Asymmetric Bandpass Filter Using a Ceramic Structure

P. L. Field, I. C. Hunter, *Member IEEE*, J. G. Gardiner

Abstract—**A ceramic bandpass filter with increased selectivity on the lower side of the passband is presented. One application is the receive filter of a mobile radio transceiver operating in the 1.8 GHz bands. Finite transmission zeros have been introduced by using a network with additional cross couplings. The circuit was constructed with four coaxial ceramic resonators capacitively coupled using a microstrip PCB. For a 75 MHz passband filter, a stopband attenuation of 37 dB was attained at 20 MHz offset from the passband. A midband insertion loss of 1.2 dB and bandedge loss of 3 dB were achieved.**

I. INTRODUCTION

PRESSURES to maximize utilization of spectrum in the 1.7–2.3-GHz bands have resulted in closer spacing of paired bands assigned to mobile and personal communications services. This in turn imposes greater demands on duplexer filter performance that results in high in-band insertion loss if symmetrical designs are used to achieve adequate out-of-band rejection. An alternative is to improve the skirt selectivity on one side of the bandpass receive filter used in the front end of a mobile radio transceiver. By introducing asymmetry into the frequency response the number of resonant sections is reduced, and hence the midband insertion loss lowered whilst still maintaining adequate transmit signal rejection.

A previous letter by the authors [1] demonstrated a microstrip circuit that produced an asymmetric frequency response by placing transmission zeros at arbitrary stopband positions. However, due to the poor unloaded resonator quality factor (Q) of the structure the insertion loss attainable was greater than that required by a practical system. However, by realizing the same design of filter network in a ceramic dielectric structure a compact low-loss design can be achieved. The circuit described gives greater flexibility in locating transmission zeros than in work applied to ceramic structures described by Nishikawa *et al.* [2] and Komazaki *et al.* [3]. Ceramic filters are generally favored in cellular radio transceivers rather than lumped element or helical resonator alternatives as they are compact, offer high Q, and are suitable for integration.

Coupled line filters can be manufactured using a solid ceramic block with a metallized exterior. Parallel inductively coupled transmission lines are realized as a series of metallized holes that are grounded at one end of the block (combline) or alternate ends (interdigital). Shunt capacitors required for resonance are placed at the opposite end. Inductively coupled circuits can also be fabricated using discrete resonators with a small air gap between coupled faces. Capacitively coupled filters can be formed from coaxial resonators grounded at one end with series couplings and shunt capacitors as microstrip elements on a PCB. This approach was adopted as standard coaxial ceramic resonators could be used by selecting appropriate capacitor values. CAD models are available to determine microstrip dimensions.

II. EQUIVALENT CIRCUIT

The filter was designed for a receive passband of 1805–1880 MHz with a transmit bandedge at 1785 MHz. Using the procedure detailed by Cameron [4], [5] the asymmetric cross-coupled low-pass prototype was synthesized. In general, couplings will be of mixed sign. However, by using a transfer function with two transmission zeros at negative frequencies on the $j\omega$-axis, a network with positive inverter values can be formed, which can be realized using capacitive couplings. A fourth order 0.1-dB insertion loss ripple function with transmission zeros at $-j1.5$ and $-j2.1$ was selected for the design, providing >40-dB stopband attenuation.

The cross-coupled low-pass prototype consisting of shunt capacitors and frequency invariant susceptances can be transformed to a bandpass network of LC shunt resonators (L_t, C_t) with susceptances, B_{lc}, and susceptance slope parameters, b_{lc}, [6]:

$$b_{lc} = \frac{\omega_r}{2}\frac{dB_{lc}}{d\omega}\Big|_{\omega=\omega_r} = \omega_r C_t, \qquad \text{where } \omega_r = \frac{1}{\sqrt{L_t C_t}}. \tag{1}$$

The equivalent circuit of the ceramic filter is shown in Fig. 1. Coaxial resonators are represented as short circuit transmission lines, which have an inductive input impedance when their electrical length is in the range $0 < \theta < \pi/2$ radians. The input inductance is L_{st}, at angular frequency, ω, and is related to the characteristic impedance of the resonator, Z_{ok}, by:

$$L_{st} = \frac{Z_{ok}\tan\theta}{\omega}. \tag{2}$$

If the same coupling values as a lumped element LC shunt resonator circuit [1] are used, the bandwidth of the filter will be reduced, as the stub inductance increases with frequency. For this reason, the required coupling values are determined from the susceptance slope parameter (b_{st}) of the stub resonator:

$$b_{st} = \frac{\omega_r C_{st}}{2} + \frac{\theta_r}{2\,Z_0 \sin^2\theta_r}, \tag{2}$$

Manuscript received April 20, 1992. This work was supported as part by of a programme funded by the UK Department of Trade and Industry and the Science and Engineering Research Council under the LINK Personal Communications Programme.

The authors are with the Telecommunications Research Group, Department of Electrical Engineering, University of Bradford, Bradford, W. Yorks, BD7 1DP, UK.

IEEE Log Number 9202598.

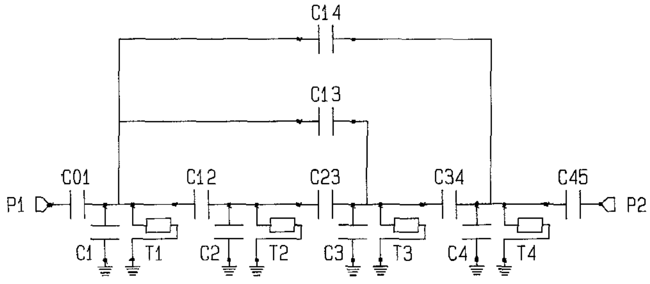

Fig. 1. Equivalent circuit.

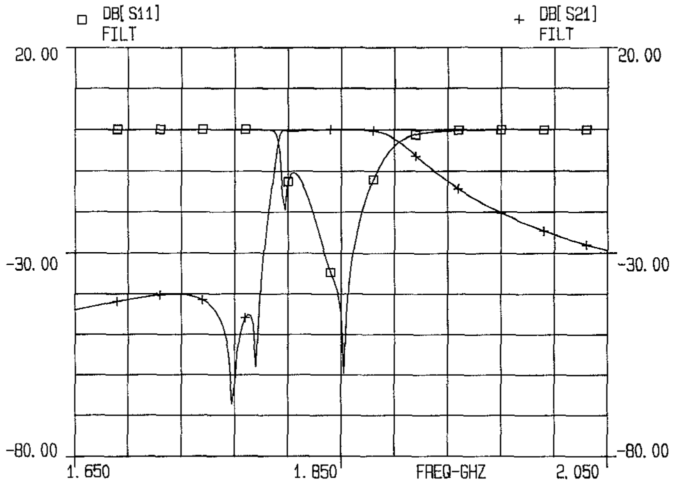

Fig. 2. Theoretical response (lossless elements).

TABLE I
ELEMENT VALUES FOR EQUIVALENT CIRCUIT

Series Capacitors/pF	Shunt Capacitors/pF	Stub Resonators T1→T4
C01 = 1.0121	C1 = 2.605	$Z = 8.5\ \Omega$
C12 = 0.2219	C2 = 3.808	$\theta = 69.09°$
C23 = 0.1409	C3 = 3.240	$f = 1842.5$ MHz
C34 = 0.3057	C4 = 2.732	
C45 = 1.0121		
C13 = 0.2116		
C14 = 0.06334		

where θ_r is the electrical length of the stub (radians) at the resonant frequency ω_r. From the selected values of θ_r and Z_{ok} the shunt capacitance, C_{st}, required for resonance can be determined from the value of L_{st} given by (1) such that $\omega_r{}^2 = 1/(C_{st}L_{st})$. The circuit can then be transformed by scaling the low-pass inverters ($J_{lp,j,k}$) to the desired bandpass values ($J_{j,k}$):

$$J_{01} = \sqrt{\frac{b_{st_1}}{b_{lc_1} R_s}}; \qquad J_{j,k|j,k=1\,\mathrm{to}\,n} = J_{lp_{j,k}} \sqrt{\frac{b_{lc_j} b_{lc_{j+1}}}{b_{st_j} b_{st_{j+1}}}};$$

$$J_{n,n+1} = \sqrt{\frac{b_{st_n}}{b_{lc_n} R_s}}, \qquad (3)$$

where n is the order of the network, and R_s is the terminating resistance. Positive inverters can be realized as $\prod$-capacitor networks, shunt capacitors can be absorbed into adjacent resonators. The values of series coupling capacitors in Fig. 1 are given by $C_{j,k} = J_{j,k}/\omega$, and the shunt capacitances by $C_n = C_{st} - \sum C_{j,k}$, as detailed in [6]. The simulated response of the equivalent circuit with ideal lossless elements is shown in Fig. 2.

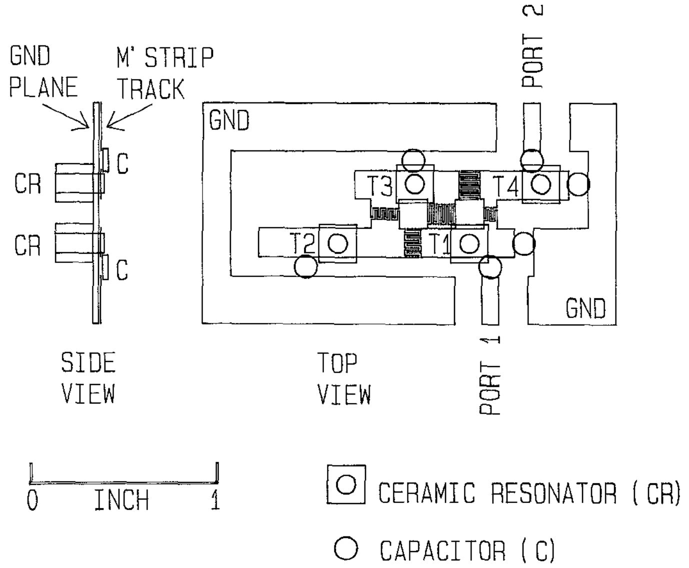

Fig. 3. Circuit layout.

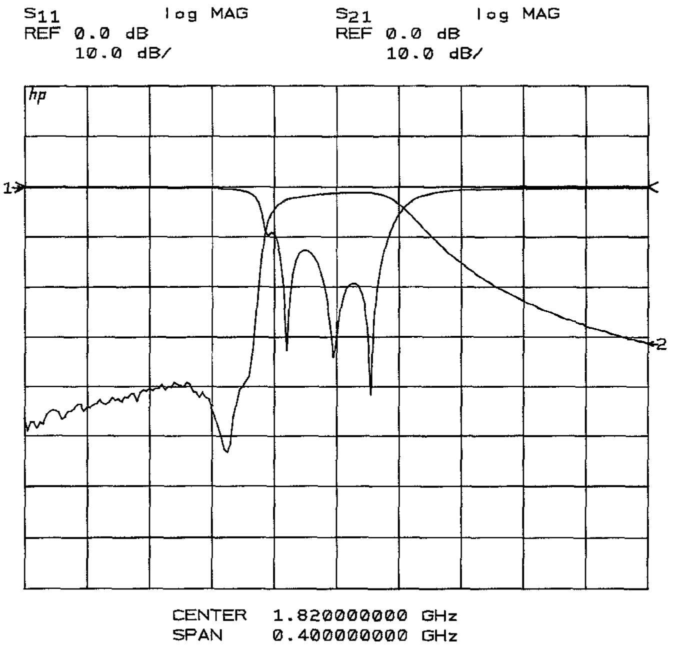

Fig. 4. Measured frequency response.

III. CERAMIC REALIZATION

The filter was designed using standard cubic cross-section quarter-wavelength coaxial ceramic resonators with resonant frequency of 2400 MHz, relative permittivity, $\epsilon_r = 38$, and outer diameter of 6 mm. From the characteristic impedance of the resonator, the input inductance can be calculated using (1),

(f_c =1842.5 MHz, Z_{ok} = 8.5 Ω, L_{st} =1.92 nH). The element values for the equivalent circuit are given in Table I. As the microstrip shunt capacitors are more lossy than the ceramic components, an increase in the length of the transmission line would improve the unloaded Q-factor of the resonator, but makes the design more sensitive to tolerances in resonator length.

The PCB was constructed using a Duriod substrate with ϵ_r = 2.33, and is shown in Fig. 3. Shunt capacitors (≈2.6–3.8 pF) were realized as electrically short lengths of low-impedance microstrip line in parallel with lumped element trimmer capacitors. Series couplings between resonators were realized as microstrip interdigital capacitors (≈0.03–0.22 pF). Lumped element trimer capacitors were used for input and output couplings (≈1 pF). The ceramic components were mounted on the PCB ground plane with a wire connection through the board from the resonator inner conductor to the microstrip shunt capacitor.

The measured frequency response is shown in Fig. 4. The center frequency measured (1821 MHz) was slightly lower than the design value (1842.5 MHz). An insertion loss of 1.2 dB was attained at the passband center and 3 dB at the 75-MHz bandwidth point. A stopband attenuation of 37 dB was achieved 20 MHz from the passband edge.

IV. CONCLUSION

An asymmetric bandpass filter has been realized using a low-loss ceramic structure. Selectivity has been improved on the lower side of passband by using a capacitively coupled network with finite transmission zeros. The cross-coupled network could be applied to an integrated structure, which could include inductive couplings. This would reduce production costs and permit transmission zeros to be placed on the upper side of the passband.

REFERENCES

[1] P. L. Field, I. C. Hunter, J. G. Gardiner, "Asymmetric bandpass filter using a novel microstrip structure," *IEEE Microwave Guided Wave Lett.*, vol. 2, pp. 247–249, June 1992.
[2] T. Nishikawa and K. Washida, "Miniaturised ceramic-based components for cellular telephones," *Microwave Eng. Europe*, pp. 23–28, Feb. 1992.
[3] T. Komazaki and K. Gunji, "Attenuation pole type dielectric filter for duplexer," *IEEE Vehicular Technol. Conf.*, Orlando, FL, May 1990, pp. 59–64.
[4] R. J. Cameron, "Fast generation of Chebyshev filter prototypes with asymmetrically- prescribed transmission zeros," *ESA J.*, vol. 6, no. 1, pp. 83–95, 1982.
[5] ——, "General prototype network-synthesis methods for microwave filters," *ESA J.*, vol. 6., no. 2, pp. 193–206, 1982.
[6] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance Matching Networks, and Coupling Structures.* Dedham: Artech House, 1980, pp. 430–431.