Microwave Filters—Applications and Technology

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Invited Paper

***Abstract*—This paper describes the development of microwave filter technology from an applications perspective. Military applications required wide-band and tunable devices for electronic support measures receivers, which led to the development of highly selective wide-band waveguide filters, coaxial resonator and suspended-substrate multiplexers, and electronically tunable filters. The satellite communications industry created demand for low-mass narrow-band low-loss filters with severe specifications on amplitude selectivity and phase linearity. These requirements resulted in the development of dual-mode waveguide and dielectric-resonator filters, and advances in the design of contiguous multiplexers. Cellular communications base-stations demanded low-loss high power-handling selective filters with small physical size, capable of being manufactured in tens of thousands at a reasonable cost. These demands led to advances in coaxial resonator, dielectric resonator, and superconducting filters, and also to methods of cost-reduction, including computer-aided alignment. Cellular radio handsets have required the manufacture of hundreds of millions of extremely small very low-cost filters, still with reasonably low loss and high selectivity. This has driven significant advances in integrated ceramic, surface, and bulk acoustic-wave active and filters using micromachined electromechanical systems.**

***Index Terms*—Dielectric resonators, diplexers, filters, microwave filters, multiplexers, resonators.**

I. Introduction

ELECTRICAL filters have the property of frequency-selective transmission, which enables them to transmit energy in one or more passbands and to attenuate energy in one or more stopbands. The design of microwave filters is normally accomplished by using a low-pass prototype network as a starting point, regardless of the eventual physical realization in transmission line, waveguide, or other media. Low-pass prototype networks are two-port lumped-element networks with an angular cutoff frequency of 1 rad/s and operating in a 1-Ω system. A typical prototype network is shown in Fig. 1. The design of such networks has developed to a remarkable degree of sophistication. The use of network synthesis [1] enables them to be designed exactly to meet a given approximating transfer function such as Chebyshev, elliptic function, linear phase, etc. [2], [3]. The ladder network prototype of Fig. 1(a) and (b) is an "all-pole" network with all its transmission zeros at infinity. This is because all the series inductors become open circuited and all the shunt capacitors become short circuited at infinite frequency. The alternative network shown in Fig. 1(c) and (d) is also used. In this case, a shunt capacitor and a pair of coupling elements, known as impedance inverters, enabling simpler realization of bandpass filters, replace a series inductor. Other prototype networks, with arbitrary location of the transmission zeros in the complex plane, may also be used, enabling highly selective transfer functions with specified phase characteristics to be realized [4]. The optimally selective transfer function for an all-pole prototype is the Chebyshev [2] one with equiripple amplitude in the passband and monotonically increasing attenuation in the stopband [see Fig. 2]. The prototype networks for more selective transfer functions require either resonant circuits or cross couplings between nonadjacent resonators [2]–[4].

More details on the theory of prototype networks will be found in [5]. However, a few points are worth mentioning. First, the selectivity of a prototype network (defined as the rate of change of signal transmission with respect to frequency) depends on the number of elements used (the degree of the filter). For example, a formula for the degree of a Chebyshev filter required to meet a given specification is given in [2]. Secondly, the low-pass prototypes of Fig. 1 may be converted into other forms, i.e., high-pass, bandpass, bandstop, etc., by applying a frequency transformation to the prototype [2]. A typical example is the capacitively coupled bandpass filter [3] shown in Fig. 3. Here, the shunt capacitors transform into parallel tuned circuits, or resonators, and the inverters are replaced by capacitive coupling elements combined with a slight retuning of the resonators. Finally, one of the challenges in filter design is that of overcoming dissipation loss. As the number of resonators is increased in order to increase the selectivity, the group delay of a filter increases. Furthermore the group delay of a bandpass filter is inversely proportional to its fractional bandwidth. In addition, the resonators used in a filter have a finite unloaded quality (Q) factor, which depends upon their physical realization. Now, for a dissipative system, as the group delay is increased, so will the insertion loss. Thus, for a specified passband loss and degree, a narrow-band filter will require resonators with a higher Q than a broad-band filter, and will be physically larger if the same type of resonators are used in both cases. This is illustrated in Fig. 4 for fourth-degree Chebyshev bandpass filters. Formulas for calculating the midband insertion loss of bandpass filters are given in [2] and [6].

It should be noted that, at microwave frequencies, lumped realizations of high-Q filters are not usually practical because the

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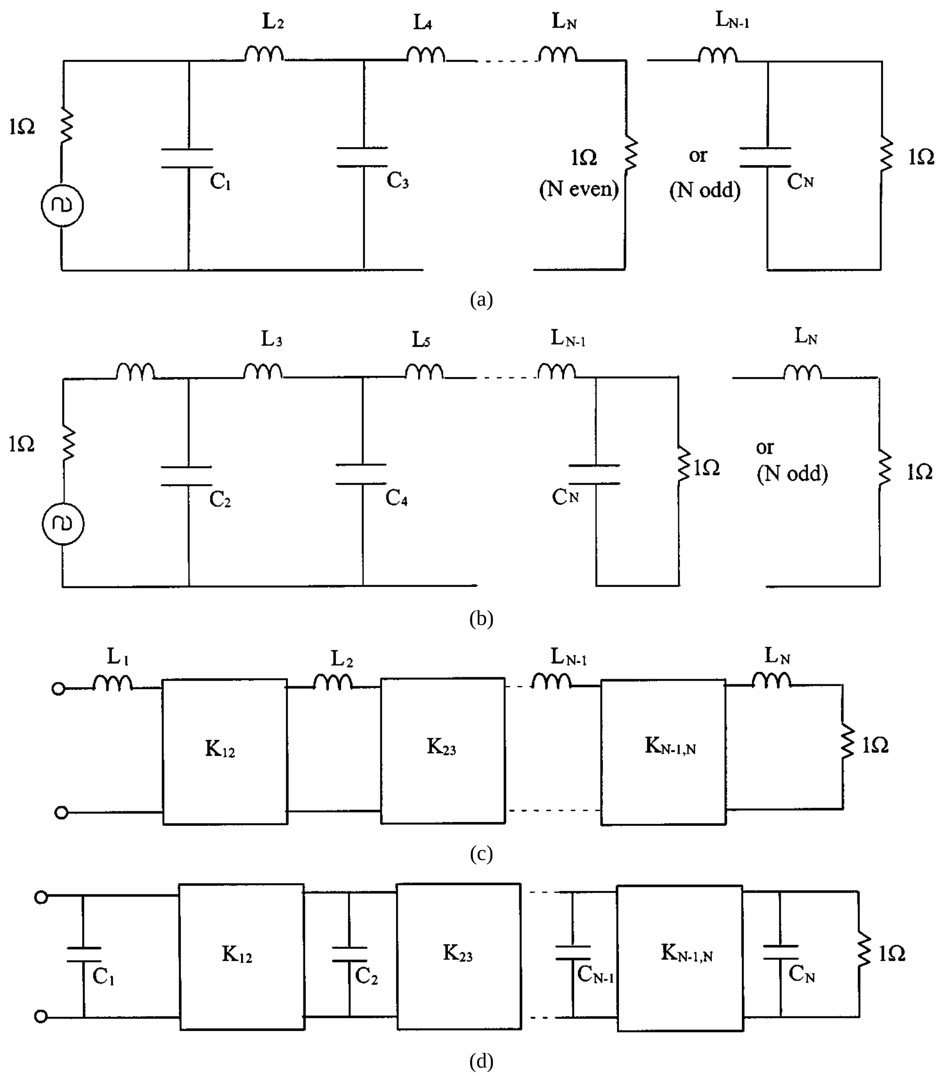

Fig. 1. (a), (b) Low-pass prototype networks for "all-pole" filters. (c), (d) Alternative low-pass prototype networks using inverters.

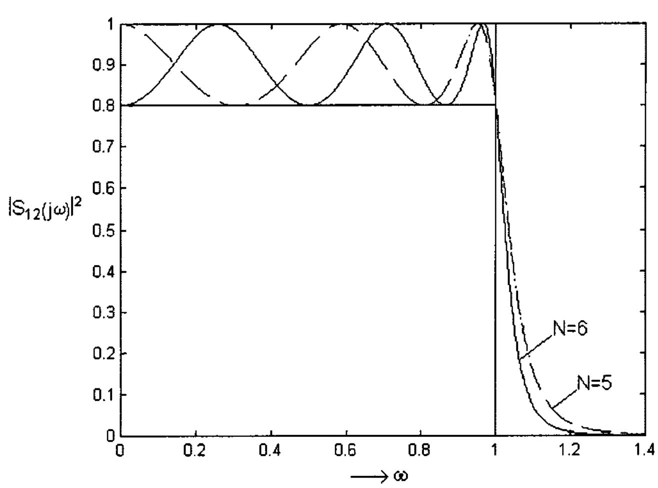

Fig. 2. Chebyshev transfer function.

wavelength becomes comparable to the physical dimensions of circuit elements. For this reason, a variety of distributed circuit element realizations are used, where one or more of the dimensions of the elements are comparable with the wavelength of operation. Connections of distributed circuit elements are described by distributed network theory [7]. For example, various types of microwave filters may be realized using transmission lines operating in the TEM mode. One example is the interdigital filter [8], which consists of an array of parallel coupled TEM lines that are shorted at alternate ends. Such devices enable practical construction of filters with resonator Q's of 1000–5000. Further increases in Q may be achieved using rectangular or circular waveguide resonators, or dielectric resonators with Q's of up to 50 000. Information on the Q's of TEM and waveguide resonators is given in [9].

In this paper, an attempt has been made to describe the exciting advances in microwave filter technology in the context of their application. In Section II, we will look at military applications, which, with the invention of radar, were the first real driver for microwave technology. This led to advances in the design of wide-band filters. Following this, we will examine satellite communications, which began in the 1960s and led to rapid advances in waveguide and dielectric resonator multiplexers. This is followed by a discussion of cellular communications, which has caused further advances in dielectric resonator, acoustic wave, and other filter technologies.

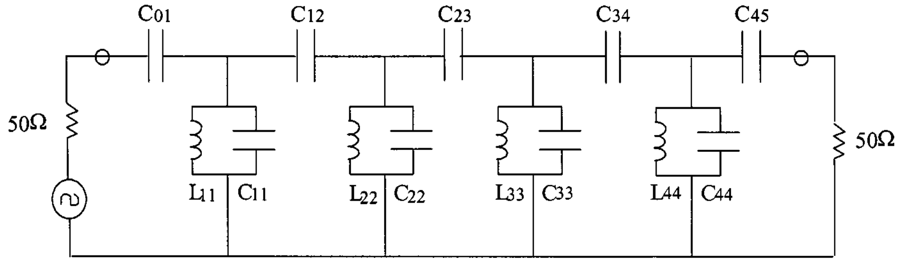

Fig. 3. Schematic of capacitively coupled bandpass filter.

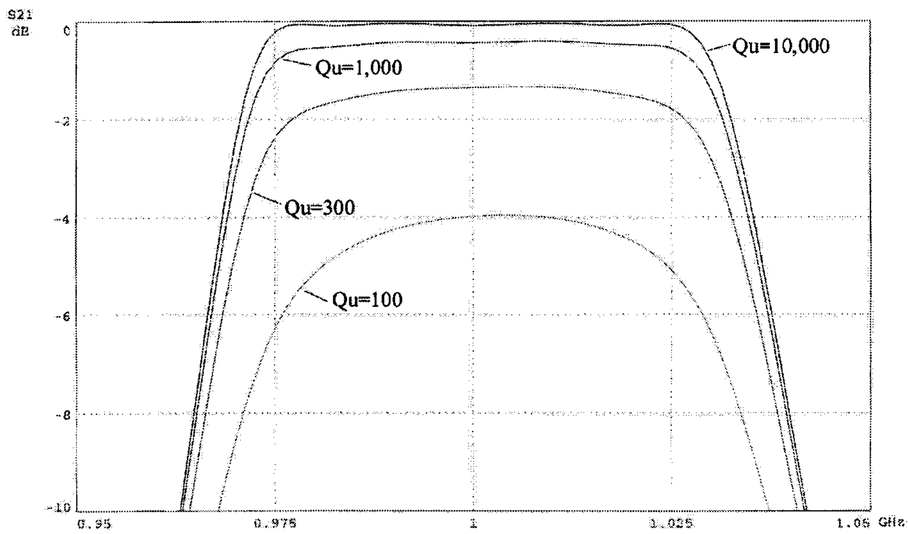

Fig. 4. Effect of finite resonator Q on the response of a fourth-degree Chebyshev bandpass filter.

II. Military Applications

World War II and the invention of radar led to significant developments in filters at various laboratories in the U.S. For example, at the Massachusetts Institute of Technology (MIT) Radiation Laboratory, Cambridge, work concentrated on narrow-band waveguide filters for radar systems. At the Harvard Radio Research Laboratory, Cambridge, MA, advances on broad-band TEM filters for electronic support measures (ESM) systems and tunable narrow-band filters for search receivers were made. Most of this work is described in [10].

One of the critical parts of any military system is the electronic countermeasures (ECM) system and its associated ESM system. The ESM system detects and classifies incoming radar signals by amplitude, frequency, pulsewidth, etc., and the ECM system can then take appropriate countermeasures, such as jamming. One method of classifying signals by frequency is to split the complete microwave band of interest into smaller sub-bands. This can be done using a contiguous multiplexer, which consists of separate bandpass filters whose passbands crossover at their 3-dB frequencies. The outputs of the individual channels can be detected, giving coarse frequency information while retaining unity probability of intercept [11]. Typical specifications might be for a quadruplexer covering the frequency range of 2–4, 4–8, 8–12, and 12–18 GHz with each channel having less than 1-dB loss over most of its band, less than 5-dB loss at crossover, and more than 60-dB attenuation for all frequencies more than 10% away from crossover.

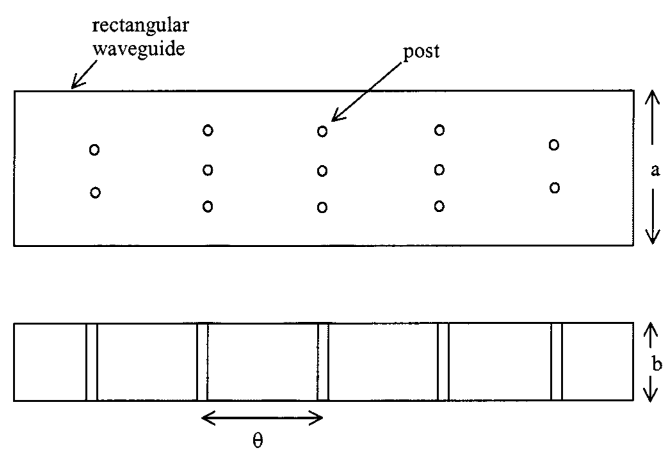

Fig. 5. Waveguide bandpass filter.

Initially work concentrated on extending the bandwidth of waveguide filters. The rectangular waveguide filter consists of a uniform section of rectangular guide with post (or other) discontinuities placed across the broad walls of the guide at approximately half-guide-wavelength intervals (Fig. 5). Usually, the waveguide is operated in its fundamental transverse electric mode (TE_{01}) mode of operation [12]. The equivalent circuit of the discontinuity is inductive [13] (Fig. 6), thus, the equivalent circuit of the filter consists of sections of uniform guide separated by inductive discontinuities. Methods of accurately designing this device for wide-band operation occupied several

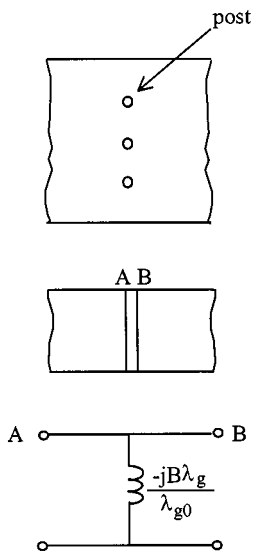

Fig. 6. Inductive waveguide discontinuity.

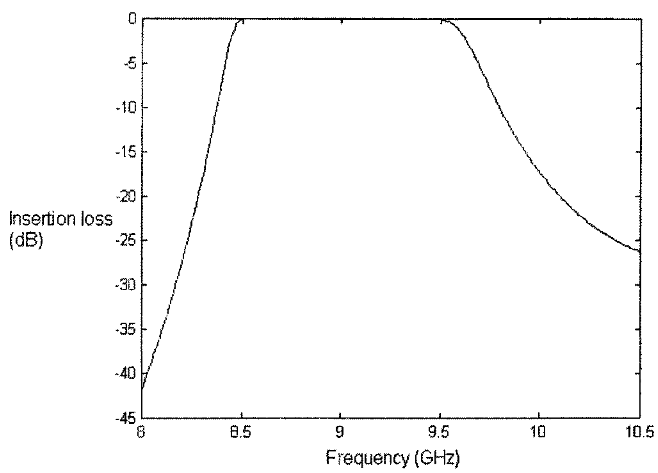

Fig. 7. Frequency response of a waveguide bandpass filter.

workers for a considerable period [14]–[16]. A waveguide is a hollow pipe without a center conductor and it has a finite cutoff frequency. Thus, waveguide bandpass filters have zero transmission at cutoff, giving an asymmetric frequency response (Fig. 7). One of the advantages of waveguide filters is that Q's of up to 10 000 at 10 GHz are achievable [9]. Secondly, they can handle very high power levels of up to several kilowatts per continuous wave, [17]. A disadvantage of using waveguide filters is the relatively large size required for low-frequency operation. The broad-wall dimension of a rectangular waveguide must be considerably greater than one-half the free —space wavelength at cutoff. Thus, a 2-GHz filter would have a broad-wall dimension of at least 10 cm.

We have already seen that the loss of a bandpass filter is inversely proportional to its fractional bandwidth; thus, high-Q waveguide realizations are not really necessary for octave bandwidth filters. Much smaller filters may be constructed using TEM transmission lines, which do not need a minimum cross-sectional dimension to ensure propagation. The most significant developments were the parallel coupled-line filter [18],

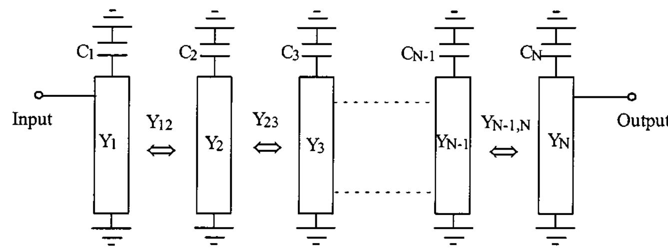

Fig. 8. Combline filter.

which has found numerous applications in microstrip subassemblies, the interdigital filter [8], and the combline filter [19]. The combline filter is of particular interest, as it has stood the test of time and variants of it are widely used in cellular radio base-stations. It consists of an array of equal-length parallel coupled conductors, each of which is short circuited to ground at the same end with capacitive loading on opposite ends (Fig. 8). The equivalent circuit is shown in Fig. 9. Here, we see that the capacitive loading will drive the resonant frequency of the resonators below that of the series couplings so that relatively strong inter-resonator couplings can occur. The combline filter has several advantages; firstly, it is compact, as the coupled conductors are typically one-eighth wavelength long. Secondly, the electrically short resonators will not re-resonate until typically six times the center frequency of the filter, giving a broad spurious-free stopband, which is not possible with wide-band interdigital filters. Thirdly, it is easier to manufacture than the interdigital filter, as all the tuning screws required for electrical alignment can be on the same face of the filter. Finally, the center frequency of the combline filter may be tuned by an octave or more without causing significant distortion to its frequency response [20]. An accurate design method for wide-band combline filters is described in [21]. The design of multiplexers using wide-band combline filters is reported in [22]–[24].

Combline multiplexers have certain disadvantages. They are relatively large and they are difficult to electrically align. Although elliptic function combline filters have been reported [25], the wide-band combline structure does not readily lend itself to realizing the most optimum transfer functions such as the generalized Chebyshev, which has an equiripple amplitude characteristic and arbitrarily placed transmission zeros [2]. Consequently, combine designs require considerably more than the optimum number of resonators. These disadvantages led to advances in printed circuit realizations of multiplexers, the most notable method being suspended substrate stripline. This consists of a thin (typically 0.1 mm) substrate suspended in an air cavity. A complex multiplexer circuit can be printed on a single substrate enabling a very accurate, almost tuning-free realization. The use of the generalized Chebyshev prototype enables realizations with the minimum impedance variation, which is essential for a printed realization. Extremely high selectivities have been reported with this approach [26]–[28]. A picture of a typical device is shown in Fig. 10.

Electronically tunable filters are used in scanning receivers as they give relatively fine frequency resolution, albeit at the expense of a relatively low probability of intercept. They are also commonly used in microwave test equipment. The most commonly used tuning mechanism consists of single crystal

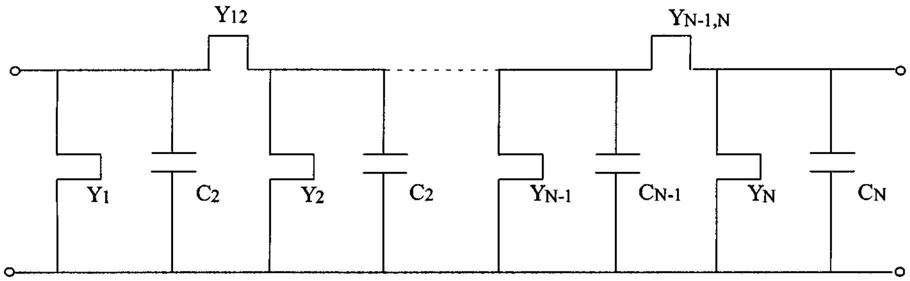

Fig. 9. Equivalent circuit of the combline filter.

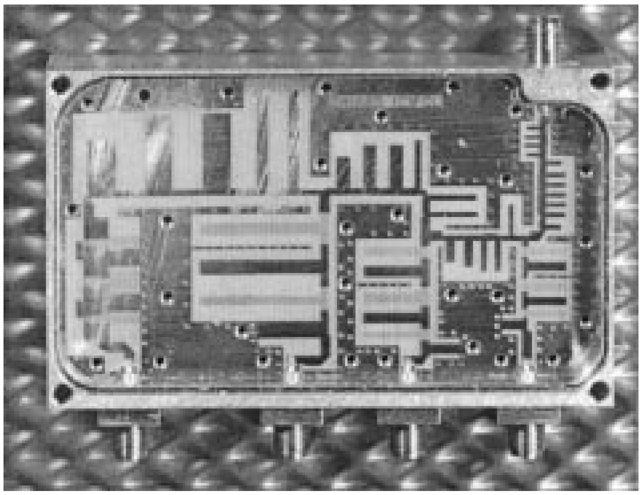

Fig. 10. Suspended substrate multiplexer (courtesy of Filtronic plc, West Yorkshire, U.K.).

spheres of ferrimagnetic yttrium–iron–garnet (YIG) with an external dc magnetic field. The resonance may be tuned higher in frequency by increasing the magnetic field. Multioctave tuning is possible with Q's of 2000 at 2 GHz and increasing with frequency, leveling out at around 5000 at 10 GHz. There is also a minimum tuned frequency when the applied magnetic field is equal to or less than the demagnetizing field. Typically, this frequency is from 1 to 2 GHz. One of the problems with YIG devices is that they are very unstable with temperature and often require heating. In addition, power limiting occurs at relatively low power levels, typically from 10 mw to 10 W. Details of the design of YIG filters are presented in [29].

III. Satellite Communications

Satellite communications began with the Intelsat I–III series of satellites, which established the viability of voice communications in the late 1960s. In 1971, the Intelsat IV series was launched. This was the first to use a channelized architecture, which is necessary to avoid problems associated with the nonlinearity of high power amplifiers. In this architecture, the 500-MHz uplink band from 5.925 to 6.425 GHz is received in the satellite, amplified using a low-noise amplifier (LNA) and mixed down to the downlink band from 3.7 to 4.2 GHz. It is then split into 12 36-MHz channels, using an input multiplexer, prior to amplification of each channel. The individual channels, known as transponders, are then recombined using an output multiplexer and fed to the downlink (transmit) antenna [30]. This created the need for high-performance filters and multiplexers as the dominant linear signal distortions (amplitude and phase) would be caused by these devices. The insertion loss of the input multiplexers is of relatively little importance as the noise figure is defined by the LNA. Thus, interactions between filters could be minimized by using a circulator-coupled approach combined with all-pass group delay equalizers. On the other hand, the insertion loss of the output multiplexer is critical, as it directly affects the link budget. The output multiplexers used rectangular waveguide filters constructed from invar (for temperature stability), which were combined in a common waveguide manifold, with each channel filter providing high isolation to signals in nonadjacent bands. Thus, two output multiplexers containing even- and odd-numbered channels are then combined onto the antenna using a hybrid combiner. These noncontiguous multiplexers provide minimum interaction between channel filters, enabling ease of alignment of the complete multiplexer.

These multiplexers were very large and heavy, approximately 4 kg per channel, and methods of reducing this were of paramount importance. The first advances were in using lightweight materials such as graphite, which were used in the first U.S. domestic satellites [31], enabling sufficient weight reduction so that 24 channels could be deployed [32]. The first major electrical innovation was the use of dual-mode filters, where size reduction is obtained by exciting two orthogonal degenerate modes in the same physical cavity. This was first reported in 1951 [33]. The first practical devices were developed by Comsat Laboratories, Canada [34]–[36]. A typical device known as a dual-mode in-line waveguide filter is shown in Fig. 11. In this structure, each cavity supports two orthogonal degenerate modes (e.g., TE_{111}). The modes in each cavity are coupled together via a tuning screw, or other discontinuity, which is oriented at 45° to the input iris. The two horizontally and vertically polarized modes in each cavity are coupled to the corresponding modes in the adjacent cavity. This structure allows certain nonadjacent resonator couplings to be present in the structure, which allowed the realization of transfer functions with transmission zeros located arbitrarily in the complex plane such as elliptic function, generalized Chebyshev, linear phase, etc. Prior to this, single-mode realizations of these transfer functions used folded waveguide structures [37]. Dual-mode filters were first launched on Intelsat IV.A in 1976, after which they became the satellite industry standard. Indeed, new types of dual-mode filter are still being reported, e.g., [38].

Triple mode designs offer further size reduction and a practical design, which allowed independent control of the TE_{111} and

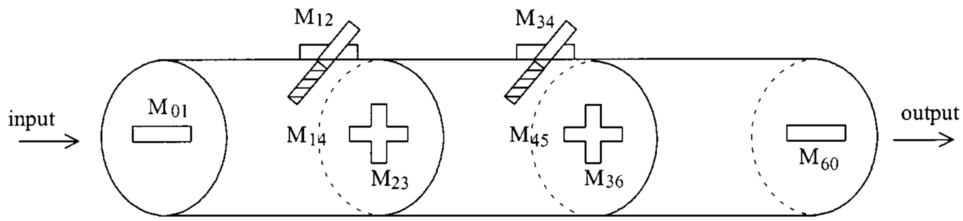

Fig. 11. Dual-mode in-line waveguide bandpass filter.

TM_{010} modes in cylindrical waveguides, was reported in [39]. Quadruple mode designs based on TE_{113} and TM_{110} modes have also been reported [40]. Unfortunately, these designs are very sensitive and suffer from poor temperature stability.

The next stimulus to research was the adoption of higher frequencies, in particular, the 14/12-GHz band for direct broadcast satellites. Standard designs at these frequencies would have an unacceptable 1 dB more loss than the 6/4-GHz band. This led to the development of TE_{10N} and TE_{11N} filters, which had N times half-wavelength cavity lengths [41].

Advances were also made in the methods and theory of multiplexing. The noncontiguous approach used in the 1970s suffered from the requirement for a complex hybrid combiner to feed the even and odd channels into the antenna. In the 1980s, a considerable effort led to the development of contiguous multiplexers [42]–[44]. This was mainly due to advances in computer-aided design and modeling. These devices provided approximately 1 dB less loss, superior channel characteristics, and higher out-of-band isolation.

Satellite communications also stimulated work on filter transfer functions and network synthesis. A new method of computing symmetrical transfer functions with otherwise arbitrary passband ripple and transmission zero locations is described in [45]. A comprehensive theory of symmetrical dual-mode filters is presented in [46]. The design of asymmetric dual-mode filters, which are required for the end channels of multiplexers, is described in [47] and [48].

The use of dielectric resonators constructed from low-loss high-permittivity (20–100) temperature stable ceramics enables high-Q (up to 100 000) filters to be realized in a fraction of the volume and weight of air-filled waveguide devices [49]. Probably the most significant development in this area was the dual-mode in-line device reported in 1982 [50]. This device is shown in Fig. 12, showing the similarity with air-filled devices. This device matched the performance of single-mode devices and has been used on several satellites.

High-temperature superconductivity shows considerable promise. In principle, superconductivity enables resonators with near-infinite unloaded Q to be constructed in a very small size, provided that the filters are cooled to 77 K [51]. A collaborative project between COM DEV, Lockheed Martin, and Du-Pont has demonstrated miniature input and output multiplexers with 50% mass saving at C-band [52]. Even more recently, collaboration between the Institut de Recherche en Communications Optiques et Microondes (IRCOM), Brive, France, and Alcatel, France, have demonstrated planar filters at 29 GHz [53]. We will hopefully soon see superconducting multiplexers working in space.

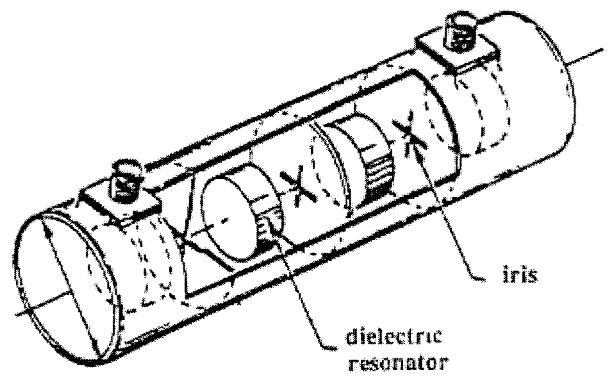

Fig. 12. Dual-mode in-line dielectric-resonator bandpass filter.

Surface acoustic wave (SAW) filters also have applications in the satellite industry. Their application is mainly limited to antialiasing filters in digital payload down-converter chains. Center frequencies of 500 MHz–1.5 GHz and bandwidths of 10–50 MHz are achievable with extremely sharp selectivities [54]. Transition bandwidths of only 1 MHz are achievable, which is superior to ceramic dielectric filters, although those devices have lower loss. They also find limited application as channelizers in IF processors, where high selectivity and linear phase is required. It is unlikely that they will be used in input or output multiplexers.

Finally power-handling issues of multipactor breakdown and passive intermodulation (PIM) are important. Multipaction is an RF breakdown mechanism, which occurs in a vacuum. It is caused by an electron resonance, which occurs when the transit time of electrons is similar to the period of one cycle of the field. Design solutions to this problem usually involve altering internal dimensions such that the electron transit time is nonresonant, dielectric filling, or filling with an inert gas [55]. PIM occurs when there are thin oxide layers on metal surfaces, mechanical imperfections in contacts, microcracks in metal surfaces, or due to the presence of dirt or metal particles on the surfaces. PIM in the output multiplexers can leak back to the receiver and must be controlled carefully. This involves a set of guidelines for workmanship standards [56].

IV. Cellular Radio

Cellular radio has provided a significant driver for filter technology since the analog systems were launched in the early 1980s. This has resulted in various innovations in filter technology for both base-stations and handsets, which have depended upon the frequency planning of the various systems standards. In the U.S., the analog Advanced Mobile Phone

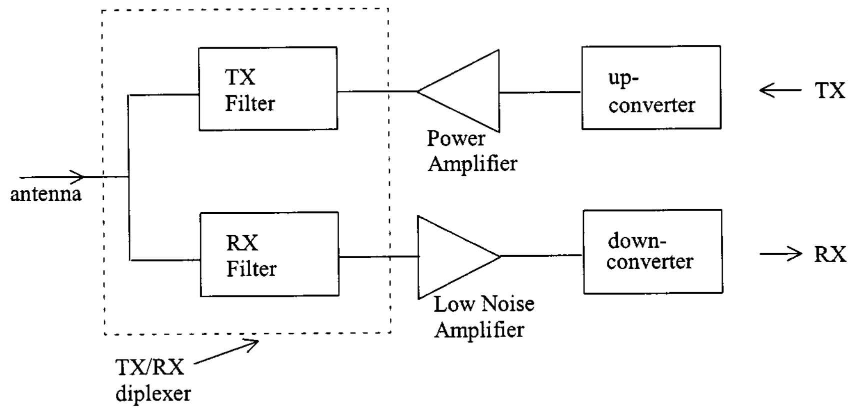

Fig. 13. RF front-end of a cellular radio base-station.

Service (AMPS) used a frequency-division multiple-access (FDMA) scheme, and was allocated 869–894 MHz for base-station transmit (mobile receive) and 824–849 MHz for base-station receive (mobile transmit). The digital time-division multiple-access (TDMA) system (IS136) occupies the personal communications system (PCS) band from 1930 to 1990 MHz and 1850 to 1910 MHz and each 60-MHz band is sub-banded and allocated to operators in three 15-MHz and three 5-MHz segments. The American code-division multiple-access (CDMA) system (IS95) occupies both of the above bands and is sub-banded into 5-, 10-, or 15-MHz segments. In Europe, the original analog total access communication system (TACS) occupied 890–905 and 935–950 MHz. This was extended (ETACS) to 872–905 and 917–950 MHz. The digital TDMA global system for mobile communications (GSM) [57] occupies 925–969 and 880–915 MHz and the bands from 1710 to 1785 and 1805 to 1880 MHz. These systems are not sub-banded. The third-generation universal mobile telecommunications system (UMTS) uses CDMA in the bands from 1920 to 1980 MHz and 2110 to 2170 MHz.

The block diagram of the RF front-end of a typical cellular radio base station is shown in Fig. 13. The systems are normally transmitting and receiving simultaneously. The transmitter will be generating relatively high power signals, e.g., for GSM two 30-W carriers, and the receiver needs to detect signals down to -100 dBm. The transmit filter must have a very high level of attenuation in the receive band, typically 90 dB in order to stop intermodulation products and noise from the power amplifier being fed into the receiver. Furthermore, the transmit filter must have low passband insertion loss, typically 0.5–1 dB, in order to satisfy power amplifier linearity and efficiency requirements. Similarly, the noise figure of the receiver dictates low insertion loss in the receive filter and this filter should have high isolation in the transmit band.

The first analog systems required filters with percentage bandwidths in the region of 2% and with reasonable guard bands between channels. These specifications may be met with asymmetric generalized Chebyshev bandpass filters, typically with six resonators with a Q of 3000 and one or possibly two transmission zeros located on one side of the passband. Such devices are relatively easy to construct using variations on the combline filter where the resonators are coupled via irises between the cavities. The transmission zeros may be realized by having additional couplings between nonadjacent resonators, in this case, around three resonators [58]. The relative sign of the additional coupling with respect to the main couplings determines whether the zeros are on the low- or high-frequency side of the passband. GSM filter specifications may also be met with this technology, although the closer proximity of transmit and receive bands means that higher Q filters with more resonators are required, typically ninth-degree filters with a Q of 5000 are used. A picture of such a device is shown in Fig. 14, and its measured performance is shown in Fig. 15. It should be noted that the proximity of the bands in the GSM case means that the filters are often temperature compensated, normally by constructing the resonators out of dissimilar metals with different temperature coefficients [59]. With these methods, temperature coefficients as low as 5 ppm/°C can be achieved. Additionally, GSM transmit filters normally have severe PIM specifications, often specified to be below -115 dBm for two 30-W carriers.

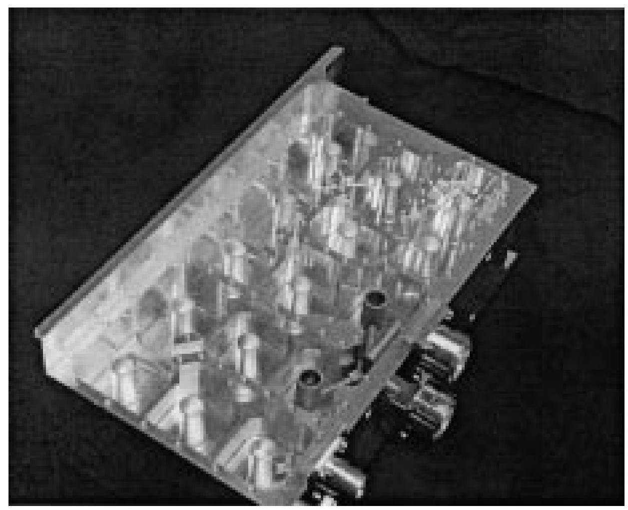

Fig. 14. Coaxial resonator realization of a cellular radio base-station filter.

The sub-banded American systems require selective filters with percentage bandwidths as low as 0.25%. These filters require higher Q resonators with much better temperature stability than achievable with coaxial resonator filters. Dielectric resonator filters have been proven useful in this respect. The most commonly used designs use a cylindrical puck of ceramic suspended on a support within a metallic housing. The fundamental

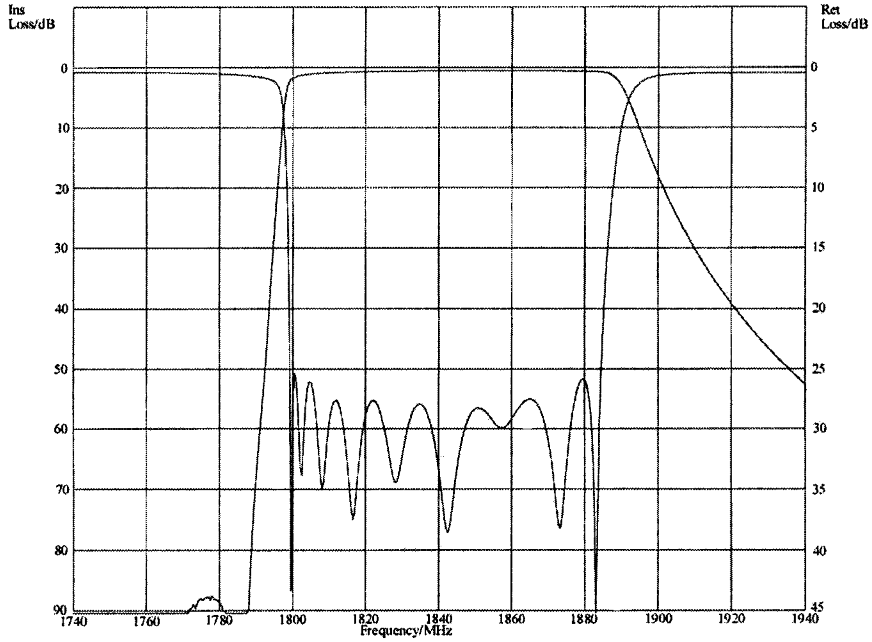

Fig. 15. Measured response of coaxial resonator filter.

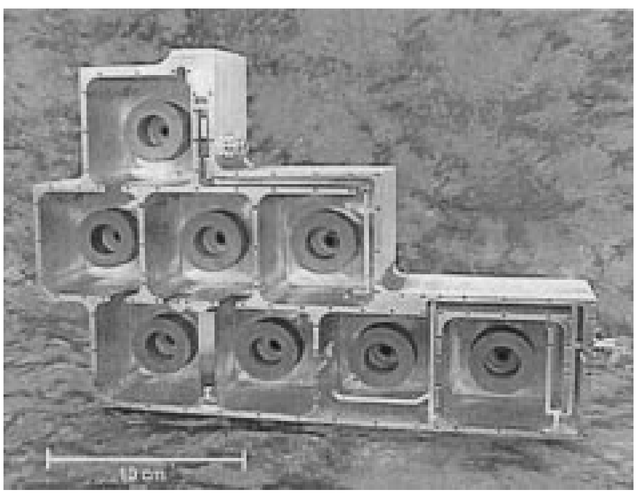

Fig. 16. Typical dielectric resonator filter.

mode of resonance is the $\mathrm{TE}_{01\delta}$ originally reported in [60]. The most commonly used material is calcium titanate–neodymium aluminate [61], which has a relative permittivity of 45, a Q of greater than 20 000 at 2 GHz, and a temperature coefficient of resonant frequency of less than 1 ppm/°C for the $\mathrm{TE}_{01\delta}$ mode. A picture of a typical device is shown in Fig. 16.

Considerable research into alternative realizations of base-station filters has been carried out and an excellent review is found in [62]. Dual-mode dielectric-resonator filters, using a conductor loaded dielectric puck, enable filters with Q's of 5000 to be realized in approximately one-half the physical volume of coaxial filters [63]. A single-mode dielectric equivalent of the combine filter, using the TM_{01} mode, enabling the realization of smaller devices with Q's from 8000 to 12 000 is described in [64]. A spherical triple-mode resonator structure used as a smaller alternative to the $\mathrm{TE}_{01\delta}$ resonator is described in [65]. A TM_{110}-mode triple-mode filter is reported in [66], which gave a fivefold volume reduction compared with coaxial filters. An alternative reflection mode approach to triple-mode dielectric-resonator filters is presented in [67]. This approach simplifies the design complexity of triple-mode filters by splitting the filter into even- and odd-mode parts connected via a 3-dB hybrid. It gives greater than 50% volume reduction compared with single-mode dielectric-resonator filters, but is restricted to stopband attenuation levels of 20–30 dB. Single-mode dielectric-loaded waveguide filters, giving significant volume reduction compared to coaxial filters with Q's of 1000–2000, are described in [68].

Superconducting base-station filters are of interest because of their high Q realizable in a very small physical size. For example, a microstrip realization of a fifth-degree Chebyshev bandpass filter with 890-MHz center frequency and 0.3% bandwidth, occupied a surface area of approximately 5 cm^2 and exhibited 1-dB passband insertion loss [69]. A complete receiver front-end including 5-MHz bandwidth filters and integrated LNAs in the PCS band at 1.9 GHz is described in [70].

Recently, one of the main drivers for base-station filters has been the desire to reduce costs. This has resulted, for example, in the machined bodies of coaxial filters being replaced by castings. Miniaturization also reduces the cost of machined bodies, but often adds the costs of ceramic resonators. Other methods of cost reduction include the introduction of automated assembly and, more importantly, computer-aided and automatic filter alignment. One reported method of achieving this involves measuring the filter and then synthesizing its equivalent low-pass prototype. This is then compared with the ideal model of the filter [71], [72]. Another method uses the reflected impulse response of the filter to determine which resonators are mistuned [73].

The filters used in cellular radio handsets have completely different requirements. The original analog handsets in the 1980s were large, bulky, and manufactured in relatively small volumes. However, these phones used an FDMA access scheme, thus, they were transmitting and receiving simultaneously. The electrical specifications for the transmit/receive diplexer were quite severe, with the transmit filter requiring 50-dB attenuation in the receive band and vice-versa. The loss specifications required unloaded Q factors of several hundred. Such specifications could be met by helical resonator filters, which use resonant inductors [74]. These devices have been made more or less obsolescent by ceramic filter technology, although small volumes are still used in 450-MHz Scandinavian systems. The ceramic filters are dielectric loaded TEM filters where the resonators are formed by metallized holes inside a plated monoblock of high-permittivity ceramic [75]. These devices give a higher Q per unit volume than helical filters and became the dominant technology for a number of years.

Handsets for second-generation TDMA systems, such as GSM, transmit and receive in different time slots. Thus, the transmit/receive diplexer can be replaced by a switch and a receive filter. The purpose of the receive filter is to protect the LNA and the mixer in the down-converter from being overdriven by extraneous signals. For example, this situation may occur if two mobiles are being operated simultaneously while in close proximity to each other. Typically, the specified stopband attenuation requirements of these filters are 15–20 dB. However, the specifications on size and price are very demanding [76]. Recent advances in SAW filter design have enabled them to compete in this market. Their main advantage is very small size (typically $3 \times 3 \times 1$ mm) and low cost. Typically, they have 3-dB insertion loss and 2-W power handling. The most significant advance in SAW technology has been the replacement of the conventional transversal designs by SAW resonators, which are formed between acoustically reflective gratings on the surface of a SAW crystal [77], as shown in Fig. 17. Here, energy is coupled in and out of the structure by placing a transducer between the gratings. The transducer has only a few fingers and is relatively broad band. The transducer couples waves into both directions, which avoids the insertion loss problems of conventional designs. In addition, a leaky SAW wave is used, which propagates faster than a Rayleigh wave, enabling higher frequency designs to be produced. Also, higher values of electromechanical coupling enable broader bandwidth designs to be produced. Finally, leaky SAWs penetrate deep into the crystal, enabling higher power-handling capacity. Although these remarkable devices offer small size and low cost, their power handling and temperature stability is poor when compared with ceramic filters. Thus, they may not be suitable for third-generation systems where the transmitter and receiver operate simultaneously. However, recent developments in film bulk acoustic resonator (FBAR) devices show very impressive performance [78], [79].

Recent advances in micromachined electromechanical systems (MEMS) have demonstrated that this technology may be suitable for handset filters. For example, micromechanical resonators with Q's of 7450 at 100 MHz have been demonstrated [80]. MEMS switches have also been used for tuning the values of lumped components within filters [81]. The use of tunable filters may be useful in handsets for future systems operating at many different frequencies. An alternative technology for tuning of filter resonators is to use ferroelectric materials with tuning accomplished by an applied electric field [82], [83]. Tuning bandwidths of up to 30% with resonator Q's of 300 are possible using these methods, although the required high field strengths of several megavolts/meter require the use of extremely thin films.

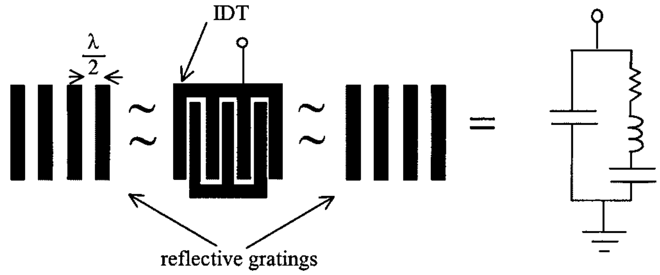

Fig. 17. SAW resonator.

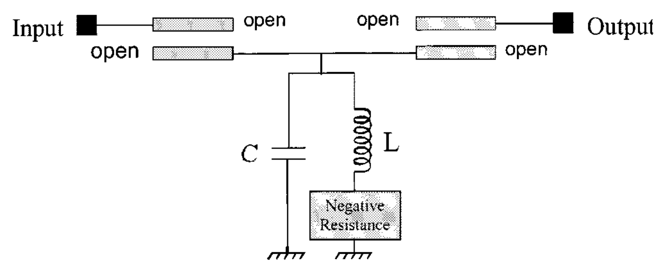

Fig. 18. Schematic of an active negative-resistance filter.

All the miniature handset filter technologies thus far discussed use passive resonators. Alternatively, several workers are investigating the use of active filters. One design approach is to compensate for the losses in physically small resonators by cancelling them with negative resistance. The negative resistance can be achieved by two distinct methods. In the first method [84], the negative resistance is achieved by connecting a series LC resonator to the drain of a single common-source transistor, with a shunt LC resonator connected to the source. Alternatively, two transistors may be connected in a feedback configuration [85]. This method has been demonstrated in the 3.8–4.2-GHz band. A spiral monolithic microwave integrated circuit (MMIC) inductor is then cascaded with this negative impedance converter to obtain a high Q inductor of 1.5 nH. This active inductance is then used to design bandpass filters. The schematic of the filter is shown in Fig. 18. This filter was realized as a MMIC in an area of 5 mm^2. An alternative technique called active matching [86], [87] used pseudomorphic high electron-mobility transistors (pHEMTs) in common-gate and common-drain configuration to match a passive filter network (Fig. 19). With this approach, an active bandpass filter operating at 31.825 GHz with 1.5-GHz 3-dB bandwidth has been demonstrated (Fig. 20). The filter had a noise figure of 4 dB at the center frequency and a 1-dB compression point of +3 dBm. It is also possible to construct active microwave filters using low-frequency techniques. For example, recursive and transversal filters have been demonstrated [88], [89]. Recursive filters require constant time delay increments, amplitude weighting elements, and signal combiners. These have been realized at microwave frequencies using low-pass elements

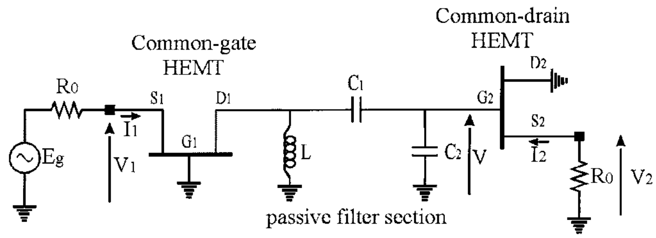

Fig. 19. Active matching method.

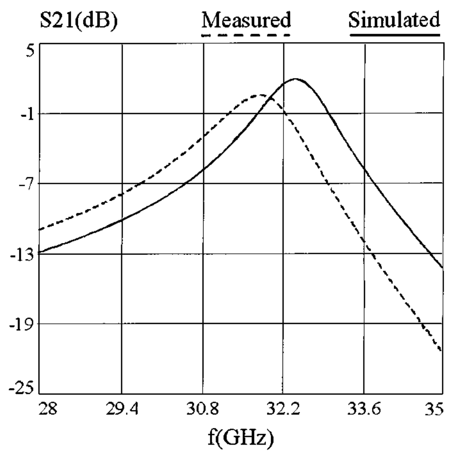

Fig. 20. Measured response of a millimeter-wave active filter.

for the time delays, amplifiers as the weighting elements, and lumped-element power dividers to realize the signal combiners. The layout of such a device, realized on gallium arsenide (GaAs), is shown in Fig. 21. The use of analog varactor tuned phase shifters also enables the realization of tunable active filters [90]. A filter that is tunable from 7 to 9.5 GHz with a constant gain of 3.5 dB and 10-dB return loss has been demonstrated. This device had a noise figure of 5 dB and a 1-dB compression point of -10 dBm.

The noise performance of active filters can be problematic and, for this reason, it is important to have a good theoretical tool for noise modeling and optimization. This may be performed using the noise wave formalism described in [91]. Another potential problem is that of nonlinearity. Active devices are inherently nonlinear and passive filters exhibit passive voltage or current magnification. For example, parallel tuned circuits exhibit a voltage gain that is inversely proportional to their fractional bandwidth. A study of active negative resistance filters [92] has shown that their third-order intercept point reduces by 6 dB each time the filter bandwidth is halved. An alternative approach is to use a low-gain LNA followed by a low-Q lossy bandpass filter, which has been correctly designed to maintain a sharp response in the presence of resonator losses [93]. The gain of the amplifier is set just high enough to establish a reasonable noise figure. The remainder of the required gain is then placed at the output of the filter in order to protect it from large out-of-band signals.

Finally, on the subject of power handling, it is worth noting that the RF voltage (or field strength) on each resonator of a coupled resonator filter is not distributed uniformly amongst the resonators. Typically, in Chebyshev bandpass filters, the second resonator experiences the largest peak voltage, at frequencies close to the passband edge [94]. By applying suitable rotations

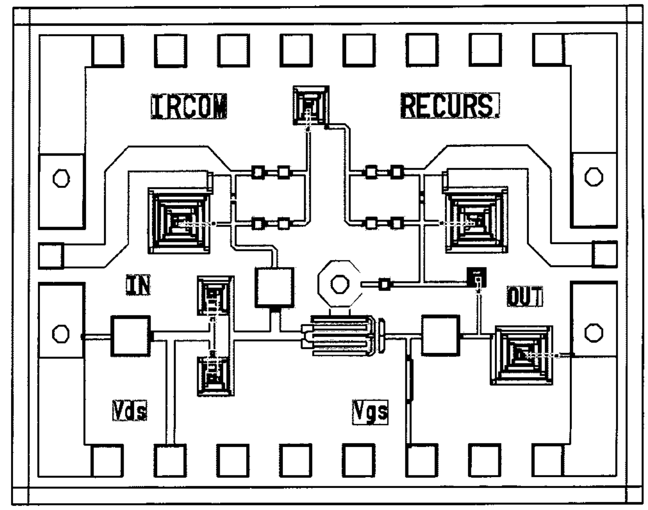

Fig. 21. Layout on GaAs of an active recursive filter.

to the admittance matrix of the low-pass prototype, it is possible to equalize the maximum voltage on each resonator, thus reducing the maximum voltage within the filter [95]. Techniques of this type may be useful in order to maximize the power handling of all the miniature filter technologies described above.

V. CONCLUSIONS

In this paper, an attempt has been made to relate some of the numerous advances in filter technology to the applications that have driven them. This has shown a continuous development in both theoretical filter design methods and in the technology used to implement them. On the theoretical front, initial work concentrated on the accurate design of waveguide and coaxial filters with Chebyshev responses. This was followed by major advances in the synthesis of filters with generalized transfer functions, dual-mode prototype networks, and contiguous multiplexers. More recently, automatic tuning, synthesis of active and lossy filters, and design of prototype networks with optimum power handling have taxed the theoreticians. On the technological front, realizations have changed from metallic waveguide and coaxial structures to ceramic resonators. More recently, more exotic structures such as superconducting, FBAR, MMIC, and MEMS filters are being investigated. Although microwave filters are often described as a mature technology, it can be seen that this is not the case and, hopefully, future applications will stimulate further advances in this exciting field.

REFERENCES

[1] M. E. Van-Valkenburg, *Introduction to Network Synthesis*. New York: Wiley, 1966.
[2] I. C. Hunter, *Theory and Design of Microwave Filters*, ser. Electromagnetic Wave 48. London, U.K.: IEE Press, 2001, pp. 49–100.
[3] J. D. Rhodes, *Theory of Electrical Filters*. New York: Wiley, 1975.
[4] J. D. Rhodes and I. H. Zabalawi, “Design of selective linear phase filters with equiripple amplitude characteristics,” *IEEE Trans. Circuits Syst.*, vol. CAS-25, pp. 989–1000, Dec. 1978.
[5] R. Levy, R. V. Snyder, and G. Matthaei, “Design of microwave filters,” *IEEE Trans. Microwave Theory Tech.*, vol. 50, pp. 783–793, Mar. 2002.
[6] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance Matching Networks and Coupling Structures*. Norwood, MA: Artech House, 1964, pp. 674–675.

[7] J. O. Scanlan, "Theory of microwave coupled-line networks," *Proc. IEEE*, vol. 68, pp. 209–231, Feb. 1980.

[8] G. L. Matthaei, "Interdigital band-pass filters," *IRE Trans. Microwave Theory Tech.*, vol. MTT-10, pp. 479–491, Nov. 1962.

[9] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance Matching Networks and Coupling Structures*. Norwood, MA: Artech House, 1964, pp. 165–173, 243–252.

[10] R. M. Fano and A. W. Lawson, *Microwave Transmission Circuits*, ser. M.I.T. Rad. Lab.. New York: McGraw-Hill, 1948, vol. 9, ch. 9, 10.

[11] J. B. Tsui, *Microwave Receivers With Electronic Warfare Applications*. New York: Wiley, 1992.

[12] I. C. Hunter, *Theory and Design of Microwave Filters*, ser. Electromag. Wave 48. London, U.K.: IEE Press, 2001, pp. 201–212.

[13] N. Marcuvitz, *Waveguide Handbook*. Stevenage: IEE, 1986, pp. 249–279.

[14] S. B. Cohn, "Direct-coupled resonator filters," *Proc. IRE*, vol. 45, pp. 187–196, Feb. 1957.

[15] L. Young, "Direct-coupled cavity filters for wide and narrow band-widths," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-11, pp. 162–178, May 1963.

[16] R. Levy, "Theory of direct coupled-cavity filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-15, pp. 340–348, June 1967.

[17] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance Matching Networks and Coupling Structures*. Norwood, MA: Artech House, 1964, pp. 889–920.

[18] S. B. Cohn, "Parallel-coupled transmission-line resonator filters," *IRE Trans. Microwave Theory Tech.*, vol. MTT-10, pp. 223–231, Apr. 1958.

[19] G. L. Matthaei, "Comb-line filters of narrow or moderate bandwidth," *Microwave J.*, vol. 6, pp. 82–91, Aug. 1963.

[20] I. C. Hunter and J. D. Rhodes, "Electronically tunable microwave bandpass filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-30, pp. 1354–1360, Sept. 1982.

[21] R. J. Wenzel, "Synthesis of combline and capacitively-coupled interdigital filters of arbitrary bandwidth," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 678–686, Aug. 1971.

[22] H. L. Schumacher, "Coax multiplexers: Key to EW signal sorting," *Microwave Syst. News*, pp. 89–93, Aug. 1976.

[23] P. M. La-Tourette, "Multi-octave combline-filter multiplexers," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1977, IEEE cat. 77CH1219-5 MTT, pp. 298–301.

[24] P. M. La-Tourette and J. L. Roberds, "Extended-junction combline multiplexers," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1978, IEEE cat. 78CH1355-7 MTT, pp. 214–216.

[25] R. Levy and J. D. Rhodes, "A combline elliptic filter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-19, pp. 26–29, Jan. 1971.

[26] J. P. Rooney and L. M. Underkofler, "Printed circuit realization of MW filters," *Microwave J.*, vol. 21, pp. 68–73, Sept. 1978.

[27] J. D. Rhodes, "Suspended substrates provide alternative to coax," *Microwave Syst. News*, vol. 9, pp. 134–143, Aug. 1979.

[28] J. D. Rhodes and J. E. Dean, "MIC broadband filters and contiguous multiplexers," in *Proceedings of the 9th European Microwave Conference*. Kent, U.K.: Microwave Exhibitions and Publishers, 1979, pp. 407–411.

[29] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance Matching Networks and Coupling Structures*. Norwood, MA: Artech House, 1964, pp. 1027–1086.

[30] C. Kudsia, R. Cameron, and W. Tang, "Innovations in microwave filters and multiplexing networks for communication satellite systems," *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 1133–1149, June 1992.

[31] C. Kudsia and M. V. O'Donovan, "A light weight graphite fiber epoxy composite (GFEC) waveguide multiplexer for satellite applications," in *Proc. 4th Eur. Microwave Conf.*, Montreaux, Switzerland, Sept. 1973.

[32] J. E. Keigler, "RCA Satcom: An example of weight-optimized, satellite design for maximum communications capacity," in *Acta Astronautics*. New York: Pergamon, 1978, vol. 5.

[33] W. G. Lin, "Microwave filters employing a single cavity excited in more than one mode," *J. Appl. Phys.*, vol. 22, pp. 989–1001, Aug. 1951.

[34] A. E. Atia and A. E. Williams, "New types of waveguide bandpass filters for satellite transponders," *Comsat Tech. Rev.*, vol. 1, no. 1, pp. 21–43, Fall 1971.

[35] ——, "Narrow-bandpass waveguide filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-20, pp. 258–265, Apr. 1972.

[36] A. E. Atia, A. E. Williams, and R. W. Newcom, "Synthesis of dual-mode filters," *IEEE Trans. Circuits Syst.*, vol. CAS-21, pp. 649–655, Sept. 1974.

[37] J. D. Rhodes, "The generalized direct-coupled cavity linear phase filter," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-18, pp. 308–313, June 1970.

[38] M. Guglielmi, O. Roquebrun, P. Jarry, E. Kerherve, M. Capurso, and M. Piloni, "Low-cost dual-mode asymmetric filters in rectangular waveguide," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Phoenix, AZ, June 2001, pp. 1787–1790.

[39] W. C. Tang and S. K. Chaudhuri, "A true elliptic-function filter using triple-mode degenerate cavities," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Boston, MA, May 1983, pp. 83–85.

[40] R. R. Bonetti and A. E. Williams, "Application of dual TM-modes to triple and quadrupole mode filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-35, pp. 1143–1149, Dec. 1987.

[41] S. Kallianteris and M. V. O'Donovan, "Technology advances in the realization of filter networks for communication satellites operating above 10 GHz," in *Proc. AIAA/CASI 6th Commun. Satellite Syst. Conf.*, Montreal, QC, Canada, Apr. 1981.

[42] R. Tong and C. M. Kudsia, "Enhanced performance and increased EIRP in communications satellites using contiguous multiplexers," in *Proc. 10th AIAA Commun. Satellite Syst. Conf.*, Orlando, FL, Mar. 1984, pp. ???–???.

[43] A. E. Atia, "Computer-aided design of waveguide multiplexers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-24, pp. 332–336, Mar. 1974.

[44] J. D. Rhodes and R. Levy, "A generalized multiplexer theory," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 99–111, Feb. 1979.

[45] C. M. Kudsia and M. N. S. Swamy, "Computer-aided optimization of microwave filter networks for space applications," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Washington, DC, May 1980, pp. 410–412.

[46] J. D. Rhodes and I. H. Zabalawi, "Synthesis of symmetric dual-mode in-line prototype networks," *Int. J. Circuit Theory Applicat.*, vol. 8, no. 2, pp. 145–160, 1980.

[47] R. J. Cameron, "Fast generation of Chebyshev filter prototypes with asymmetrically prescribed transmission zeros," *ESA J.*, vol. 6, pp. 83–95, 1982.

[48] R. J. Cameron and J. D. Rhodes, "Asymmetric realizations for dual-mode bandpass filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-29, pp. 51–58, Jan. 1981.

[49] S. J. Fiedziuszko, I. C. Hunter, T. Itoh, Y. Kobayashi, T. Nishikawa, S. N. Stitzer, and K. Wakino, "Dielectric materials, devices, and circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 50, pp. 706–720, Mar. 2002.

[50] S. J. Fiedziuszko, "Dual-mode dielectric resonator loaded cavity filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-30, pp. 1311–1316, Sept. 1982.

[51] M. J. Lancaster, *Passive Microwave Device Applications of High-Temperature Superconductors*. Cambridge, U.K.: Cambridge Univ. Press, 1997.

[52] R. R. Mansour, S. Ye, S. Peik, B. Jolley, V. Dokas, T. Romano, and G. Thomson, "HTS filter technology for space applications," presented at the IEEE MTT-S Filter Technol. Commun. Syst. Workshop, Baltimore, MD, June 1998.

[53] C. Lascaux, F. Rouchaud, V. Madrangeas, M. Aubourg, P. Guillon, B. Theron, and M. Maignan, "Planar Ka-band high temperature superconducting filters for space applications," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Phoenix, AZ, June 2001, pp. 487–490.

[54] R. C. Peach, "SAW filters for space applications," presented at the IEEE MTT-S Filter Technol. Commun. Syst. Workshop, Baltimore, MD, June 1998.

[55] C. Kudsia, R. Cameron, and W. Tang, "Innovations in microwave filters and multiplexing networks for communications satellite systems," *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 1133–1149, June 1992.

[56] R. C. Chapman *et al.*, "Hidden threat: Multi-carrier passive component IM generation," in *AIAA/CASI 6th Commun. Satellite Syst. Conf.*, Montreal, QC, Canada, Apr. 1976, paper 76-296.

[57] I. A. Glover and P. M. Grant, *Digital Communications*. Englewood Cliffs, NJ: Prentice-Hall, 1997, pp. 146–149.

[58] I. C. Hunter, *Theory and Design of Microwave Filters*, ser. Electromag. Wave Series 48. London, U.K.: IEE Press, 2001, pp. 90–99.

[59] H.-W. Yao and A. E. Atia, "Temperature characteristics of combline resonators and filters," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Phoenix, AZ, June 2001, pp. 1475–1478.

[60] S. B. Cohn, "Microwave bandpass filters containing high-Q dielectric resonators," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-16, pp. 218–272, Apr. 1968.

[61] W. Wersing, "Microwave ceramics for resonators and filters," *Current Opinion Solid State Phys. Mater. Sci.*, vol. 1, pp. 715–729, 1996.

[62] K. Wakino, T. Nishikawa, and Y. Ishikawa, "Miniaturization techniques for dielectric resonator filters for mobile communications," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 1295–1300, July 1994.

[63] I. C. Hunter, J. D. Rhodes, and V. Dassonville, "Dual-mode filters with conductor loaded dielectric resonators," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 2304–2311, Dec. 1999.

[64] C. Wang, K. A. Zaki, A. E. Atia, and T. G. Dolan, "Dielectric combline resonators and filters," *IEEE Trans. Microwave Theory Tech.*, vol. 46, pp. 2501–2506, Dec. 1998.

[65] H. Tanaka, H. Nishida, and Y. Ishikawa, "Spherical dielectric resonator filter coupled with NRD guide," in *Proc. IEICE Jpn. Spring Conf.*, 1991, C-103.

[66] T. Nishikawa, K. Wakino, H. Wada, and Y. Ishikawa, "800 MHz channel dropping filter using TM_{110} triple mode resonance," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1985, K-5, pp. 289–292.

[67] I. C. Hunter, J. D. Rhodes, and V. Dassonville, "Triple mode dielectric resonator hybrid reflection filters," *Proc. Inst. Elect. Eng.*, pt. H, vol. 145, pp. 337–343, 1998.

[68] V. Walker and I. C. Hunter, "Dielectric loaded waveguide filters," *Proc. Inst. Elect. Eng.*, pt. H, vol. 148, no. 2, pp. 91–96, Apr. 2001.

[69] D. Zhang, G.-C. Liang, C. F. Shih, M. E. Johansson, and R. S. Withers, "Narrow-band lumped-element microstrip filters using capacitively loaded inductors," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 3030–3036, Dec. 1995.

[70] E. Soares, K. F. Raihn, and J. D. Fuller, "Dual 5 MHz PCS receiver front-end," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Phoenix, AZ, June 2001, pp. 1981–1984.

[71] M. Kahrizi, S. Safavi-Naeini, and S. K. Chaudhuri, "Computer diagnosis and tuning of microwave filters using model-based parameter estimation and multi-level optimization," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Boston, MA, June 2000, pp. 1641–1644.

[72] D. Ibbetsen, "A synthesis based approach to automated filter tuning," in *IEE Microwave Filters Multiplexers Colloq.*, London, U.K., Nov. 2000, Paper 00/117, pp. 11/1–11/3.

[73] J. Dunsmore, "Simplify filter tuning in the time domain," *Microwaves RF*, vol. 38, no. 4, pp. 68–84, Mar. 1999.

[74] N. P. Spencer, "A review of the design of helical resonator filters," *J. Inst. Electron. Radio Eng.*, vol. 57, no. 5, Sept. 1987.

[75] S. Kobayashi and K. Saito, "A miniaturized ceramic bandpass filter for cordless phone systems," in *IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 2, 1995, pp. 391–394.

[76] K. Wakino, "Miniaturization trend of filters for mobile communication handsets," presented at the IEEE MTT-S Int. Microwave Symp. Miniaturization Filters Commun. Handsets Workshop, Anaheim, CA, June 1999.

[77] T. Tagami, H. Ehera, K. Noguchi, and T. Komaski, "Resonator type SAW filter," *Oki Tech. Rev.*, vol. 63, p. 59, 1997.

[78] J. D. Larson, R. C. Ruby, P. Bradley, J. Wen, S. Kok, and A. Chien, "Power handling and temperature coefficient studies in FBAR duplexers for 1900 MHz PCS band," in *IEEE Ultrason. Symp. Dig.*, 2000, pp. 869–874.

[79] R. Weigel, D. P. Morgan, J. M. Owens, A. Ballato, K. M. Lakin, K. Hashimoto, and C. C. W. Ruppel, "Microwave acoustic materials, devices, and applications," *IEEE Trans. Microwave Theory Tech.*, vol. 50, pp. 738–749, Mar. 2002.

[80] C. T.-C. Nguyen, "Transceiver front-end architectures using high-Q micromechanical resonators," presented at the IEEE Eur. MIDAS MEMS for High-Q Filters Workshop, Surrey, U.K., July 2000.

[81] D. Peroulis, S. Pacheco, K. Sarabandi, and L. P. B. Katehi, "Tunable lumped components with applications to reconfigurable MEMS filters," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Phoenix, AZ, 2001, pp. 341–344.

[82] M. J. Lancaster, J. Powell, and A. Porch, "Thin-film ferroelectric microwave devices," *Superconduct. Sci. Technol.*, no. 11, pp. 1323–1334, 1998.

[83] I. Vendik, O. Vendik, V. Sherman, A. Svishchev, V. Pleskachev, and A. Kurbanov, "Performance limitation of a tunable resonator with a ferroelectric capacitor," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Boston, MA, June 2000, pp. 1371–1374.

[84] A. Brucher, C. Cenac, M. Delmond, P. Meunier, L. Billonet, B. Jarry, P. Guillon, and S. E. Sussman-Fort, "Improvement of microwave planar active filters with MMIC technology," in *Proc. Eur. GaAs Related III–IV Compounds Applicat. Symp.*, Apr. 1994, pp. 315–318.

[85] S. E. Sussman-Fort, "An NIC-based negative resistance circuit for microwave active filters," *Int. J. Microwave Millimeter-Wave Computer-Aided Eng.*, vol. 4, no. 2, pp. 130–139, Apr. 1994.

[86] W. Mouzannar, L. Billonet, B. Jarry, and P. Guillon, "A new design concept for wide-band frequency tuneable and high order MMIC active recursive filters," *Microwave Opt. Technol. Lett.*, vol. 24, no. 6, pp. 380–385, Mar. 2000.

[87] K. B. Niclas, "Active matching with common-gate MESFET," *IEEE Trans. Microwave Theory Tech.*, vol. 43, pp. 492–499, 1996.

[88] C. Rauscher, "Microwave active filter based on transversal and recursive principles," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, Dec. 1985.

[89] L. Billonet *et al.*, "Design concepts for microwave recursive and transversal filters using Lange couplers," in *IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 2, June 1992, pp. 925–928.

[90] M. Delmond, L. Billonet, B. Jarry, and P. Guillon, "Microwave tuneable active filter design in MMIC technology using recursive concepts," in *IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 3, May 1995, pp. 105–108.

[91] S. Wedge and D. Rutledge, "Wave techniques for noise modeling and measurement," *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 2004–2012, Nov. 1992.

[92] I. C. Hunter and S. R. Chandler, "Inter-modulation in active microwave filters," *Proc. Inst. Elect. Eng.*, pt. H, vol. 145, no. 1, pp. 7–12, 1998.

[93] J. D. Rhodes and I. C. Hunter, "Synthesis of reflection-mode prototype networks with dissipative circuit elements," *Proc. Inst. Elect. Eng.*, pt. H, vol. 144, no. 2, pp. 97–104, 1998.

[94] C. Wang and K. A. Zaki, "Analysis of power handling capacity of band pass filters," in *IEEE MTT-S Int. Microwave Symp. Dig.*, Phoenix, AZ, June 2001, pp. 1611–1614.

[95] C. Ernst, V. Postoyalko, R. Parry, and I. C. Hunter, "Filter topology with minimum peak stored energy," in *IEEE MTT-S Int. Microwave Symp.*, Phoenix, AZ, June 2001, pp. 1631–1634.

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