Tunable Microwave and Millimeter-Wave Band-Pass Filters

Jaroslaw Uher, *Member, IEEE,* and Wolfgang J. R. Hoefer, *Fellow, IEEE*

Abstract —**This paper presents an overview of tunable microwave and millimeter-wave band-pass filters realized in different technologies. Some general design principles are described. Recent progress in the performance of various tunable filters is reported. The paper surveys magnetically tunable filters (ferrimagnetic resonance filters, MSW filters, evanescent waveguide filters, *E*-plane printed circuit filters), electronically tunable filters, and mechanically tunable filters. The typical performance parameters are summarized and compared in terms of suitability for different applications.**

I. Introduction

TUNABLE band-pass filters can be realized in many techniques, but whatever the method of tuning may be, they must conserve as much as possible their transmission and reflection characteristics over a given tuning range. The tuning can be accomplished by varying either the length or the inductive or capacitive loading of the resonators. Research on tunable filters has spanned more than three decades, resulting in a large number of attractive realizations and applications.

Fig. 1 shows a block diagram of the proposed common spare payload for European DBS systems. The two output multiplexers can be tuned via telecommand, thus allowing broadcasts in any frequency band allocated for DBS operating countries. The manifold multiplexers incorporate five four-pole elliptic filters which are mechanically tunable, with a tuning range of $+/-40$ MHz [51]. Fig. 2 demonstrates another typical application of tunable filters in switched tracking preselectors–mixers [15]. This circuit contains a three-sphere YIG-tuned preselector and a single YIG sphere which acts as a discriminator generating an error signal to lock the preselector frequency. Further common applications for tunable band-pass filters include frequency hopped receivers, Doppler radar, and troposcatters.

The above examples demonstrate that the spectrum of applications for tunable filters includes all major areas of microwave engineering. Obviously, different requirements for microwave systems have led to the development of various types of tunable filters with a performance matched to the system demands. Most tunable filters described in the literature fall into three basic types: mechanically tunable,

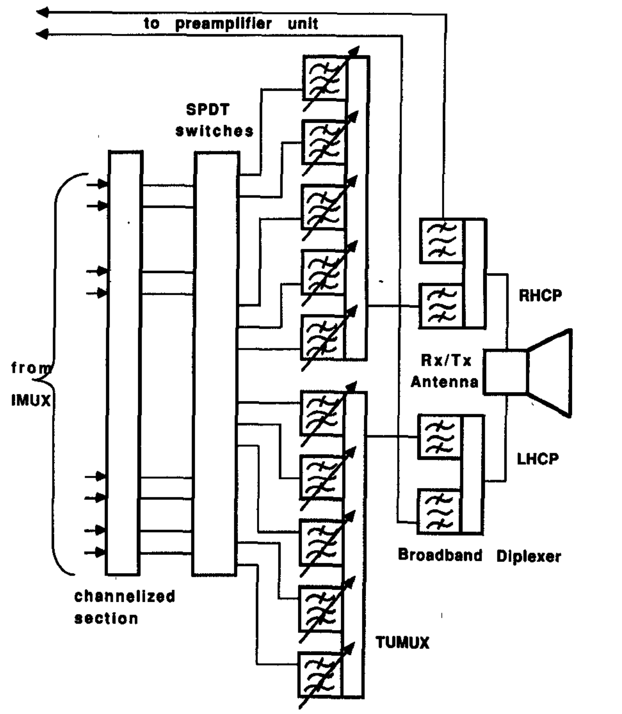


Fig. 1. Block diagram of proposed common spare payload for European DBS containing tunable output multiplexers (after [51]).

magnetically tunable, and electronically tunable filters. Since the number of problems associated with the theory and design of tunable filters is extremely large, it is not possible to deal with all of them in a single paper. Moreover, some of the tuning methods described in the literature are technically extremely complex (e.g. plasma–dielectric multilayer structures in continuously varied electric field [54]) and therefore they have never found practical application. The intention of this paper, therefore, is to highlight only the most important innovations in tunable filters, emphasizing design theory and the resulting improvement in performance.

II. Theory of Tunable Microwave Filters

A. Definitions

The performance of tunable band-pass filters may be described by the same criteria as used for high-quality fixed components (low insertion loss, high selectivity, high dynamic range) and additionally by certain parameters relevant only

Manuscript received February 26, 1990; revised December 10, 1990.
J. Uher was with the Electrical Engineering Department, University of Ottawa, Ottawa, Ontario, Canada. He is now with Spar Aerospace Limited, 21025 Trans-Canada Highway, Ste-Anne-de-Bellevue, Quebec, Canada H9X 3R2.
W. J. R. Hoefer is with the Electrical Engineering Department, University of Ottawa, Ottawa, Ontario, Canada K1N 6N5.
IEEE Log Number 9042495.

0018-9480/91/0400-0643$01.00 ©1991 IEEE

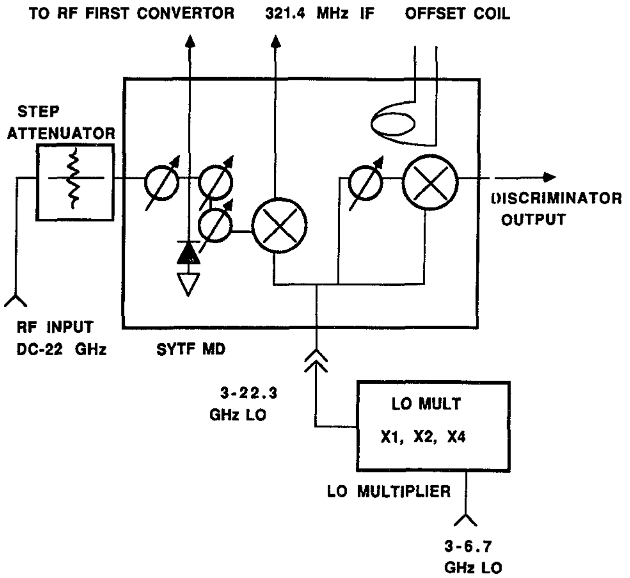


Fig. 2. Magnetically tunable YIG filter in switched-tracking preselector mixer.

to tunable filters. The most important of them are:

Tuning range—defined as the difference between the lowest and the highest midband frequency which is achievable within acceptable limits for insertion loss, bandwidth, and response distortion.

Tuning speed—defined as the time which is necessary to change the filter response to another steady state by the unit frequency shift.

Tuning linearity—the maximum deviation of the center frequency versus a parameter which enforces the tuning (coil current, voltage, static magnetic field, or resonator length variation) from a best fit straight line over the specified operating frequency range.

Tuning sensitivity or tuning efficiency—this can be written as

$$\eta[\Delta f_0] = \frac{f_0(X_2) - f_0(X_1)}{X_2 - X_1} \tag{1}$$

where $f_0(X_2)$ and $f_0(X_1)$ are center frequencies corresponding to X_2 and X_1; X_i is the variable enforcing the tuning.

B. Design Principles and Example

Even though each type of tunable filter has its specific theoretical description, some universal design principles can be formulated. The basic requirements which all tunable filters must satisfy are a constant filter response shape and constant bandwidth over the tuning range. For narrow-band-pass filters, with direct inductive coupling, two independent conditions for constant bandwidth and constant response shape have been derived [1]. If the conditions for constant response and bandwidth are derived from the expressions for the external Q factor of the end resonators, they take the following form:

$$x_1 = (x_1)_m \left[\frac{f_0}{(f_0)_m}\right]^3 \qquad x_n = (x_n)_m \left[\frac{f_0}{(f_0)_m}\right]^3 \tag{2}$$

where $(x_1)_m$ and $(x_n)_m$ are the resonator slope parameter values at the mean tuning frequency $(f_0)_m$.

However, if the same condition is calculated from the expressions for coupling coefficients between resonators, a different formulation is obtained:

$$x_j|_{j=1 \text{ to } n} = (x_j)_m \left[\frac{f_0}{(f_0)_m}\right]^2. \tag{3}$$

The contradiction in these two equations is due to different reference points in calculating the slope parameters. Equation (2) refers to the slope parameters of resonators 1 and n "seen" from the end-coupling reactances X_{01} and $X_{n,n+1}$, while (3) expresses the slope parameters which are "seen" from the interresonator reactances X_{12} and $X_{n-1,n}$. In order to reconcile conditions (2) and (3) and thus minimize bandwidth and change of response shape across the tuning range, coupling reactances X_{01} and $X_{n,n+1}$ must be different from X_{12} and $X_{n-1,n}$.

Good examples of tunable band-pass filters where the number of simplifying assumptions in the design theory is minimized are magnetically tunable printed circuit filters. The features of various E-plane filter types are discussed in another part of this paper. In this section, the design of a ferrite-loaded single metal insert filter (Fig. 3(a)) will be considered. A fixed metal insert band-pass filter belongs to the class of filters with direct inductively coupled cavities and has been analyzed extensively in the literature [4]–[6]. To design tunable filters with resonators partially filled with E-plane ferrite slabs, one must consider some additional aspects. Although the basic synthesis from a lumped/distributed element low-pass prototype followed by computer-aided optimization remains unchanged, the optimization goal will be specific to tunable filters. In this step it will be required that the tuning range become maximum and that other performance parameters (insertion loss, 3 dB bandwidth variation, 40 dB bandwidth variation) not exceed acceptable limits.

The filter is synthesized using the formalism derived by Cohn [7] and generalized by Young [8], Levy [9], and Rhodes [10]. The equivalent circuit of an E-plane inductive septum (Fig. 3(b)) was introduced in [4] and will be used to realize the required impedance inverter. As design specifications we assume a lower and upper band-edge frequency for the mid tuning band $(f_L)_m, (f_H)_m$, out-of-band rejection L (dB), and ripple $(\epsilon)_m$. Then the design procedure can be summarized as follows:

1) Compute the midband guide wavelength $(\lambda_{g0})_m$ at the mid tuning frequency by solving

$$(\lambda_{gL})_m \sin\left[\pi(\lambda_{g0})_m/(\lambda_{gL})_m\right] + (\lambda_{gH})_m \sin\left[\pi(\lambda_{g0})_m/(\lambda_{gH})_m\right] = 0 \tag{4}$$

where $(\lambda_{gL})_m$ and $(\lambda_{gH})_m$ are the guide wavelengths in the resonators at $(f_L)_m$ and $(f_H)_m$. The above equation can be solved numerically after finding roots of the complex transcendental equation associated with the five-layer geometry shown in Fig. 3(c). The method for solving such an equation has been discussed in [36]. At

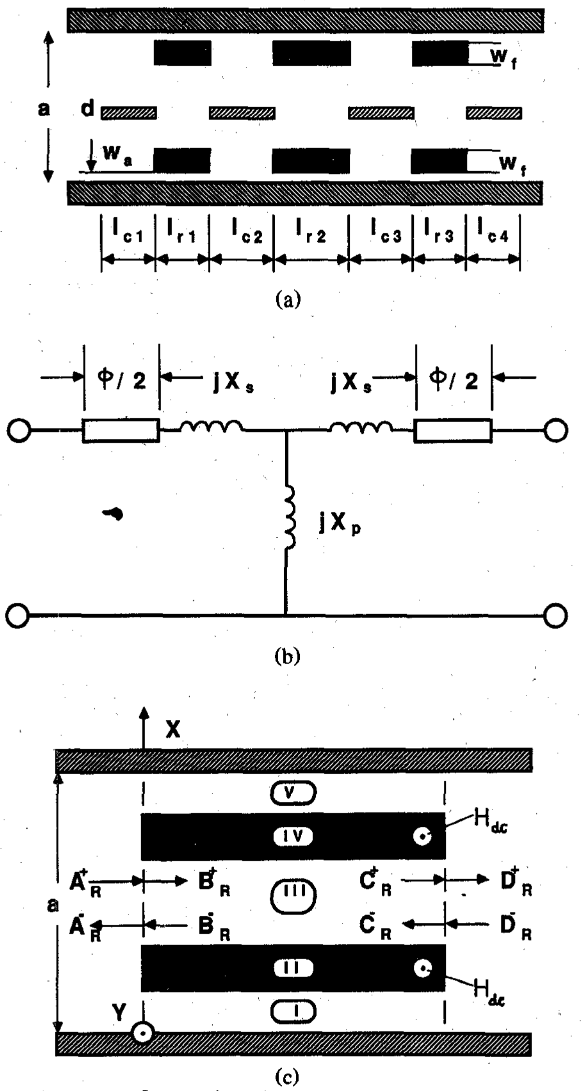


Fig. 3. Magnetically tunable E-plane metal insert filter: (a) general topology; (b) equivalent circuit of impedance inverter; (c) ferrite loaded resonator section.

this point a suitable ferrite material and slab geometry must be determined. For simplicity, the air gap between the ferrite slab and the narrow waveguide wall (due to the nonzero radius of the waveguide corner) should be kept constant while the ferrite slab thickness (w_f), as well as the material data (M_s, ϵ_r), should be varied. For appropriate selection of the resonator loading, both the real and the imaginary part of the cavity eigenvectors must be examined, and the correct solution must satisfy the following conditions:

The maximum value of the attenuation constant of the dominant mode corresponds to a power loss factor of about 0.1 dB/mm (which comes to approximately 1 dB of loss per resonator). The H_{dc} field at which this attenuation occurs is the maximum biasing field, and the corresponding resonant frequency is the upper tuning range limit.

The phase constant variation checked at zero and maximum bias must be maximized in order to make the tuning range as wide as possible. The midband guide wavelength at the center tuning frequency (eq. (4)) can then be computed.

2) Calculate the scaling parameter α as

$$\alpha = (\lambda_{g0})_m \Big/ \left[(\lambda_g)_m \sin\left(\pi (\lambda_{g0})_m / (\lambda_{gL})_m \right) \right]. \quad (5)$$

3) Determine the number of resonators N.

If T_n is the first-kind Chebyshev polynominal of degree N, and λ_g is the guide wavelength at the designated stopband frequency f_s, then the number of resonators can be computed by finding the minimum value of N for which the most severe constraints on the rejection L satisfy

$$L_m = 10 \log \left(1 + (\epsilon)_m \cdot T_n^2 \left[\alpha \frac{(\lambda_g)_m}{(\lambda_{g0})_m} \sin\left(\frac{\pi (\lambda_{g0})_m}{(\lambda_g)_m} \right) \right] \right). \quad (6)$$

If an another type of the filter response is required (e.g. Butterworth, Bessel), (6) must be adequately modified.

4) Calculate the impedances of the distributed elements and the normalized impedance of the inverter. Suitable expressions can be found in [6].

5) Determine the geometry of the coupling sections. Using the equivalent circuit of the impedance inverter (Fig. 3(b)), we obtain

$$K_{j-1,j} = \left| \tan\left(\frac{1}{2}\Phi_j + \tan^{-1} X_{sj} \right) \right|,$$

$$\Phi_j = -\tan^{-1}(2X_{pj} + X_{sj}) - \tan^{-1} X_{sj}. \quad (7)$$

After converting the scattering matrix for the dominant mode of the two-port E-plane septum into an impedance matrix and equating it to the impedance matrix of the equivalent T circuit, one obtains

$$jX_s = \frac{1 - S_{12} + S_{11}}{1 - S_{11} - + S_{12}} \qquad jX_p = \frac{2S_{12}}{(1 - S_{11})^2 - S_{12}^2}. \quad (8)$$

The scattering coefficients are a function of the septum geometry and have been computed using the mode matching technique [37], [38]. Therefore, in order to realize the required inverters, the computation of the S parameters with variable septum lengths must be repeated until these values have been obtained.

6) Compute the lengths of the resonators:

$$(l_r)_j = \frac{(\lambda_{g0})_m}{2\pi} \left[\pi - \frac{1}{2}(\Phi_j + \Phi_{j+1}) \right]. \quad (9)$$

This combined analytical and numerical synthesis procedure yields excellent starting values for the final optimization of the filter. For fixed E-plane filters, a similar synthesis procedure leads to a filter response that satisfies all design specifications [6]. However, for tunable filters some discrepancy between theoretical and experimental curves can be expected. This is mainly due to uncertainty in characterization of the ferrite, whose theoretically predicted permeability

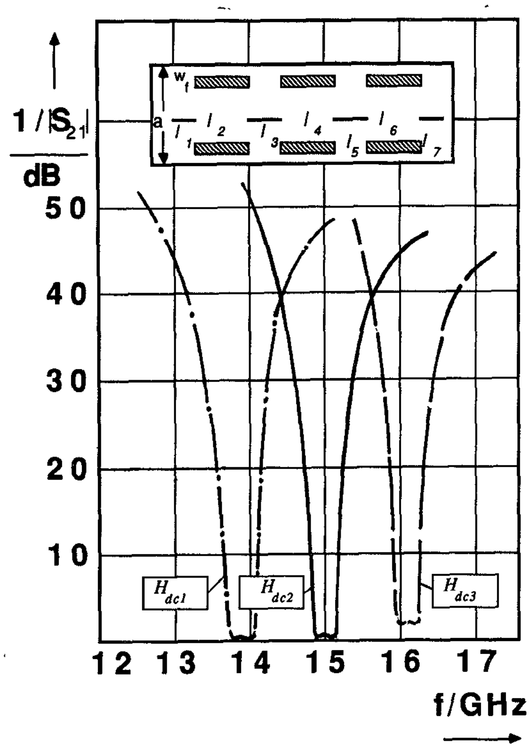


Fig. 4. Computer-optimized Ku-band magnetically tunable E-plane metal insert filter. Frequency response.

tensor parameters do not always reflect the real frequency response of the material. Some improvement in filter performance can be achieved by slightly varying the dimensions of the resonators and coupling sections. Therefore, an adequate optimization routine involving also the widening of the tuning range should be implemented in the design procedure.

As the frame algorithm for optimization, an evolution strategy method has been applied [40]. The frequency response of the filter is computed using a modal S-matrix approach with exact expressions given in [37] and [38].

Fig. 4 shows the computer-optimized frequency response of a Ku-band filter design. The design is based on the following specifications:

> Tuned center frequency: 15 GHz; minimum tuning range: $+/-600$ MHz; passband: 260 MHz; lower rejection: 40 dB at 14.4 GHz; upper rejection: 40 dB at 15.6 GHz; ripple: 0.05 dB, maximum 3 dB bandwidth variation at the tuning range edges: $+/-5\%$; maximum 20 dB bandwidth variation across tuning range: $+/-7\%$.

These requirements have been realized in the following design: waveguide housing: R140; number of resonators: 3; ferrite material ($M_s = 2.74\cdot 10^5$ A/m; $\epsilon_r = 12.8$; slab thickness $w_f = 1.1$ mm); narrow wall spacing $= 0.15$ mm; septum thickness $= 0.19$ mm; $l_{c1} = l_{c4} = 2.219$, $l_{c2} = l_{c3} = 8.855$, $l_{r1} = l_{r3} = 10.478$, $l_{r4} = 10.513$. The filter tuning range has been optimized and goes from 13.85 to 16.20 GHz with the insertion loss 0.8–2.2 dB. The tuned center frequency corresponds to a biasing field of $1.9\cdot 10^5$ A/m.

A wide class of tunable waveguide filters can be designed in a similar way. The most important steps in the analytical/numerical procedure are solution of the eigenvalue problem in the resonator and modal S-matrix computation of the key building blocks of a filter. With this analysis a number of components can be designed. Such devices include metal insert filters, multiple insert filters, large-gap finline filters, inductive/resonant iris filters, evanescent waveguide filters (with E-plane ferrimagnetic slabs in resonators), and mechanically tunable filters. In all other practical cases (YIG filters, varactor-diode filters) a rigorous field theory design is extremely complex and has not been presented so far. However, the filter synthesis from the low-pass prototype including all mentioned preclusions and limitations also yields satisfactory results as will be shown in some examples.

III. Magnetically Tunable Filters

A. YIG Filters

A filter circuit which make use of gyromagnetic coupling was first developed by deGrasse [12] and was reported in 1958. Since these filter types usually contain single-crystal YIG spheres in their resonators, the ferrimagnetic resonance filters are commonly termed YIG filters. However other low-loss ferrites ($(LiFe)_{0.5}Fe_2O_4$ or barium ferrites) can also be used. YIG filters are perhaps the most popular tunable microwave filters because of their multioctave tuning range, very high selectivity, spurious-free response, and compact size. The most remarkable fact in the operating principle of YIG filters is that, if the anisotropy effects can be neglected, the center frequency for a given shape of the resonator does not depend on its size but only on the biasing magnetic field. The resonant frequency of a ferrimagnetic resonator is given by the Kittel equation [13]:

$$\omega_r = |\gamma|\sqrt{\left[H_0 + (N_x + N_x^e - N_z)M_s\right]\left[H_0 + (N_y + N_y^e - N_z)M_s\right]} \tag{10}$$

where N_x^e and N_y^e are the effective demagnetization factors [14], considered as the measure of magnetic anisotropy; N_x, N_y, and N_z are the demagnetization factors in the x, y, and z directions, respectively. The design principles for multistage YIG filters operating from 0.5 to 40 GHz are described in monographs [1], [2] and also in the papers by Carter [20] and Fierstad [16].

Recently, significant progress in millimeter-wave magnetic resonance filters has been reported [15], [19]. Instead of using YIG spheres, the millimeter-wave filters employ highly anisotropic hexagonal ferrites, thus reducing dramatically the value of the external magnetic field, which is necessary for resonance. The filter reported in [15] has been tuned from 50 to 75 GHz with an insertion loss of about 6 dB and an off-resonance isolation (ORI) of 35–40 dB. A W-band filter (75–110 GHz) showed an insertion loss of about 8 dB and an ORI of about 25–30 dB.

A further development in YIG filter technology is the extension of the operating range toward lower frequencies. If the anisotropy effect is neglected, the lowest resonant frequency for the spherical YIG resonator ($M_s = 1.4\cdot 10^5$ A/m) is 1700 MHz, and for Ga-doped YIG it is about 1000 MHz. By using a disk-shaped resonator, one

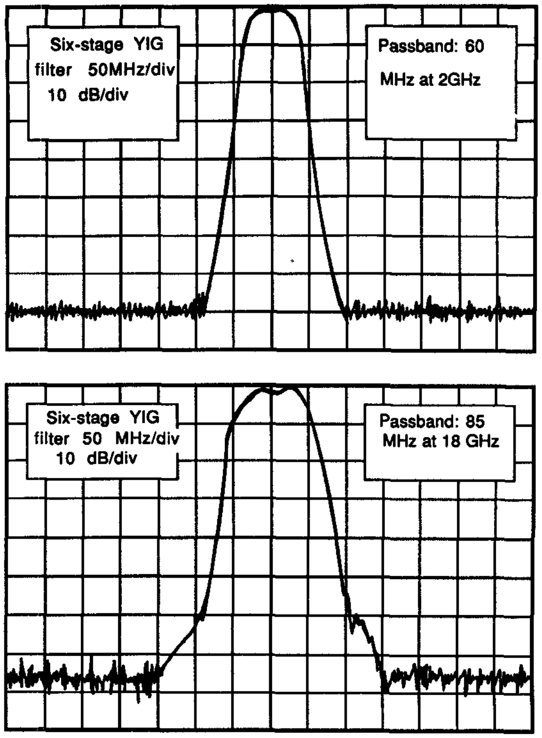


Fig. 5. Frequency response of six-stage YIG filter (after [23]).

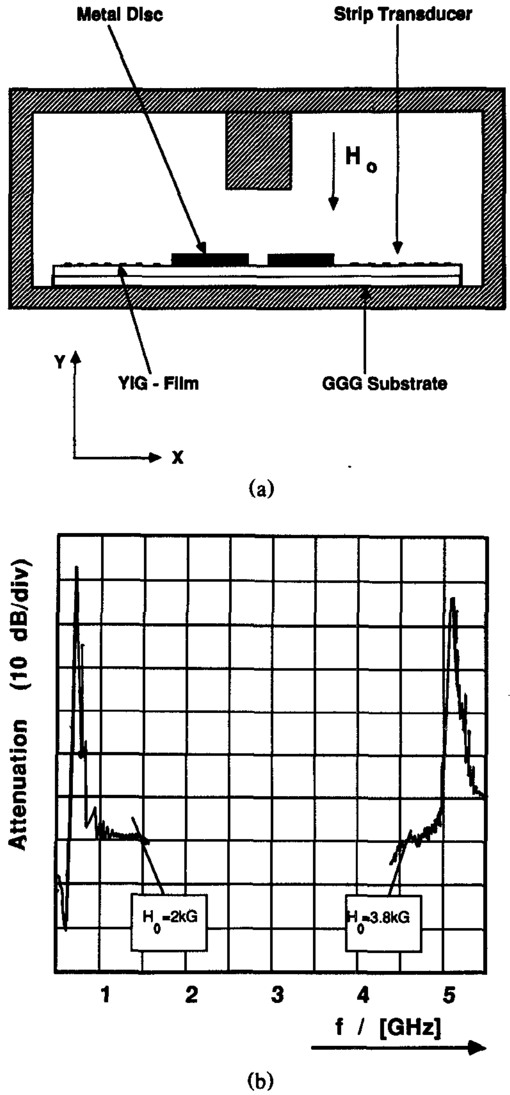


Fig. 6. Magnetostatic wave tunable filter. (a) Cross-sectional view of parallel transducer MSW filter. (b) Frequency response. Geometrical dimensions: YIG film: $20\times5\times0.043$ mm; transducer: $5\times0.2\times0.05$ mm, spacing: 0.3 mm; disk: $\phi = 3$ mm (after [35]).

can reduce this frequency theoretically to almost zero. However, until recently, the lowest resonant frequency achieved with planar resonators was 500 MHz. This discrepancy between theory and experiment has been explained [26] by taking the effective demagnetization factors (eq. (10)) into account. If a single-crystal planar resonator with rotational symmetry is configured so that the axis of rotation is oriented, with respect to the crystalline lattice of the ferrimagnetic material, parallel to a (100) plane for a material with a negative anisotropy and parallel to a (110) plane inside the acute angle formed by two [111] axes for a material with a positive anisotropy, then the lowest resonant frequency of such resonator is as low as 50 MHz.

Another problem associated with YIG filters is their moderate tuning speed, which is usually not below some milliseconds/GHz. The reason for low tuning speed is the induction of eddy currents in the magnetic circuit during the tuning process. More recently, a novel fast tunable type of YIG filter has been reported [24]. A typical biasing circuit containing an electromagnet has been replaced by two orthogonal Helmholtz coils. The YIG resonator is planar in shape and the coupling is accomplished by two orthogonal striplines short-circuited just behind the resonators. The dc magnetic field can be changed either in amplitude or in direction by varying the current intensity in the coils. These changes have no eddy-current delay effect; thus the tuning can be accomplished significantly faster.

Further progress in YIG filter technology has been achieved by introducing six-stage structures. Until recently the maximum number of spheres used in YIG filters was 4. The problems associated with alignment of the magnetic field, which must be uniform for all resonators, were too severe to obtain six-stage filters with overall good performance. This difficulty, however, has recently been overcome [23]. By adding two stages, the passband is widened while the selectivity rises to 36 dB/octave (Fig. 5). The suppression of undesirable out-of-band responses is also significantly improved.

B. Magnetostatic Wave Filters

Magnetostatic wave (MSW) devices have been developed as an extension of SAW components at microwave frequencies [30]–[34]. MSW filters are made of thin YIG films grown by liquid phase epitaxy on gadolinium gallium garnet substrates. Three modes of magnetostatic wave propagation (surface wave, forward volume, backward volume) are possible, depending on the direction of the static magnetic field which is biasing the YIG films to the resonance. The coupling of the microwave energy to magnetostatic modes is accomplished by means of microstrip transducers in meander line or grating configuration. The design of MSW filters is usually based on the computation of the radiation impedance for a given transducer geometry. The thickness of the YIG

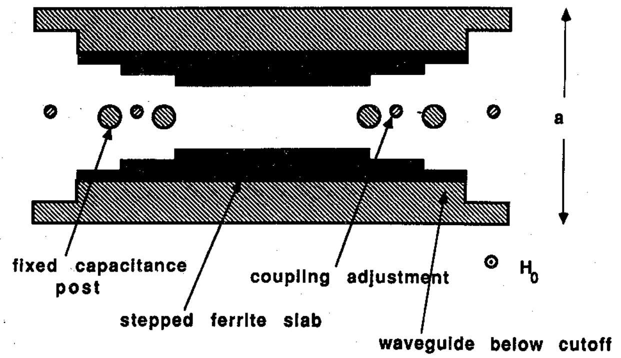


Fig. 7. Evanescent waveguide magnetically tunable band-pass filter.

film determines the bandwidth of the filter, which is usually very narrow (0.1%–1%.) The typical insertion loss of such filters is, however, very high (20–30 dB.)

Recently, a low-insertion-loss MSW band-pass filter (forward volume) has been proposed [35]. The filter configuration (Fig. 6(a)) consists of parallel microstrip transducers (three to seven) and three MSW disk resonators positioned on YIG–GGG substrate. By carefully optimizing the radiation resistance, an average insertion loss of 6 dB within a 0.7–5.2 GHz tuning range (Fig. 6(b)) has been achieved.

C. Evanescent Waveguide Filters

The major drawback of ferrimagnetic resonance filters is their low power handling capability. Even though YIG resonators are coupled by sections of a rectangular waveguide, the power handling capability of such a filter is usually lower than 200 mW. High-power signals excite spurious modes in the nonlinear ferrite and degrade the filter response. The intensity of this undesirable phenomenon is significantly reduced in off-resonance magnetically tunable filters. Two types of filters operating in the off-resonance region have been proposed so far. Historically, the first comprises evanescent waveguide filters [3], [41], [42]. Fig. 7 shows an evanescent waveguide filter with stepped ferrite loading [42]. This filter is a modification of a fixed tuned evanescent filter, and it features a very wide stopband, high stopband attenuation, and compact size. By symmetrically loading the evanescent section with ferrite slabs, the waveguide cutoff frequency is varied. Thus the filter center frequency can be tuned. For the filter design, a regular lumped-element equivalent network synthesis has been applied. Although the synthesis procedure described in [42] made an attempt to compensate for the bandwidth variation in the tuning range, the experimental results were rather unsatisfactory. The bandwidth was strongly reduced within a moderate tuning range. The stopband was narrow, with a relatively low isolation.

D. Printed E-Plane Filters

The drawbacks in the performance of evanescent waveguide filters have been overcome to a large extent with the recently introduced printed E-plane filters [36]–[39]. The operating principle of these filters is based on the well-known fact that the resonant frequency of a resonator partially loaded with ferrite can be tuned by varying the biasing magnetic field. The following two types of microwave structures are represented in this filter class: metal insert filters with symmetrical (Fig. 8(a)) or nonsymmetrical (Fig. 8(b)) ferrite loading, and large-gap finline filters using printed ferrite substrate (Fig. 8(c)) or printed dielectric substrate with symmetrically attached ferrite slabs (Fig. 8(d)). Both types feature low insertion loss and high power handling capability, and they can be designed very accurately using previously described methods. The potential applications range from communication systems (troposcatters, variable output multiplexers) to radar and navigation systems. The most critical aspects of design of these devices are as follows.

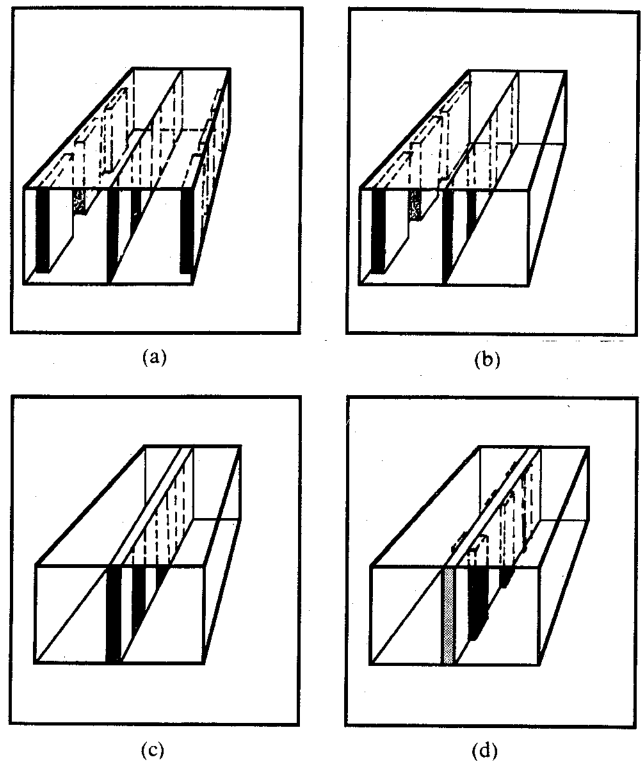


Fig. 8. Magnetically tunable E-plane printed circuit filters. (a) Symmetrically ferrite-loaded metal insert filter. (b) Nonsymmetrically loaded metal insert filter. (c) Large-gap finline filter on a ferrite substrate. (d) Large-gap finline filter on a dielectric substrate.

- There is an absence of ferrite loading in the coupling sections (ferrite parts in the coupling region cause stopband narrowing and large bandwidth variations).
- All air gaps between ferrite and waveguide narrow wall (caused by finite corner radius) must be taken into account while evaluating the eigenvalue problem. If the air gaps are neglected, a serious error in the computation of resonator eigenvectors can occur.
- High power handling capability requires some criteria to be considered for a suitable choice of a corresponding ferrite material and biasing dc field. Ferrites with relatively constant thermal characteristics and low losses are recommended to avoid midband frequency shift when a part of the microwave energy is dissipated in the waveguide housing and ferrite. Subsidiary resonance losses may be reduced if a ferrite material with a suitable spin wave line width combined with appropriate biasing field range is selected. For a device with variable bias, operation below subsidiary resonance is recommended. Some

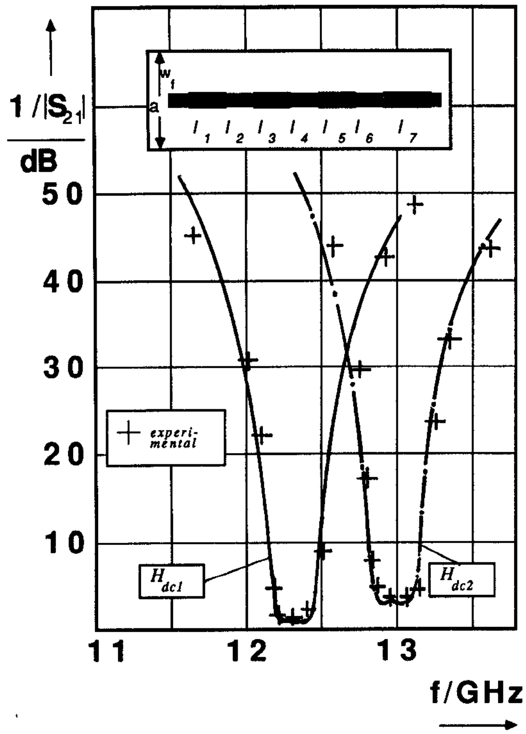


Fig. 9. Ku-band magnetically tunable large-gap finline filter. Computed and measured frequency response. Design data: Ferrite TTVG-1200, $a = 2b = 15.799$ mm, $t = 15$ μm, $l_1 = l_9 = 1.96$ mm, $l_2 = l_8 = 2.136$ mm, $l_3 = l_7 = 6.157$ mm, $l_4 = l_6 = 7.262$ mm, $l_5 = 6.204$ mm, $H_1 = 0$, $H_2 = 2.3 \cdot 10^5$ A/m, $w = 0.7$ mm.

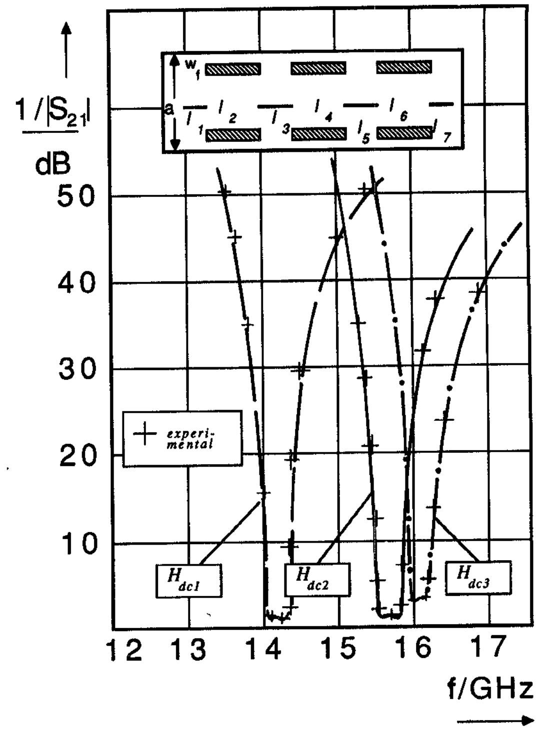


Fig. 10. Ku-band magnetically tunable metal insert filter. Computed and measured frequency response. Design data: Ferrite: TTI-2800, $a = 2b = 15.799$ mm, $t = 0.19$ mm, $l = 50$ mm, $w = 1$ mm, $l_1 = l_7 = 3.59$ mm, $l_2 = l_6 = 8.916$ mm, $l_3 = l_5 = 9.417$ mm, $l_4 = 8.921$ mm, $H_1 = 0$, $H_2 = 1.72 \cdot 10^5$ A/m.

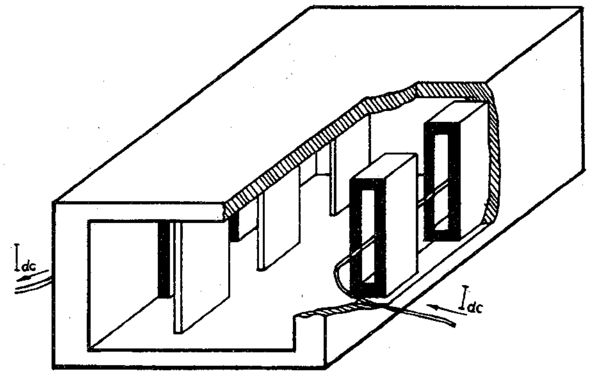
examples of maximum values for the magnetic field H_{dc} for high-power tunable filters are given in [37].

- The millimeter-wave filter design calls for very high accuracy in the manufacturing of the filter circuit. The ferrite slabs for frequencies above 40 GHz are too thin and brittle to be handled safely. Therefore, thin-layer deposition techniques rather than bulk material must be considered for millimeter-wave structures.

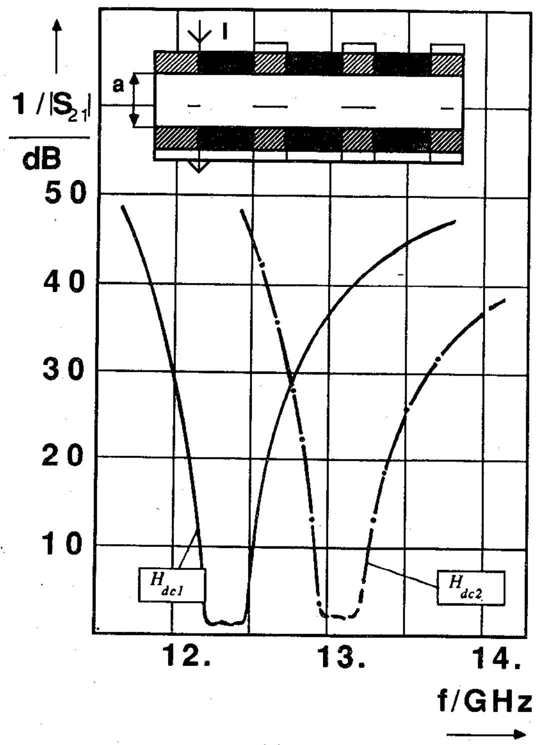
Since all E-plane filters can be treated analytically with high accuracy, the resulting designs show very good performance and excellent agreement with the theoretically predicted response. Fig. 9 shows the calculated and measured responses of a magnetically tunable finline filter for the Ku band. This filter type was manufactured by etching a finline circuit from a Cu-sputtered TTVG-1200 (Transtech Inc.) ferrite substrate. The thickness of the ferrite slab was 1 mm. The measured insertion losses ranged from 1.3 to 2.3 dB. Since the electrical lengths of both the copper-clad coupling sections and the resonator regions depend equally on the biasing magnetic field, the bandwidth decreases relatively quickly as the filter is tuned. Fig. 10 shows the computed and measured responses of a tunable metal insert filter for the Ku band. The resonators are symmetrically filled with lateral TTI-2800 (Transtech Inc.) ferrite slabs. The measured insertion loss was about 1.3 dB in the center of the tuning range. The insertion loss can be further reduced by loading only one side of the resonator with ferrite. The measured insertion loss was about 0.8 dB for this filter type. The trade-off is mostly in reduced tuning efficiency and narrower stopband. The tuning speed of all the filters presented is of the order of milliseconds and is comparable to the tuning speed of common YIG filters. Much faster tuning can be achieved if the homogeneous ferrite slabs are replaced by small ferrite toroids (Fig. 11(a)). For this filter type only a single wire loop is necessary for biasing. The typical switching time for the phase shifters utilizing similar toroids is about 1–3 μs. Thus, the tuning speed will be improved by two to three orders of magnitude for such filters. Fig. 11(b) shows the calculated Ku-band response of this filter type. In this design a ferrite material with a large built-in magnetic field ($2.8 \cdot 10^5$ A/m) has been assumed. This was necessary to obtain a sufficient level of premagnetization since the biasing field generated by a single wire is usually too low. The midband of the filter can be tuned between 12.3 and 13.1 GHz with an average insertion loss of 1.4 dB. The design theory, as well as experimental verification of such a filter, was described in [53].

IV. Electronically Tunable Band-Pass Filters

Electronically tunable filters can be tuned very fast (about 1 GHz/μs) over a wide (an octave) tuning range, and they offer compact size. They have found wide application in ESM receivers [43]. In most electronically tunable filters, GaAs varactor diodes are used. The varactor diode capacitance varies with reverse (negative) voltage. When a varactor is in series with a resonator circuit or element, this capacitance change alters the resonant frequency. A typical varactor-tuned filter utilizes a combline circuit (Fig. 12(a)) constructed in suspended stripline technique [44], [46]. The small-signal equivalent circuit usually assumes resonators consisting of distributed inductors in parallel with lumped capacitors, while the coupling is represented by distributed

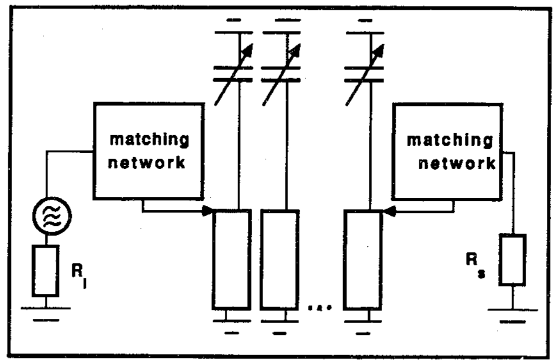


(a)

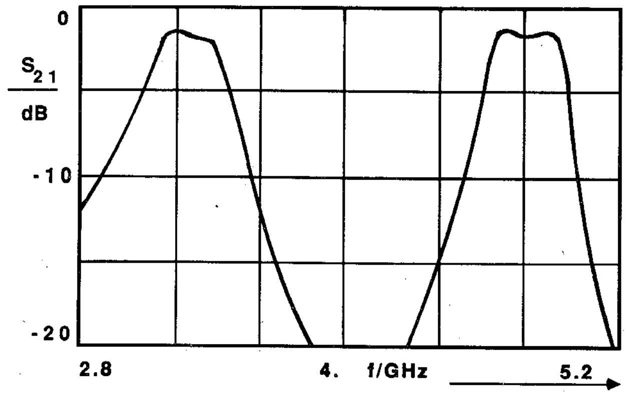


(b)

Fig. 11. Fast tunable E-plane passband filter. (a) Positioning of ferrite toroids in a waveguide housing. (b) Frequency response of Ku-band design. Design data: $a = 2b = 15.799$ mm, $a_1 = 19.29$ mm, $w_f = 0.6$ mm, $w_d = 0.4$ mm, $w_a = 0.15$ mm, $H_{dc1} = 0$, $H_{dc2} = 2.75 \cdot 10^5$ A/m, $M_s = 1.96 \cdot 10^5$ A/m.



(a)



(b)

Fig. 12. Varactor diode tuned combline filter. (a) Topology of the filter circuit. (b) Frequency response, after [46].

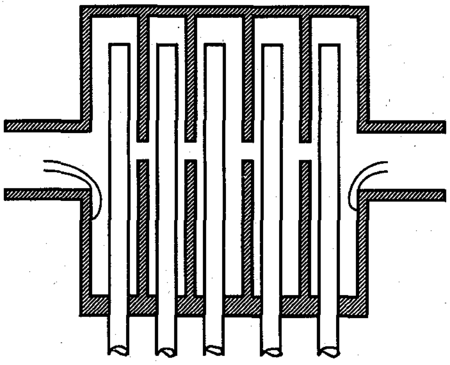
series inductors. The frequency dependence of internal coupling can be compensated by a matching network which also takes into account the varactor Q effect. The Q factor of a varactor GaAs diode is usually low ($Q = 75$–125 at 5 GHz); therefore the insertion loss of the varactor-tuned filter depends mainly on the Q factor of the diode. The maximum reported insertion loss of two-resonator filters tuned from 3 to 5 GHz is 5.5 dB [44]. Recently, a similar filter has been designed using a more rigorous theory, resulting in a reduced insertion loss: maximum 1.5 dB [46]. Fig. 12(b) shows the computed and measured filter responses for different bias values. The variations of bandwidth within the tuning range have also been reduced and do not exceed 10%.

The main advantage of varactor-tuned filters resides in their superior tuning speed. However, they also have some serious disadvantages. Two-resonator filters are characterized by a relatively low selectivity (12 dB/oct) and low stopband isolation (30 dB). By adding more resonators, an improved selectivity can be achieved, but only at the expense of larger insertion loss. Another problem associated with varactor-tuned filters is their low power handling capability. Since a varactor diode is a nonlinear device, larger signals generate harmonics and subharmonics. Large-signal testing of such filters indicated a third-order intercept point of 7 dBm at zero bias and 37 dBm at 30 V bias [44].

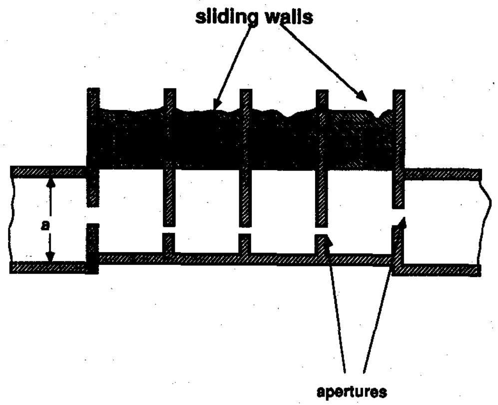
Varactor-tuned filters can also be realized in rectangular waveguide technique. Examples of such filters are discussed in [45]. Contrary to common tunable band-pass filters, they feature variable bandwidth while maintaining a constant center frequency. This feature has been obtained by coupling the resonators through variable capacitance. A varactor-tuned filter with similar passband behavior realized in planar technique has been described in [49]. Electronically tuned band-pass filters using ceramic resonators have also been reported [47], [48].

V. Mechanically Tunable Filters

Mechanically tunable band-pass filters still draw considerable attention among filter designers. Their large power handling capability and low insertion loss are often decisive factors if a tunable filter for long-distance communication (satellite or troposcatter) or radar systems is required. Me-



(a)



(b)

Fig. 13. Mechanically tunable filters. (a) Mechanically tunable band-pass filter with coaxial resonators. (b) Waveguide tunable band-pass filter.

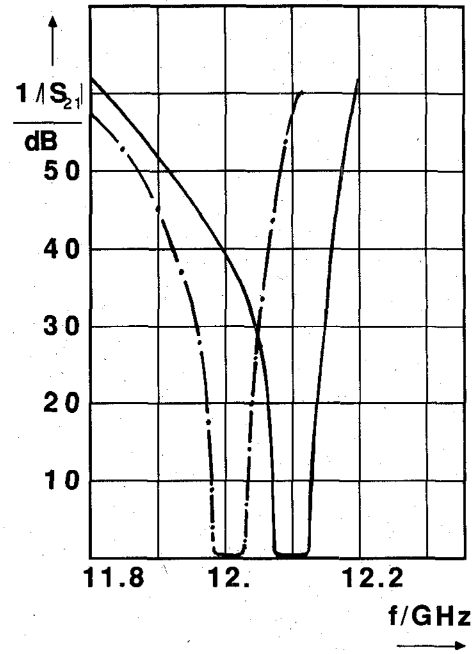


Fig. 14. Mechanically tunable TE_{113} dual-mode band-pass filter—frequency response (after [51]).

chanically tunable filters are usually realized using either coaxial or waveguide resonators [1]. A systematical design theory of tunable waveguide filters was presented in [50]. Fig. 13(a) shows a four-resonator coaxial band-pass filter. The resonators operate in the TEM mode and are $\lambda_g/4$ long at resonance. The input and output ports are coupled to the resonators by means of a magnetic loop. All resonators are approximately coaxial in the cross section with a somewhat flattened coupling region to keep the irises as thin as possible. The coaxial filters are manufactured with between two and six resonators and cover the frequency range from 400 MHz to 12 GHz with a typical tuning range between 20% and 60% of a standard waveguide band.

Fig. 13(b) shows a waveguide counterpart of a coaxial filter. The filter is tuned by a sliding wall on one side of each resonator cavity. The resonators are coupled by apertures which are shifted from the positions of the aperture coupling at the input and output. This reduces the variations of bandwidth and response shape as the filter is tuned. The cavities of mechanically tunable filters (usually between two and six) can be either rectangular or cylindrical. In [52] a TE_{011}-mode filter with four cylindrical cavities was reported. The tuning can be accomplished by a movable plunger of a diameter smaller than or equal to that of the cavity. The TE_{011}-mode field distribution allows for a nontouching piston because the wall currents are purely circumferential. Therefore the Q factor of the cavity remains high ($Q = 9000$–$11\,000$). Such filters possess a very high power handling capability (500 W CW at 12 GHz).

The main drawbacks of the TE_{011} filters are spurious resonances, which are relatively close to the main resonance and must be suppressed to an acceptable level, and the large size and mass of such filters. This size can be reduced by using dual-mode techniques. In [51] a TE_{113} dual-mode tunable filter was presented. The two cavities are arranged side by side in split block technique. The filter is tuned by only two plungers which are located opposite to the coupling side. Since for this particular filter only a moderate tuning range is required ($+/-50$ MHz at 12 GHz), only a small part of the end plate is used as a plunger. Fig. 14 shows the measured frequency response of the filter [51]. The return loss of this filter is better than 25 dB, and the equiripple bandwidth varies by about 10%.

The biggest disadvantage of the mechanically tunable filters is their low tuning speed. The filters can be tuned manually or via telecommand if the filter is combined with a remotely controlled motor.

VI. Conclusions

In Table I, the performance of all the filters presented is summarized. From this comparison it is obvious that none of these devices can simultaneously satisfy all requirements for perfect tunable filters. For microwave systems where multi-octave tuning is essential, the YIG filter is an obvious choice. In systems where the requirement of high power handling capability combined with low insertion loss predominates, mechanically tunable filters and magnetically tunable E-plane filters are recommended. If the tuning speed is a crucial requirement, varactor-tuned filters or E-plane filters with ferrite toroids are devices of choice. For millimeter-wave design, the most promising structures are ferrimagnetic resonance filters utilizing hexagonal ferrite resonators or, up to

TABLE I
TYPICAL PERFORMANCE PARAMETERS OF MICROWAVE TUNABLE BAND-PASS FILTERS

Performance Parameter	Mechanically Tunable Filter	YIG Filter	MSW Filter	*E*-Plane Filter	Varactor Tuned Filter
Bandwidth [%]	0.3–3	0.2–3	0.2–0.5	1–10	2–20
Insertion loss [dB]	0.5–2.5	3–8	6–10	0.7–2.5	0.3–2.5
Selectivity [dB/oct]	12–24	12–36	24	12–24	12–24
Rejection att. [dB]	> 50	40–60	> 45	> 50	> 30
Power handling capability [W]	100–500	0.1–1	0.05	50–200	0.05–0.1
Tuning range [% of WG-band]	5–20	multi-octave	multi-octave	60–70	~ octave
Tuning speed [GHz/ms]	very low	0.5–2	0.5–2	0.5–10^3	10^3
BW variation [%]	10–20	10–40	10–40	5–10	10–20
Tuning linearity [MHz]	+/−15	+/−10	+/−10	+/−15	+/−35
Temp. Stability [MHz]	20	15	15	25	25
Millimeter-wave capability	no	yes	no	yes	no

60 GHz, magnetically tunable *E*-plane printed circuit filters. Further developments in the field of microwave tunable filters will depend strongly on the progress in the relevant microwave technologies. The most important fields for research seem to be in MMIC integration, high-Q varactors, and low-cost manufacturing methods for single-crystal ferrite layers. Superconductive cavities will also have a strong impact on the future development of high-performance tunable filters.

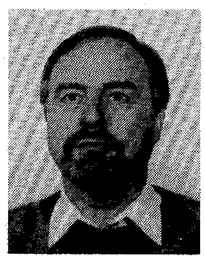
REFERENCES

[1] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance Matching Networks, and Coupling Structures*. New York: McGraw-Hill, 1964, ch. 17.
[2] J. Helszajn, *YIG Resonators and Filters*. New York: Wiley, 1985.
[3] G. Craven, *Evanescent Mode Waveguide Components*. Boston: Artech House, 1987.
[4] Y. Konishi and K. Uenakada, "The design of a bandpass filter with inductive strip-planar circuit mounted in waveguide," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, pp. 1209–1216, 1974.
[5] R. Vahldieck, J. Bornemann, F. Arndt, and D. Grauerholtz, "Optimized waveguide *E*-plane metal insert filters for millimeter-wave applications," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, pp. 65–69, Jan. 1983.
[6] L. Q. Bui, D. Ball, and T. Itoh, "Broad-band millimeter-wave *E*-plane bandpass filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, pp. 1655–1659, Dec. 1984.
[7] S. Cohn, "Direct-coupled-resonator filters," *Proc. IRE*, vol. 45, pp. 187–196, Feb. 1957.
[8] L. Young, "Direct-coupled cavity filters for wide and narrow bandwidths," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-11, pp. 162–178, 1963.
[9] R. Levy, "Theory of direct-coupled cavity filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-15, pp. 340–348, 1967.
[10] J. D. Rhodes, *Theory of Electrical Filters*. New York: Wiley, ch. 4.
[11] I. Wolff, *Felder und Wellen in Gyrotropen Mikrowellenstrukturen*. Braunschweig: Verlag Vieweg and Sohn, 1973.
[12] R. W. deGrasse, "Low-loss gyromagnetic coupling through single crystal garnets," *J. Appl. Phys.*, vol. 30, pp. 1555–1559.
[13] C. Kittel, "On the theory of ferromagnetic resonance absorption," *Phys. Rev.*, vol. 77, pp. 155–161, 1948.
[14] R. Soohoo, *Microwave Magnetics*. New York: Harper and Row, 1985.
[15] H. Tanbakuchi, D. Nicholson, B. Kunz and W. Ishak, "Magnetically tunable oscillators and filters," *IEEE Trans. Magn.*, vol. 25, pp. 3248–3253, Sept. 1989.
[16] R. F. Fjerstad, "Some design considerations and realizations of iris-coupled YIG-tuned filters in the 12–40 GHz region," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-18, pp. 205–212, Apr. 1970.
[17] G. H. Thiess, "Theory and design of tunable YIG-filters," *Microwaves*, Sept. 1964.
[18] K. D. Gilbert, "Dynamic tuning characteristics of YIG-devices," *Microwave J.*, no. 6, pp. 36–40, June 1970.
[19] D. Nicholson, "Ferrite tuned millimeter wave bandpass filters with high off-resonance isolation," in *IEEE 1988 MTT-S Int. Microwave Symp. Dig.*, May 1988, pp. 867–870.
[20] P. S. Carter, "Equivalent circuit of orthogonal-loop-coupled magnetic resonance filters and bandwidth narrowing due to coupling resonance," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-18, pp. 100–105, Feb. 1970.
[21] P. S. Carter, "Side-wall coupled, strip-transmission line tunable filters employing ferrimagnetic YIG resonators," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-13, pp. 306–315, May 1965.
[22] L. Rhymes, "Two-sphere YIG multiplier/filter ensures purity," *Microwaves/RF*, pp. 109–116, Apr. 1988.
[23] M. Korber, "Six-stage filters," *Watkins-Johnson Company Technical Notes*, vol. 15, no. 5, Sept./Oct. 1988.
[24] C. Vittoria, "High speed frequency tunable microwave filter," U.S. Patent 4 197 517, Apr. 8, 1980.
[25] C. Vittoria, "Tunable microwave filters utilizing a slotted line circuit," U.S. Patent 4 590 448, May 20, 1986.
[26] G. U. Sorger and D. Raicu, "Magnetically tunable resonators and tunable devices such as filters and resonant circuits for oscillators using magnetically tuned resonators," U.S. Patent 4 555 683, Nov. 26, 1985.
[27] L. Young and D. B. Weller, "A 500–1000 MHz magnetically tunable bandpass filter using two YIG-disc resonators," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-15, pp. 72–86, Feb. 1967.
[28] "Improving Microwave Measurement with Ferre-trac Electr. Tuned Filters 0.5–26 GHz," Technical Note "Ferretec"-FT3, Ferretec Inc., Fremont, CA.
[29] "YIG-Filters," Technical Note,—"Siversima," Culver City, CA.
[30] Y. Murakami, T. Ohgihara, and T. Okamoto, "A 0.5–4.0 GHz tunable bandpass filter using YIG-film grown by LPE," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-35, pp. 1192–1197, Dec. 1987.

[31] J. P. Castera and P. Hartemann, "A multipole magnetostatic volume wave resonator filter," *IEEE Trans. Magn.*, vol. MAG-18, pp. 1601–1603, 1982.
[32] E. Huijer and W. Ishak, "MSSW resonators with straight edge reflectors," *IEEE Trans. Magn.*, vol. MAG-20, pp. 1232–1234, 1984.
[33] J.D. Adam, "A MSW tunable bandpass filter," in *Proc. IEEE Ultrasonics Symp.*, 1985.
[34] W. S. Ishak and K. W. Chang, "Tunable microwave resonators using magnetostatic wave in YIG-films," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-34, pp. 1383–1393, Dec. 1986.
[35] T. Nishikawa *et al.*, "A low-loss magnetostatic wave filter using parallelstrip transducer," in *1989 IEEE MTT-S Int. Microwave Symp. Dig.*, June 1989, pp. 153–158.
[36] J. Uher, J. Bornemann, and F. Arndt, "Magnetically tunable rectangular waveguide *E*-plane integrated circuit filters," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 1014–1022, June 1988.
[37] J. Uher, J. Bornemann, and F. Arndt, "Computer-aided design and improved performance of tunable ferrite-loaded *E*-plane integrated circuit filters," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 1841–1858, Dec. 1988.
[38] J. Uher, J. Bornemann, and F. Arndt, "Ferrite tunable millimeter wave printed circuit filters," in *IEEE 1988 MTT-S Int. Microwave Symp. Dig.*, May 1988, pp. 871–874.
[39] J. Uher and W. J. R. Hoefer, "Analysis and tuning efficiency optimization of magnetically tuned printed *E*-plane circuit filters," in *IEEE 1989 MTT-S Int. Microwave Symp. Dig.*, June 1989, pp. 1273–1276.
[40] H. Schmiedel, "Anwendung der Evolutionsoptimierung bei Mikrowellenschaltunden," *Frequenz*, vol. 35, pp. 306–310, Nov. 1981.
[41] R. F. Skedd and G. Craven, "Magnetically tunable multisection bandpass filters in ferrite-loaded evanescent waveguide," *Electron. Lett.*, vol. 3, pp. 62–63, Feb. 1967.
[42] R. Snyder, "Stepped-ferrite tunable evanescent filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, pp. 364–371, Apr. 1981.
[43] C. B. Hofmann and A. R. Baron, "Wideband ESM receiving systems," *Microwave J.*, no. 9, pp. 24–34, Sept. 1980.
[44] I. C. Hunter and J. D. Rhodes, "Electronically tunable microwave bandpass filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-30, pp. 1354–1360, Apr. 1982.
[45] S. Toyoda, "Variable bandpass filters using varactor diodes," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, pp. 356–362, Apr. 1981.
[46] S. Kumar and Y. Liang, "Varactor tuned suspended substrate combline filter," in *Proc. Can. Conf. Elect. and Comp. Eng.*, Sept. 1989.
[47] M. A. Harris, "High-power tunable filter," U.S. Patent 4 692 724, Sept. 8, 1987.
[48] J. B. West, "Ceramic TEM resonator bandpass filters with varactor tuning," U.S. Patent 4 721 932, Jan. 26, 1988.
[49] A. Schwarzmann, "Adjustable passband filter," U.S. Patent 4 250 457, Feb. 10, 1981.
[50] B. Rawat, "Tunable waveguide filters—A practical design procedure," *Microwaves and RF*, no. 9, pp. 97–102, Sept. 1983.
[51] U. Rosenberg, D. Rosowski, W. Rummer, and D. Wolk, "Tunable manifold multiplexers—A new possibility for satellite redundancy philosophy," in *Proc. MIOP Int. Conf.* (W. Germany), Feb. 1989.
[52] M. A. Kunes and G. G. Connor, "A digitally controlled tunable, high output filter for space applications," in *Proc. European Microwave Conf.* (London), Sept. 1989.
[53] J. Uher and W. J. R. Hoefer, "Fast tunable bandpass filter with high power handling capability," in *MIOP 1990 Conf. Dig.*
[54] G. C. Tai, C. H. Chen, and Y.-W. Kiang, "Plasma-dielectric sandwich structure used as a tunable bandpass microwave filter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, pp. 111–113, Jan. 1984.

Jaroslaw Uher (M'88) received the M.Sc. degree in electronic engineering from the Technical University of Wroclaw, Poland, in 1978 and the Dr.-Ing. degree in microwave engineering from the University of Bremen, Bremen, West Germany, in 1987.

From 1978 to 1982 his work dealt with planar ferrite technology at the Institute of Electronic Technology, Technical University of Wroclaw. From 1983 to 1988 he was with the Microwave Department of the University of Bremen, where his research activities involved field problems of ferrimagnetic slab discontinuities in waveguides structures and the design of ferrite control components and tunable filters. From September 1988 to August 1990 he was a Research Engineer at the University of Ottawa, Ottawa, Canada, where he was involved in research on quasi-planar components and numerical techniques. In September 1990 he joined SPAR Aerospace Ltd., where he is presently working as a Senior Member of the Technical Staff, developing numerical tools for microwave components design for satellite hardware.



Wolfgang J. R. Hoefer (M'71–SM'78–F'91) received the diploma in electrical engineering from the Technische Hochschule Aachen, Aachen, Germany, in 1964 and the D.Ing. degree from the University of Grenoble, France, in 1968.

After one year of teaching and research at the Institut Universitaire de Technologie, Grenoble, France, he joined the Department of Electrical Engineering, University of Ottawa, Ottawa, Ont., Canada, where he is currently a Professor. His sabbatical activities have included six months with the Space Division of AEG-Telefunken in Backnang, Germany, six months with the Electromagnetics Laboratory of the Institut National Polytechnique de Grenoble, France, and one year with the Space Electronics Directorate of the Communication Research Centre in Ottawa, Canada. His research interests include microwave measurement techniques, millimeter-wave circuit design, and numerical techniques for solving electromagnetic problems.

Dr. Hoefer is a registered Professional Engineer in the province of Ontario, Canada.