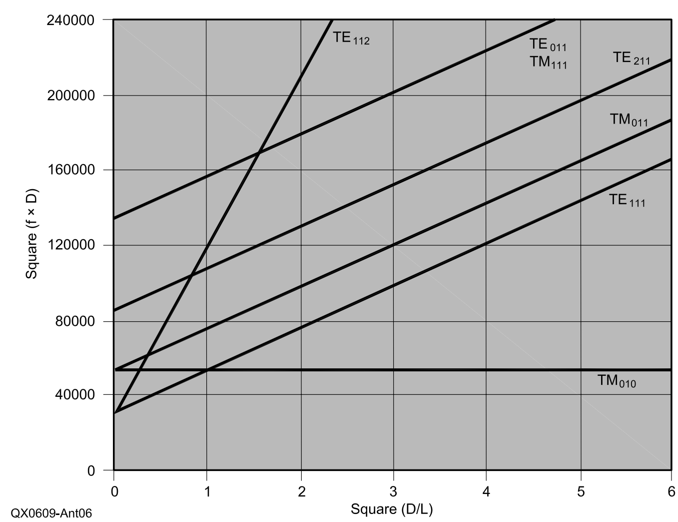

Figure 6 — Mode chart for microwave cavities (frequency in GHz and D&L in mm).

small cavity height, and therefore the selectivity of the filters is lower. A suggested, but expensive, way is to use a precision TM_{010} machined cavity with a tuning screw limited to the Fine Tuning function. In a copper cavity with D = 21.5 mm and L = 14 mm the resonance (TM_{010}) is at ~10.7 GHz with a calculated $Q_U \approx 9500$ (see Note 2). The nearest TE_{111} spurious mode is at about 13.5 GHz.

We have measured the Q and bandwidth of the two TM_{010} copper cavities of Figure 4 with $D = 21.5 \pm 0.05$ mm and L = 14 mm. The ratio between the loaded Q (Q_L) and unloaded Q (Q_U) starting from the resonant bandwidth and the insertion loss of the single cavity with 50-Ω attenuators at the input and output was obtained as shown in Figure 5. For the first cavity (left in Figure 4) excited using small probes (about 2 mm height) or loops on the printed circuit, we measured (with specific microstrip circuits) about 0.8 dB insertion loss because of the feed lines only, and a very bad final unloaded Q_L of 450.

For the second cavity (right in Figure 4) the driving probes are realized using microwave SMA connectors with *only 1.0 mm* of pin height. The measured insertion loss was only 0.6 dB and the resonance bandwidth = 31.6 MHz, with a final Q_L = 336 and Q_U = 4804. This is a very interesting value for a TM_{010} cavity where the theoretical maximum Q_U (perfectly polished internal walls) is about 9500. The measured resonant frequency was 10.627 GHz.

It is useful to be able to predict the spurious mode frequencies to gain insight into the spurious performance of a cavity filter. The easiest way to predict these modes is to use a Mode Chart as shown in Figure 6.[3] The dominant mode of a cavity is the lowest frequency at which it can resonate.

To prevent the secondary (or spurious) mode resonances, the TM_{010} mode cannot be excited ove the entire D / L range reported in Figure 3, but only in the part of the graph were the ratio $D / L > 1$. At $D / L \leq 1$, the dominant mode becomes TE_{111}. Practical values of the D / L ratio to be used are 1.4 to 2.4 for TM_{010} and 0.45 to 0.5 for TE_{111} modes. Frequently, the tuning screw is very long and the cavity is reduced to a standard coaxial resonator in T_{EM} mode.

Another interesting mode is TE_{011}, which is not a dominant mode, so care must be used to choose a coupling system that does not

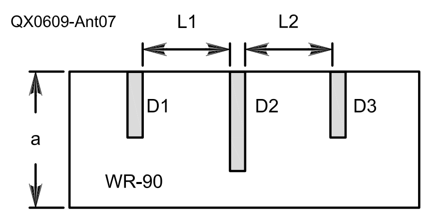

Figure 7 — Layout of a simple two-pole microwave iris filter.

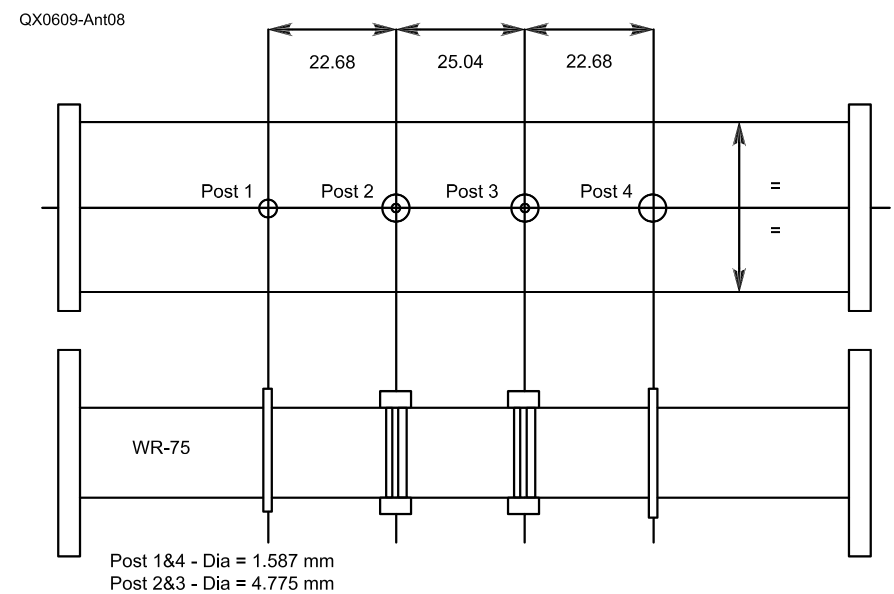

Figure 8 — Mechanical data of a simple "no-tune" post filter with a WR-75 guide.

excite the other possible modes that could resonate within the frequency tuning range of the cavity.

To couple power into or out of a resonant cavity, either waveguide or coaxial loops, probes, or apertures may be used. An inductive loop is inserted in the resonator at a desired point where it can couple to a strong magnetic field. The degree of coupling may be controlled by rotating the loop. A capacitive probe is inserted in the resonator at a point where it is parallel to, and can couple to, a strong electric field. The degree of coupling is controlled by adjusting the length of the probe.

The TE_{011} mode is very interesting for its very high quality factor Q_U that is about *three times greater* than that of TM_{010} or TE_{111} cavities. Another advantage of TE_{011} is that there is no axial current. This means that the end plate can be free to move to change the resonant frequency without significant loss due to the current that is parallel to the circular caps of the cylinder. With $D / L = 1.5$, the diameter of the TE_{011} cavity is about two times that of a TM_{010} cavity. For a more professional filter design in X-band, the solution is probably a waveguide filter.

Waveguide Band-Pass Filters

There are several types of filters that can be used to perform the job of separating RF and LO or image frequencies, but most are either difficult to construct or have losses that are unacceptable. The waveguide approach is attractive from the construction point of view and, at 10 GHz, is relatively simple and small.

Waveguide band-pass filters originated in Bell Telephone Laboratories and the MIT Radiation Laboratory of World War II. In 1957, a milestone paper introduced direct-coupled waveguide band-pass filters that quickly evolved into a preferred configuration.[4] For the waveguide filter structures and design see also a classic reference from 1964 (mentioned in the text of Note 3), the RSGB Microwave Handbook, Vol. 2, and an interesting method of verifying the realized filter on-line.[5,6] Typical band-pass structures consist of propagating sections of standard waveguide (WR-75 and WR-90 for the 10-GHz band) acting as half-wavelength resonators, which are coupled via discontinuities in the guide such as iris (see Figure 7) or posts.

Filters Using Posts

For the filters using posts, we include only a three-pole 10.315 GHz example from a W1VT design.[7] The most important mechanical dimensions of this no-tune filter using four inches of WR-75 copper waveguide are shown in Figure 8. The measured performance is very interesting and our proposed idea is for a quasi-no-soldering solution. This is very important because the soldering phase in waveguide is not simple and not without problems. The tested filter is shown in Figure 9.

The main advantage of the no-tune design is the ease of home construction without the need for expensive test equipment. Reliability is also enhanced because of the impossibility of going out of alignment (no screws) if the filter is a part of a transverter at the top of a tower submitted to vibration. Remember that this kind of filter must be carried out with a suitable mechanical accuracy (better than 0.1 mm) to meet performance goals.

To receive the frequency $f_0 = 10.368$ GHz and using $f_{IF} = 144$ MHz, the LO frequency becomes $f_{LO} = f_0 - f_{IF} = 10.224$ GHz, since most people prefer a lower LO frequency. In the case of our filter, composed of three waveguide resonators (n = 3), if an attenuation A = 30 dB is requested at the image frequency, the bandwidth can be calculated by:

$$BW = (2 f_{IF}) / (10^{A / (20 n)}) = 91 \text{ MHz} \quad \text{(Eq 1)}$$

Our test filter was designed *for laboratory use only* at *fo* = 10.315 GHz to verify the no-tune precision.

The measured performance was:
Insertion Loss = 1.4 dB
SWR < 1.3:1 in-band
BW = 74 MHz (–3 dB) and 250 MHz (–20 dB)

For the complete frequency response, see Table 1. Two homemade waveguide WR-75 to SMA adapters are used for testing the waveguide filter. If the mechanical errors are < 0.1 mm (post positioning), the results are quite good (calculated $f_o = 10.315$ GHz and measured $f_o = 10.294$ GHz).

Iris-Coupled Filters

In waveguide filters that employ iris-coupled resonant sections, we see a pattern similar to that of capacitively coupled resonator filters. If, in rectangular waveguide capable of propagating the dominant mode only,

Table 1
Measured Response of the No-Tune Three-Pole Filter Designed for f_o = 10.315 GHz

Frequency (GHz)	*Attenuation (dB)*
10.166	–20
10.228	–10
10.248	–6
10.266	–3
10.273	–2
10.282	–1
10.300	0
10.328	–1
10.334	–2
10.340	–3
10.346	–6
10.368	–10
10.414	–20

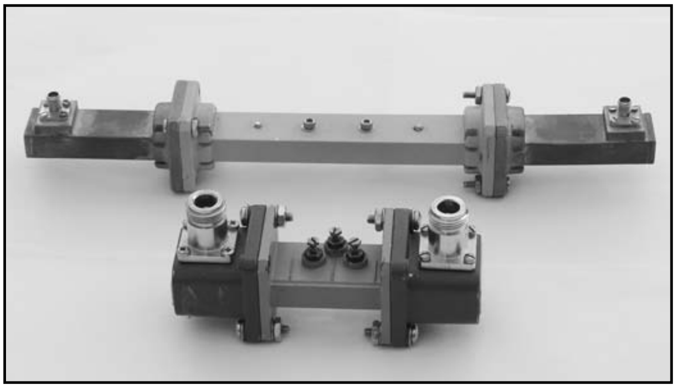

Figure 9 — Photo of the two waveguide band-pass filters (no-tune and iris).

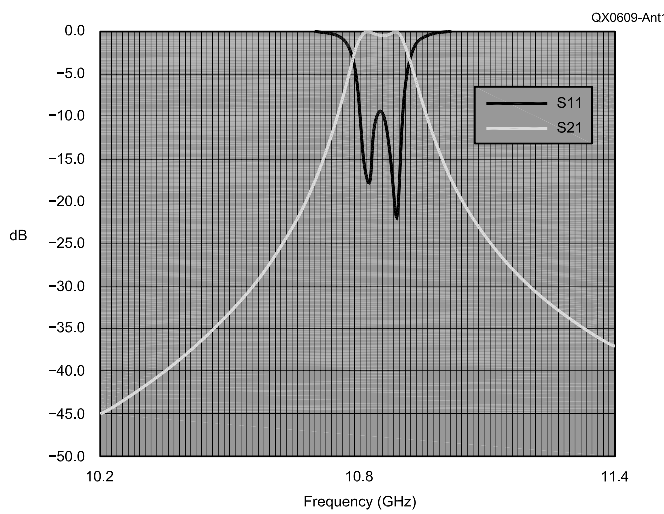

Figure 10 — On-line calculated response of the two-pole filter.

a thin metal partition is inserted in such a way that the edge of the partition is parallel to the electric field, the iris so formed is equivalent to a shunt inductance.[8] The value of the susceptance has been accurately calculated.[9, 10]

The waveguide can be cut with a saw, and the iris material soldered on, before plating. As with other metallic filters, it is not unusual for a designer to take advantage of the opportunity to have tuning screws to accommodate small tolerance variations in the filter.

The unloaded *Q* may be substantially reduced by losses in the loading capacitance. If the capacitance is realized as a simple brass screw, the threads exposed within the waveguide should be removed and the remaining smooth metal should be plated with a highly conductive material such as copper, silver or gold. A low-resistance contact should be insured at the grounded end of the tuning screw.

The layouts of Figure 7 and Table 2 are used as an example to verify, via the Web, the performance of the filter. In our case, the filter is designed for a 10.368 GHz nominal frequency, but the on-line software gives a 10.8 GHz center frequency because the tuning screws are not considered. After the final adjustment the filter is tuned exactly at 10.368 GHz. From the layout data in Figure 7, it's clear that the filter is asymmetric: realized with iris reactances only in one side of the waveguide and with a guide WR-90, where side *a* is 22.86 mm. The Java real-time software available free at Guided Wave Technology (see Reference 6) uses a mode-matching simulation method with 13 modes considered in the analysis. The frequency response and the input/output matching (S11 = S22) of the filter are shown in Figure 10.

Design of a Two-Pole Filter Using Asymmetrical Iris

The first things to be defined are the lower and the upper frequencies at which the filter must operate. To define the filter working frequencies, the values will be 3 to 5% higher to permit fine tuning using suitable screws placed at the center of each resonator. In our case the two frequencies are: $f_1 = 10.805$ GHz and $f_2 = 10.895$ GHz. The tuning screws, to shift the resonant frequency, are simple brass M4 machine screws.

At this point, we select the waveguide to be used: WR-90, with a cut-off frequency of 6.56 GHz (λ_c = 45.72mm), is a good candidate to satisfy our requirements (see Table 3).

We have all the elements to calculate both the free-space and the corresponding guide wavelengths:

$\lambda_1 = 299.8 / f_1 = 27.75$ mm
$\lambda_2 = 299.8 / f_2 = 27.52$ mm

$$\lambda_{g1} = \frac{\lambda_1}{\sqrt{1-\left(\frac{\lambda_1}{\lambda_c}\right)^2}} = 34.91 \text{ mm}$$

$$\lambda_{g2} = \frac{\lambda_2}{\sqrt{1-\left(\frac{\lambda_2}{\lambda_c}\right)^2}} = 34.46 \text{ mm}$$

where f_1 and f_2 are in GHz; λ_1, λ_2 and λ_c in mm.

Table 2
Design Data for the 10.4-GHz Two-Pole Iris Filter

a	Waveguide Width (mm)	22.8
W	Iris Thickness (mm)	0.5
d_1	Iris no. 1 Length (mm)	13.2
L_1	First Resonator Length (mm)	16.4
d_2	Iris no. 2 Length (mm)	16.5
L_2	Second Resonator Length (mm)	16.4
d_3	Iris no. 3 Length (mm)	13.2

Table 3
Waveguide Data (Standard WR-90 and WR-75)

Waveguide Designation	*Inner Dimensions (mm)*	*Dominant Mode*	*λ_c (mm)*	*f_{cutoff} (GHz)*	*Frequency Range (GHz)*
WR-90	22.86 × 10.16	TE_{10}	45.72	6.56	8.2 – 12.5
WR-75	19.05 × 9.53	TE_{10}	38.10	7.87	9.8 – 15.0

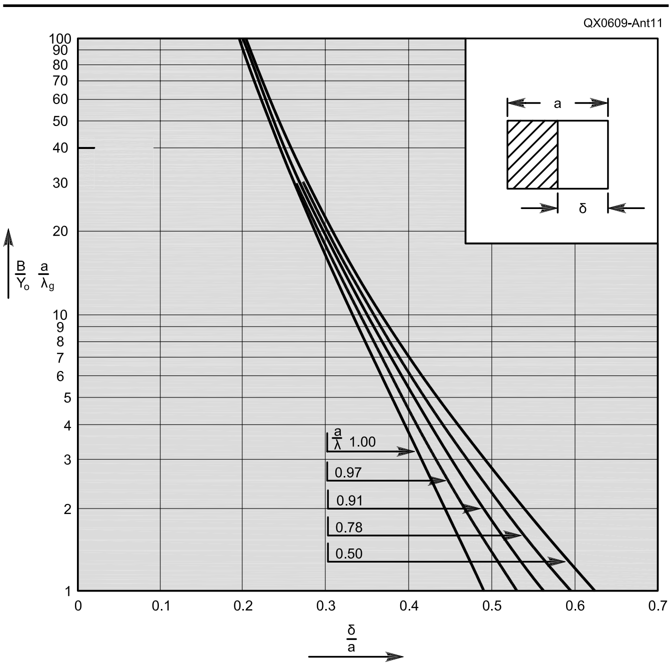

Figure 11 — Susceptance of asymmetrical iris in waveguide.

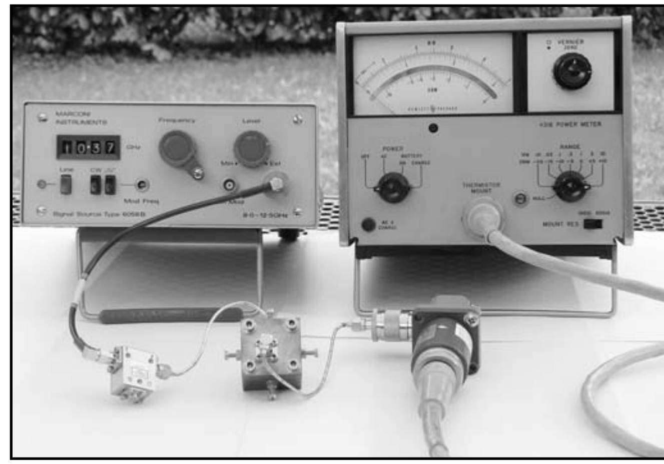

Figure 14 — Test setup: a Marconi 6058B generator, a Narda 7 to 12 GHz circulator and the HP 431B power meter plus thermistor mount

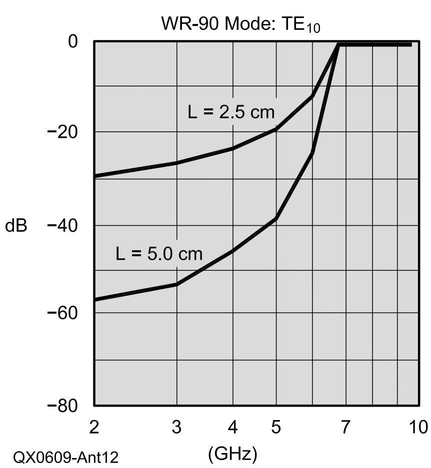

Figure 12 — Calculated response of a high-pass filter using WR-90 in the cutoff zone.

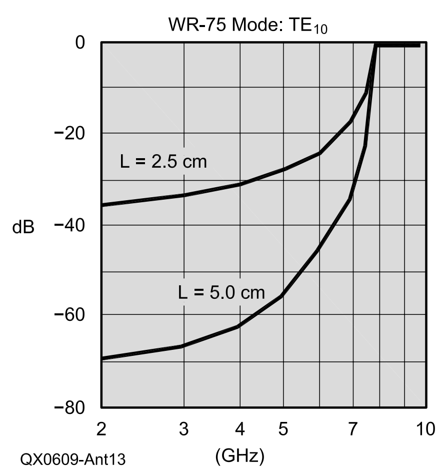

Figure 13 — Calculated response of a high-pass filter using WR-75 in the cutoff zone

The center waveguide wavelength can be computed by the following:

$\lambda_{g0} = (\lambda_{g1} + \lambda_{g2}) / 2 = 34.68$ mm (Eq 2)

This corresponds to the free-space wavelength $\lambda_0 = 27.63$ mm.

Now we determine the fractional bandwidth from the waveguide wavelength with the equation:

$\omega_\lambda = (\lambda_1 - \lambda_2) / \lambda_0 = 0.01307$ (Eq 3)

It is not the same thing as the fractional bandwidth calculated in terms of frequency. This is related to the dependence of the guide wavelength to the guide physical dimensions.

At this point, we define the filter response required by our application. We assume to realize a two-pole filter (N = 2) that will work with a 0.2-dB-ripple Chebyshev response. Table 4 gives the g-value normalized elements that allow us to calculate the coupling coefficients. The values to be used are: $g_0 = 1$, $g_1 = 1.0378$, $g_2 = 0.6745$, $g_3 = 1.5386$:

$$K_{01} = \sqrt{\frac{\pi\omega_\lambda}{2 g_0 g_1}} = 0.14064 \quad \text{(Eq 4)}$$

$$K_{12} = \frac{\pi\omega_\lambda}{2\sqrt{g_1 g_2}} = 0.02453 \quad \text{(Eq 5)}$$

$$K_{23} = \sqrt{\frac{\pi\omega_\lambda}{2 g_2 g_3}} = 0.14064 \quad \text{(Eq 6)}$$

Note that, in case of a filter with three or more cells, the equation to be used for the whole central element is similar to the one of the second coefficient K_{12} (with a suitable index).

Then we calculate the normalized reactances, to $Z_0 = 1\ \Omega$, needed at each iris plate position.

The results, without the relevant units, are pure numbers due to the normalization:

$X_{01} = X_{23} = K_{01} / (1 - K_{01}^2) = 0.14348$ (Eq 7)

$X_{12} = K_{12} / (1 - K_{12}^2) = 0.02455$ (Eq 8)

We can see that $X_{01} = X_{23}$ and so there is no need to work with the whole values, but only with the first two. The asymmetric iris (layout of Figure 7) appears as an inductor across the guide. After the calculation of the reactances, we are able to compute the spacing between iris into the waveguide (about a λ / 2 cell) both in terms of electric angle (radians) and in millimetres:

$\theta_1 = \theta_2 = \theta = \pi - [\arctan(2\,X_{01}) - \arctan(2\,X_{12})] / 2 = 3.0264$ radians (Eq 9)

$L_1 = L_2 = (\theta\lambda_{g0}) / (2\,\pi) = 16.71$mm (Eq 10)

The iris sizes are obtained through the normalized susceptances:

$B_{01} = 1 / X_{01} = 6.97$ (Eq 11)

$B_{12} = 1 / X_{12} = 40.73$ (Eq 12)

With the graph of Figure 11 for the iris design, we calculate the value of the ordinate:[10]

$(B_{01}\,a) / \lambda_{g0} = 4.59$ (Eq 13)

$(B_{12}\,a) / \lambda_{g0} = 26.85$ (Eq 14) and the ratio $a / \lambda_0 = 0.83$ for a correct choice on the diagram:

$$\delta_1 / a = \delta_3 / a = 0.42 \qquad \text{(Eq 15)}$$

$$\delta_2 / a = 0.28 \qquad \text{(Eq 16)}$$

The asymmetric iris (0.5 mm metal sheet) lengths can be obtained by:

$$d_1 = d_3 = a\,(1 - 0.42) = 13.26 \text{ mm} \qquad \text{(Eq 17)}$$

$$d_2 = a\,(1 - 0.28) = 16.46 \text{ mm} \qquad \text{(Eq 18)}$$

At the end of the computation phase, the filter was also simulated using the on-line software (see Note 6). The relevant results are reported in Figure 10: f_0 = 10.75 GHz, BW = 104 MHz at –3 dB, insertion loss < 1 dB, return loss > 15 dB, attenuation > 25 dB at $f_0 \pm 250$ MHz. From our laboratory tests : f_0 = 10.368 GHz, BW = 98 MHz at –3 dB. The differences between simulation and measurements are related to the tuning screws that cannot be simulated.

In Figure 9, note the two band-pass filters analyzed in the paper [three-pole with posts (rear) and iris two-pole (front)] with the waveguide-to-coax adapters (WR-90 and WR-75) used in the tests.

High-Pass Filters

A very easy method to realize a high-pass filter consists in the use of a standard waveguide, of defined length, in the cut-off region. If the involved guide has a different size than the used one, suitable adapters are used to match source and load.

The equation that describes the shape of the waveguide attenuation, in decibels, under cutoff frequency is:

$$A = 54.58 \frac{d}{\lambda_c} \sqrt{1 - \left(\frac{\lambda_c}{\lambda} \right)^2} \qquad \text{(Eq 19)}$$

where λ_c is the cutoff, λ the working wavelength and d is the guide length. To apply Equation 19 correctly, all dimensions must be expressed using the same units. Take note that the total attenuation obtained (in dB) is directly proportional to the guide length.

The cutoff wavelength can be calculated using the formula applicable to the rectangular guide:

$$\lambda_c = 2\,a \qquad \text{(Eq 20)}$$

where a is the largest dimension of the guide cross section considering operation in the TE_{10} mode. A new equation is applied for the circular waveguides operating in the TE_{11} dominant mode:

$$\lambda_c = 3.412\,a \qquad \text{(Eq 21)}$$

where, in this case, a becomes the waveguide radius.

Two examples are shown in Figure 12 using a WR-90 waveguide (X-band = 8.2 to 12.4 GHz) and in Figure 13 using a WR-75 (9.8 to 15 GHz). See also Table 3. The considered lengths are about one and two inches (25 and 50 mm).

Table 4
Chebyschev 0.2-dB Ripple

N	g_0	g_1	g_2	g_3	g_4
1	1.000	0.4342	1.000		
2	1.000	1.0378	0.6745	1.5386	
3	1.000	1.2275	1.1525	1.2275	1.000

Simple Measurements

In the photo of Figure 14, see the simple setup used for our measurements, which includes a Marconi 6058B microwave generator and a well known HP 431B power meter with waveguide thermistor mount X486A. Two waveguide to coaxial adapters, a 7 to 12 GHz SMA Narda isolator and a fixed 7 dB SMA attenuator are the only other components of the tests circuit. An HP waveguide attenuator model X382A and a 10 dB waveguide directional coupler model X752C (with 40 dB directivity) was used to analyze the input /output matching. A high-stability counter with 13-GHz prescaler is used to obtain high-precision frequency measurements. All these components are available today at very low cost on the surplus market.

Appendix: Cavity Resonating Modes

The general classification of transverse electric (**TE**) and transverse magnetic (**TM**) modes is true both for cylindrical and rectangular cavities. The denomination derives from the one already used for the waveguide. In the transverse electric mode, the entire electric field is in the transverse plane, which is perpendicular to the length of the waveguide (direction of energy travel). Part of the magnetic field is parallel to the length axis. In the transverse magnetic mode, the entire magnetic field is in the reverse plane and has no portion parallel to the length axis.

In the rectangular waveguide resonators the modes are named $\mathbf{TE}_{mnp}$ or $\mathbf{TM}_{mnp}$ where the integer "m" designates the number of half waves of electric or magnetic field in the "x" direction (the bigger dimension of waveguide cross section), while "n" denotes the number of half cycles in the "y" direction (the smaller dimension of waveguide cross section) and "p" indicates the number of half wave in the "z" direction (perpendicular to the waveguide cross section).

In cylindrical resonators the same type of index system is used, but changes the meaning of its subscripts: "m" represents the number of half waves variation along the circumference that constitutes the base of the cylinder (as a function of its angle at the centre), "n" express the number of half cycles change along the base radius direction, "p" define the number of half wave change along the cylinder symmetry axis.

Notes

[1]H. J. Meise, DK2AB, *Complementation of Cylindrical Resonator Cavities*, Dubus, 2/1987, pp 95-97.

[2]*Reference Data for Engineers*, SAMS, Seventh Edition, 1989, pp 30-20 to 30-22.

[3]G. Matthaei, L. Young and E. M. T. Jones, *Microwave Filters, Impedance-Matching Networks and Coupling Structures*, Artech House, 1980.

[4]S. B. Cohn, "Direct-Coupled Resonator Band-Pass Filters," *Proc of IRE*, pp 187-195, Feb 1957.

[5]*RSGB, Microwave Handbook*, 1991, Vol. 2, pp 12.23 to 12.27.

[6]**www.guidedwavetech.com/IRISHTML/iris2p.htm**

[7]Zack Lau, W1VT, "A No-Tune 10 GHz Filter", *QEX*, May/Jun 2000, pp 52-55.

[8]C. G. Montgomery, Principles of Microwave Circuits, 1948, pp 163-167.

[9]N. Marcuvitz, *Waveguide Handbook*, 1951

[10]*Microwave Journal Handbook*, Feb 1968, pp 26.

Paolo Antoniazzi was licensed as IW2ACD and became a member of the Associazione Radiamatori Italiani (ARI) in Italy in 1961. He worked for Siemens, GTE, Sprague and ST Microelectronics in the fields of RF and telecommunications marketing and applications until his retirement in November 2003. Paolo is particularly interested in simulation and measurements (and writing papers) about narrow-band LF, 2.4 GHz, and more recently, the 10 GHz band. Space communications with the future P3E and P5A satellites is a priority interest for 2006 - 2007.

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