Miniature Microwave Filters for Communication Systems

Ian C. Hunter, *Senior Member, IEEE*, Stephen R. Chandler, David Young, and Andrew Kennerley

Abstract—**This paper will present new developments in miniature integrated microwave filters suitable for application in mobile communication handsets, wireless LAN's and microwave fixed links. Recent measured results for active microwave filters will be presented with emphasis on their nonlinear behavior. The use of high permittivity ceramic substrates in parallel plate TEM mode stripline filters is described and recent measured results will be presented for a prototype 2-GHz filter occupying a volume of 0.5-cubic centimeters. E-Plane filters for millimeter wave applications will be discussed and measured results will be presented for a novel E-Plane filter with asymmetric frequency response which was realized in a planar structure. Finally, a new type of ceramic loaded E-Plane filter will be proposed**

I. Introduction

THE work described in this paper was driven by the need for new and possibly improved methods of manufacturing miniature microwave filters. Such components find increasing application as transmit/receive diplexers and intermediate frequency filters in cellular radios, wireless LAN's and microwave fixed links. These systems cover a broad range of the spectrum from UHF to EHF. In the UK the analogue cellular system (TACS) handsets use bandwidth around 900 MHz and the digital microcell Personal Communications Network (PCN) handsets operate around 1.8 GHz. PCN base stations will be connected via line of sight links at 38 and 58 GHz . Wireless LAN's operating in the deregulated 2.4-2.5-GHz band are being produced and the DTI has authorised a band at 5.2 GHz for Hiperlan. All of these systems will require the large scale production of microwave bandpass filters. Demanding specifications are placed on these filters, they must be miniaturised while having low passband insertion loss, narrow passband bandwidth, high skirt selectivity and high stopband attenuation. They must also be cheap. Factors of cost and miniaturisation dictate the use of integrated surface mount filters operable over the broad frequency range described.

Various integrated filter technologies will be discussed in this paper with the exception of Surface Acoustic Wave (SAW) filters which suffer from frequency limitations. The next section will consider the use of active filter technology with emphasis upon their large signal behavior.

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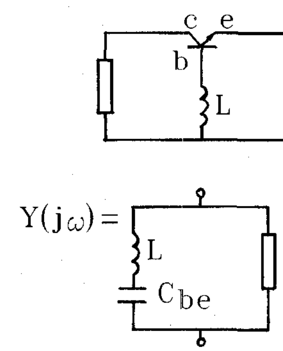

$$G(\omega) = \frac{-g_m}{\left(\frac{\omega^2}{\omega_0^2} - 1\right)}$$

$$\omega_0 = \frac{1}{\sqrt{LC_{be}}}$$

Fig. 1. Active negative resistance circuit.

II. Active Microwave Filters

Active filters are a mature technology which on the surface appears an attractive approach for miniature integrated circuit microwave filters. The circuitry is compatible with monolithic microwave integrated circuit (MMIC) technology and potentially provides a miniaturised solution, enabling a transmit/receive diplexer to be realized as part of an MMIC. Low frequency active filters use combinations of resistors, capacitors and operational amplifiers, thus avoiding the need for large low Q inductors. In principle this technique could be extended to microwave frequencies. However, this would require active devices with a large gain bandwidth product at the operating frequency in order to approximate anything like an ideal operational amplifier. Thus an MMIC active filter based upon this approach might need 20 GHz technology to realize a filter with a 1-GHz center frequency. Second, inductive microwave circuit elements are not particularly problematic since these are achievable in a small physical size using either lumped or distributed techniques.

In order to avoid high frequency operational amplifiers, the use of active negative resistance filters was considered. This method uses a conventional microwave filter consisting of N coupled resonators and N microwave transistors [1]. Each transistor is combined with a matching circuit such that the resultant one port network has an input impedance with a negative real part at the center frequency of the filter (Fig. 1). Each one port network is then coupled into one of the filters' resonators. In this way the resonator resistance of a low Q transmission medium such as microstrip on GaAs is cancelled by the negative resistance of the active devices.

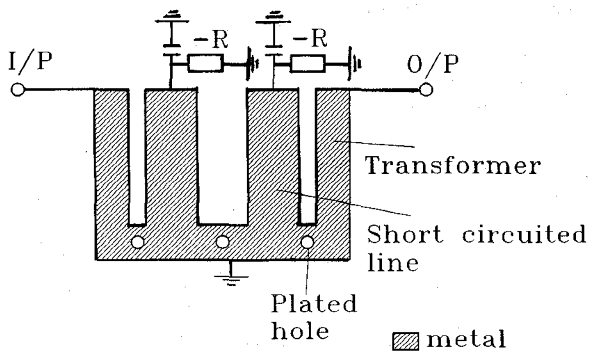

Fig. 2. Active microstrip bandpass filter.

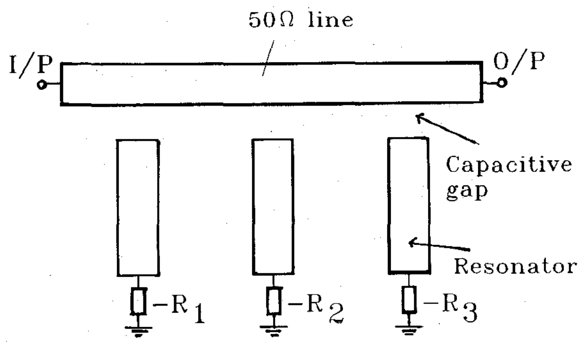

Fig. 3. Active microstrip bandstop filter.

The advantage of this approach is that the active devices only require gain at the center frequency of the filter. This approach has been applied to the design of narrow band microwave bandpass and bandstop filters. A bandpass filter of this type was constructed using a microstrip combline filter and bipolar transistors (Fig. 2) and exhibited approximately zero midband insertion loss [2]. Similarly a microstrip bandstop filter (Fig. 3) exhibited high selectivity and stopband attenuation [3]. It would be relatively simple to miniaturize these circuits using a lumped element realization. At first glance these filters appear to exhibit infinite unloaded resonator Q factor. Unfortunately their performance degrades with increased input signal power due to power saturation within the active devices. Saturation causes a gradual reduction in filter Q which manifests itself as increased passband attenuation for a bandpass filter or decreased stopband attenuation within a bandstop filter [3]. Under some circumstances this may be acceptable, but more importantly, the nonlinear device characteristics which cause saturation also produce inter- modulation products which can be within the filter passband. A typical value of two tone third order intercept point (IP3) for a bandpass filter was 10 dBm at 5-mA bias. An active filter whose sole purpose is to reject unwanted signals will thus generate other unwanted signals. It is possible to increase the intercept point by increasing the bias current supplied to the active filter (Fig. 4) and a 6 dB increase in intercept point will be obtained by doubling the current. However this is not always a practical proposition. It is also worth noting that intermodulation becomes more pronounced in narrow band filters because resonator current magnification causes a 6 dB per octave reduction in intercept point as the filter bandwidth is reduced. In addition active filters have finite noise figure. Measured results for bandpass filters were 4–7 dB

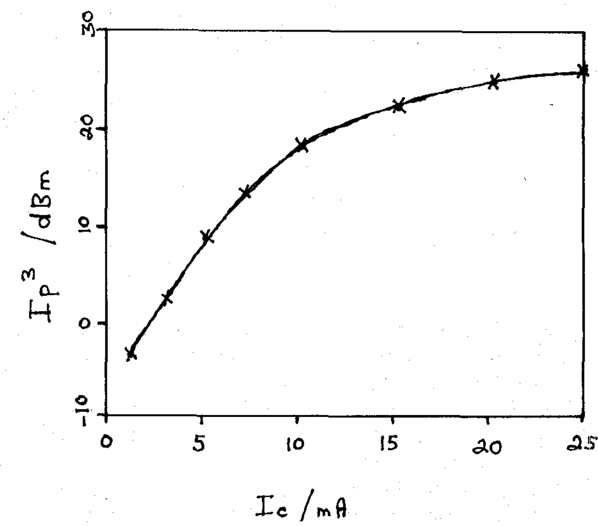

Fig. 4. Third-order intercept point of active bandpass filter.

resulting from the insertion loss of the passive part of the filter plus a contribution from the noise figure of the active devices.

From a mobile communications system point of view the receiver noise figure is not particularly important as these systems are interference rather than noise limited. However, nonlinear effects in a transmit/receive diplexer will affect the overall system performance. For example the high level transmit signal could mix with an incoming signal to produce an intermodulation product which is then mixed down to the I. F. stage. Another unwanted effect is that of cross modulation where the modulation on the transmitted carrier is transferred on to the received carrier. Simulated results for cross modulation in the bandstop filter were generated using the harmonic balance simulator in Academy. The input power at which 1% cross modulation was observed is plotted as a function of device bias in Fig. 5.

It has been shown that active filters have potential application in communication systems but nonlinear effects caused by device saturation will affect system performance. An alternative approach is the use of miniature integrated passive filters. The next section of this paper will discuss the realization of such filters using high dielectric constant ceramic materials to achieve miniaturisation.

III. High Dielectric Constant Ceramic Microwave Filters

Microwave filters presently used in cellular radio handsets are commonly manufactured using high dielectric constant ceramic materials. The use of a high dielectric constant typically in the range of 36-40 reduces the wavelength of electromagnetic waves within the structure by the square root of the relative dielectric constant of the material. Consequently a transmission line filter realized in such a material will have all dimensions reduced by a factor of six or more compared to an air dielectric device. Most presently manufactured devices used for cellular radio handsets employ a rectangular block of ceramic containing cylindrical holes [4], [5]. The external surfaces of the structure and the interiors of the holes are selectively metallised (Fig. 6). The composite structure is

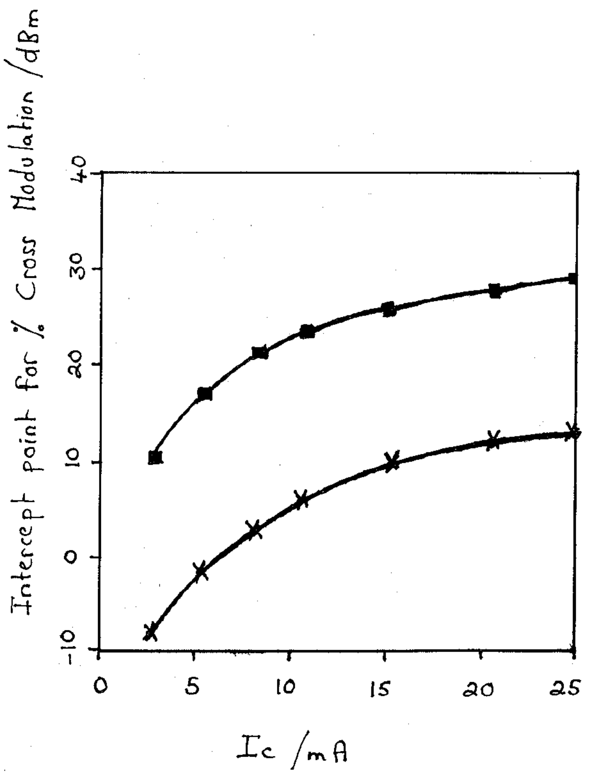

Fig. 5. Cross modulation in active bandstop filter.

equivalent to an array of parallel coupled transverse electromagnetic (TEM) mode transmission lines which can be designed to have a bandpass filter response. For example, by insulating opposite ends of adjacent holes from the ground the resultant structure is an interdigital filter. In this case the length of the holes should be one quarter wavelength within the dielectric at the filter center frequency and the spacing between the holes determines the filter bandwidth. The use of lumped capacitors in combination with the ceramic block to produce a combline filter has also been reported [6]. These filters suffer from the disadvantage that the external dimensional tolerances of the ceramic block are linearly proportional to resonant transmission line lengths. Thus a tight tolerance on filter center frequency is difficult to achieve without some mechanical alignment or tuning of the filter. This would appear to be one reason why low cost filters of this type are not generally available from European manufacturers. The effect of a fixed dimensional tolerance on frequency accuracy increased as the square of filter center frequency. For example, a quarter wave resonator with εr of 38 and a +/− 0.1-mm dimensional tolerance would have a center frequency accuracy of approximately +/− 8 MHz at 1 GHz and +/− 200 MHz at 5 GHz which is obviously of little value.

A further difficulty is that the in-line resonator structure of these filters is unsuitable for the realization of certain filter transfer characteristics. For example, an asymmetric filter transfer function using cross couplings between nonadjacent resonators would result in increased filter thickness. This would be particularly disadvantageous if a surface mount filter were required for a PCMCIA card.

These above problems may be alleviated by printing the transmission line filter pattern on a thin microwave substrate.

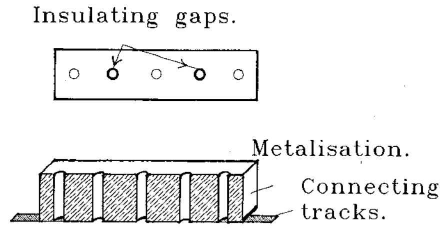

Fig. 6. Bandpass filter realization in monolithic ceramic block.

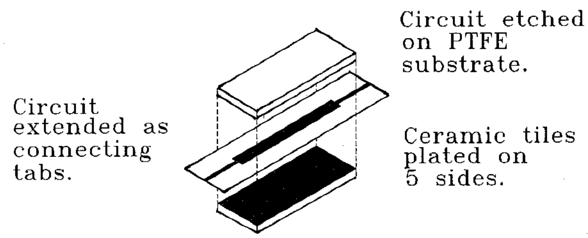

Fig. 7. Ceramic stripline configuration.

Frequency accuracy is then largely determined by photolithographic tolerances and the printed circuit allows an effectively planar filter realization for cross coupled circuits. A multilayer printed approach has been reported in [7] and further discussion on the printed approach will be presented in the nexxt section.

IV. Ceramic Stripline Filters

The combination of size reduction, good frequency tolerance and circuit design flexibility is possible using the stripline filter realization shown in Fig. 7. The filter is composed of a printed microwave substrate symmetrically located between two high dielectric constant tiles. The outside faces of the filter are metallised and the printed circuit extends beyond the edges of the ceramic tiles. Provided that the operating frequency is restricted so that the width of the structure is less than one half wavelength within the dielectric then the dominant mode of electromagnetic propagation is TEM and higher order waveguide modes are not excited. With appropriate substrate metallisation the composite structure may have a variety of filter transfer characteristics. The external dimensions of the ceramic tiles have no effect n resonator lengths and, hence, on the center frequency accuracy. They will affect the inter-resonator couplings and hence the filter bandwidth but this is of lesser importance. The resonant frequency accuracy is determined by the resonators' physical lengths and the tolerance on the dimensions of the ceramic substrates. For example, a quarter wave line resonant at 1 GHz with εr of 38 will have a length of 12.16 mm. A reasonable printed circuit dimensional tolerance of 0.01 mm gives a center frequency accuracy of +/− 0.8 MHz at 1 GHz and +/− 4 MHz at 5 GHz. The effect of dielectric constant variation must also be considered. This produces a frequency independent percentage tolerance with a 1% variation in permittivity causing a 0.5% variation in frequency due to the square root dependence of wavelength upon dielectric constant. An achievable dielectric constant of 38 +/− 0.25 will produce a +/− 3 MHz accuracy at 1 GHz and 15 MHz at 5 GHz.

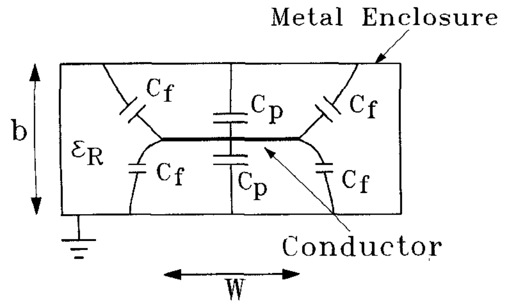

Fig. 8. Static capacitances in stripline.

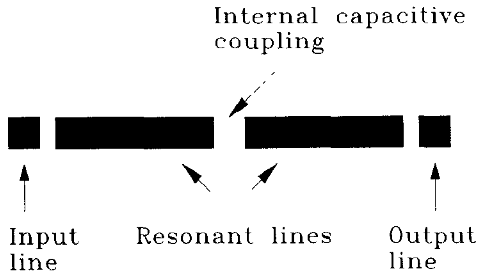

Fig. 9. End coupled bandpass filter.

The printed nature of the filter enables complex filter transfer characteristics to be realized by the appropriate use of cross coupling between nonadjacent resonators or by the use of extracted pole resonators.

The unloaded Q factor and consequent passband attenuation of the ceramic stripline filter can be calculated as follows. First we calculate the characteristic impedance of a single line from its static capacitance. The static capacitance c/t is the sum of the parallel plate and fringing capacitances as shown in Fig. 8. and

$$\frac{c}{\epsilon} = \frac{4w}{b} + 4CF$$

and for a TEM line

$$\sqrt{\epsilon r} Zo = \frac{377}{c/\epsilon}.$$

For a 4-mm ground plane with $\varepsilon R = 38$, a 3-mm-wide transmission line has a characteristic impedance of 12.7 Ω. The unloaded Q factor of the transmission medium Q_u is proportional to the ground plane spacing and the square root of the frequency of operation and is calculated from a graph in [8]. For the above transmission line this yields $\mathrm{Qu}_c = 500$ at 1 GHz assuming ideal copper conductors. Assuming a realistic Quc of 400 and a loss tangent of 0.001 we obtain an unloaded Q factor of 285. The midband insertion loss of a Chebychev bandpass filter can be calculated [9] using

$$\mathrm{I.L} \cong 8.686 \frac{\mathrm{C_N fo}}{\Delta \mathrm{fQu}}$$

where $Fo/\Delta F$ is the ratio of center frequency to bandwidth of the filter. C_N is a contant related to the degree and ripple level of the Chebychev network. For a fourth degree network with a 0.1-dB ripple level, $C_N = 2.4$ yielding a midband insertion loss for a 1 GHz filter with 50-MHz bandwidth of 1.4 dB. This is a reasonable level of loss for a mobile handset application.

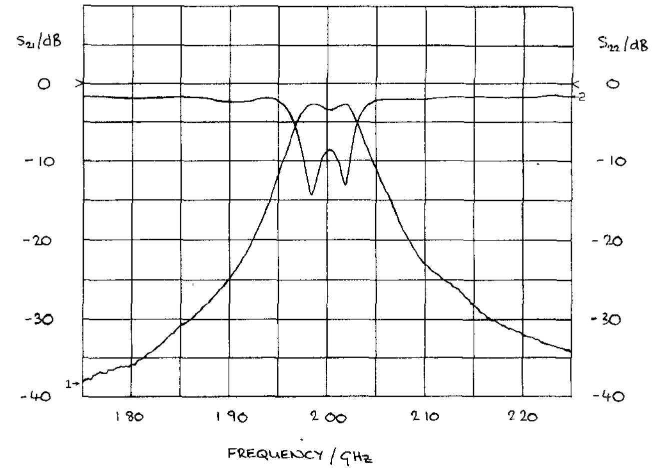

Fig. 10. Measured performance of ceramic stripline bandpass filter.

A large variety of stripline filter circuits are available to the designer. One of the simplest is the end coupled filter shown in Fig. 9. This consists of approximately half wavelength transmission lines separated by capacitive discontinuities. This circuit was chosen because it does not require any short circuit connections. It also has the advantage that it utilizes space well. The transmission lines can be designed to have a low impedance so that they occupy most of the filter's width. Consequently the lines will have maximum Q factor. Second, the effects of higher order mode propagation are minimized as the width of the filter is much less than one half wavelength at the filter center frequency. This maximizes the upper stopband bandwidth.

A two section bandpass filter with 50-MHz bandwidth at 2 GHz, center frequency was designed and constructed. The filter was constructed using 2-mm-thick ZTS substrates with a relative permittivity of 38. The external metallisation was copper plating. The dimensions of the filter were 25 × 6 × 4 mm. The circuit pattern was printed on to a 0.125 mm duroid substrate with a relative permittivity of 2.3. The input and output connections to the filter were achieved by extending the circuit beyond the end faces of the filter. The integrated filter can then drop into a rectangular slot in a sub-system circuit board. Further ground connections can be made to the filter by extending the circuit board beyond the sides of the filter.

Measured results for an experimental prototype filter are shown in Fig. 10. Although the device was slightly mismatched a midband insertion loss of approximately 2 dB was observed. Taking account of reflection losses the filter exhibited an unloaded Q factor of greater than 300.

The measured results are very encouraging, demonstrating a low cost miniature realization for filters in the 1–3-GHz range. However the half wave end coupled resonator approach may be physically too long if large number of sections were required. For example a sixth degree filter with a center frequency of 2 GHz using a ceramic with a relative permittivity of 38 would be at least 75 mm long. This problem is now being addressed by using coupled line filters. Initially the interdigital filter was considered but was discarded because the coupling between interdigital lines is strong thus narrow band

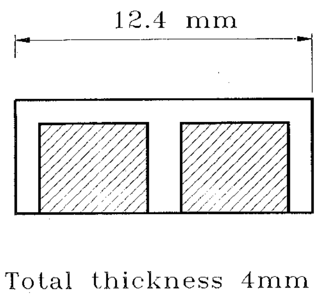

Fig. 11. Combline filter artwork.

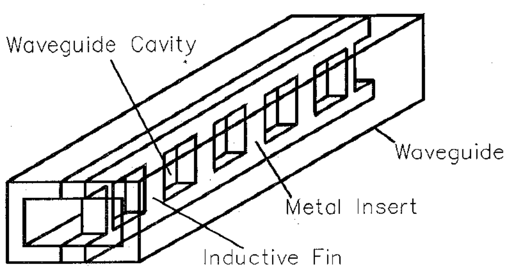

Fig. 12. Chebychev E-Plane filter configuration.

filters require large spacings between the interdigital lines. A better solution is to use a combline filter consisting of an array of coupled short circuited lines with lumped capacitances to ground terminating the open circuited ends of the lines. The capacitors can be realized using the fringing capacitance to ground from the open circuited ends of the transmission lines. As an example, a 3.5-mm-wide transmission line in a ground plane spacing of 4 mm and a relative permittivity of 38 has an impedance of 11.5 ohms and an end loading capacitance of approximately 1.5 pF. The required electrical length of the resonator for resonance at 2.5 GHz is then 75 degrees. This is approaching one quarter wavelength and thus the coupling impedance between resonators is relatively weak and consequently narrow band filters can be realized with small spacings between resonators. The only disadvantage introduced by this approach is that the filter will be more selective on the high frequency skirt than the low frequency skirt. This is due to the attenuation poles introduced by the couplings becoming resonant at the quarter wavelength frequency. The dimensions of a 2 pole filter centerd at 2.5 GHz are shown in Fig. 11. Work is now progressing on fabricating this filter using a thin film metal pattern printed directly onto the ceramic tiles.

V. E-Plane Filters

E-Plane technology is attractive at millimeter wavelengths where it offers a low cost relatively high Q miniature integrated waveguide transmission medium. The basic structure for an E-Plane bandpass filter is shown in Fig. 12. This has all transmission zeros at infinity realized by half wave lengths of uniform waveguide. These are separated by metal fins across the E-plane of the waveguide forming inductive coupling susceptances. Admittance inverters may be formed by combining the metal fins with appropriate negative lengths of waveguide. This type of structure is simple and is suitable for Chebychev bandpass filter realizations. With a purely metal insert Q factors of 1200 are achievable at millimetric frequencies. Printing the fins on a thin substrate produces a Fin-Line filter which is suitable for integration with other components albeit with some degradation of Q factor, typically 800 can be achieved.

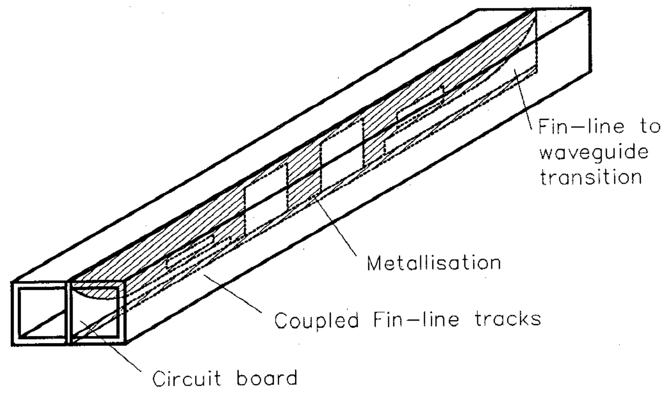

Fig. 13. Asymetric E-Plane filter.

We were particularily interested in the metal insert E-Plane filter as a building block in a 58–60-GHz transmit-receive diplexer for PCN base stations. In this case the most important filter specifications are low passband loss combined with high stopband attenuation in the adjacent band. The optimum way to achieve this is to use E-Plane filters with an asymmetric frequency response, i.e with finite frequency transmission zeros on one side of the filter passband. This can be achieved with the addition of shunt waveguide stubs to the basic E- Plane filter or with the introduction of cross couplings between nonadjacent resonators within the filter structure. Neither of these realizations were considered desirable as they both result in a more complex physical structure than for the basic E-Plane filter. Our solution was to use coupled finline transmission lines to produce extracted pole resonators yielding finite frequency transmission zeros and with inductively coupled waveguide sections to realize transmission zeros at infinity [10]. The resultant filter structure is shown in Fig. 13. This was designed to have an asymmetric Generalized Chebychev transfer characteristic . The realization is simple with the entire filter being realized on a single metal insert within a uniform waveguide. The measured insertion loss of a prototype 4th degree filter with two finite frequency transmission zeros on the high side of the passband is shown in Fig. 14. The measured results show the improved performance on the high side of the passband. The filter response was not perfectly tuned. This was due to the need to model the circuit accurately. This is presently being pursued using a transverse mode matching technique.

It is thus shown that a simple E-Plane filter structure can provide a complex filter transfer characteristic with poles of infinite attenuation located at finite frequencies on either side of the passband. This will be particularily useful for the realization of transmit/receive diplexers at the front end of millimeter wave radio transceivers. In addition there is no reason why the ceramic filter and E-Plane filter technologies may not be combined. That would yield a miniature surface mount waveguide filter with a higher Q factor than a TEM mode ceramic filter. As an example a 5.2 GHz E-plane filter

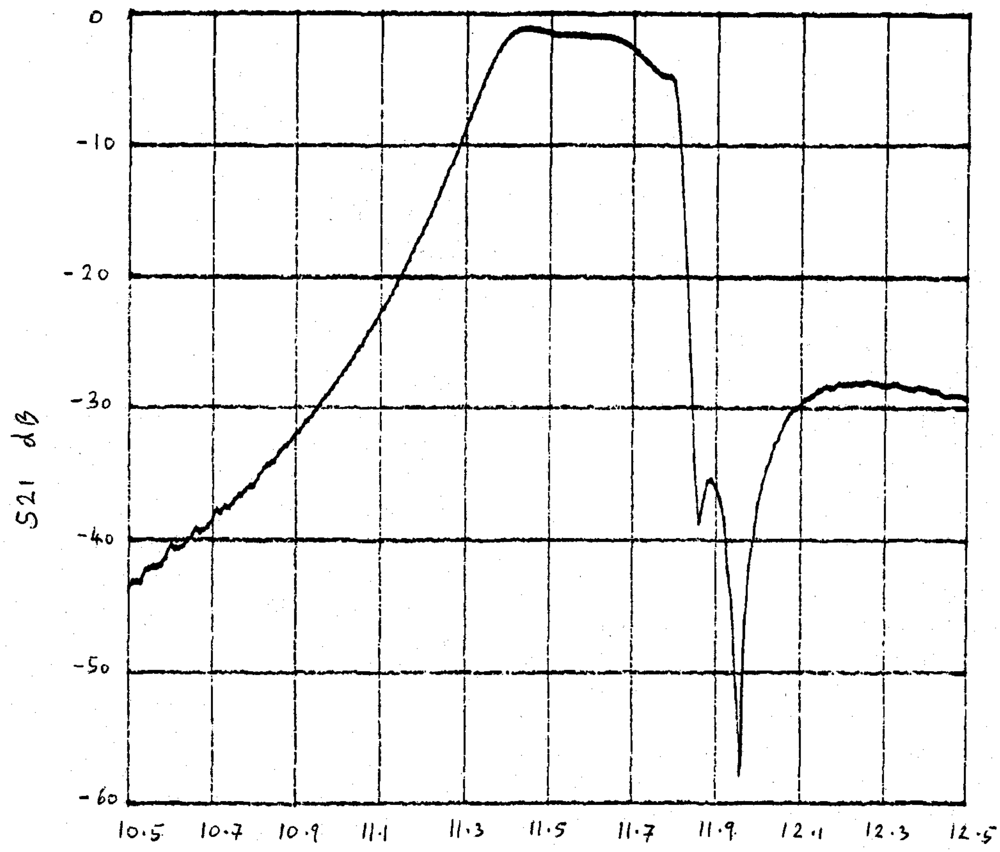

Fig. 14. Measured performance of asymetric E-Plane filter.

realized with a relative permittivity of 38 would have a broadwall dimension of 6mm and an unloaded Q factor of approximately 600. This is an exciting possibility which is presently being pursued.

VI. Conclusion

Various filter technologies suitable for use in communications systems have been investigated. Active microwave filters have been constucted and exhibit infinite Q-Factor. This could be a useful technique for realising MMIC bandpass filters with low insertion loss. However the inherant nonlinear behavior of active devices will cause intermodulation products to be produced when multiple signals are filtered. For this reason the authors believe that miniature passive filters are ultimately a better solution for products such as wireless LAN's. To this end high dielectric ceramic materials used in a miniature stripline configuration have been investigated and encouraging results for a 2 pole end coupled filter have been demonstrated. The prototype filter exhibited an unloaded resonator Q factor of greater than 300. With this technique miniature surface mount filters can be realized with a variety of transfer characteristics. E-Plane filter technology for use in millimeter wave radio systems has been investigated and a new type of E-Plane filter with an asymmetric frequency response has been demonstrated. This will be of particular use in diplexer applications. It is further proposed to combine the E-Plane and ceramic technologies to produce miniature high Q waveguide filters.

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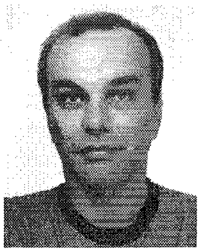

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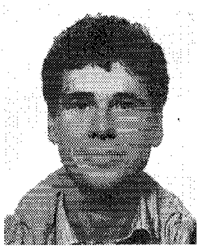

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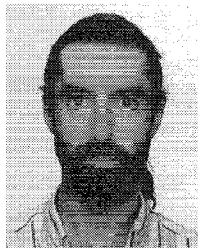

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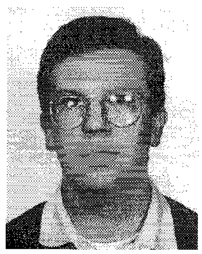

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