Asymmetric Bandpass Filter Using a Novel Microstrip Circuit

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Abstract—**A method for increasing the skirt selectivity on one side of a bandpass filter is presented. One such application is the receive filter of a PCN mobile terminal. An asymmetric synthesis method has been used, realising finite transmission zeros as cross couplings. A novel microstrip structure has been devised to demonstrate the electrical design. Although the insertion loss measured was greater than that required for a practical system, the procedure described is applicable to compact dielectric resonator structures with potentially lower loss capability.**

I. Introduction

As THE requirements of microwave filters become more stringent, there is an increasing need to introduce asymmetry in order to increase the selectivity of one of the attenuation skirts. One such application is the front-end receive filter of a UK Personal Communication Network (PCN) transceiver unit. The DCS-1800 standard to be adopted at 1.8 GHz is based upon the 900-MHz GSM Pan-European digital cellular system. The receive filter (passband 1805–1880 MHz) must provide >50-dB isolation from the transmit signal (1710–1785 MHz) for full duplex, and >30 dB for time division duplex operation. Whilst the intended air-interface standard does not require full duplex operation by filtering, service enhancements may be achievable by providing this feature. This implies a greater selectivity requirement than the previous cellular systems. There are restrictions on insertion loss, cost, size and weight. Hence, it is desirable to realize the mobile RF filter using an asymmetric frequency response to minimize the number of resonant sections required.

II. Synthesis Of Asymmetric Filter

A filter synthesis software package has been developed to produce a low-pass prototype with an asymmetric frequency response, based upon previous work applied to microwave waveguide filters described by Cameron [1], [2]. By placing a finite transmission zero on the s-plane imaginary axis a notch can be obtained at a predetermined frequency in the stop-band of the filter. A further package has been developed to transform to a bandpass lumped element circuit. The use of a synthesis method, rather than direct formulae [3], [4], permits greater flexibility in the placing of zero positions.

The transfer function $S_{12}(\omega)$ of a lossless reciprocal 2-port Nth degree Chebyshev network can be defined by (1), where ϵ, the equiripple factor, is related to the prescribed insertion loss ripple, α:

$$|S_{12}(\omega)|^2 = \frac{1}{1+\varepsilon^2 F_N(\omega)},$$

where

$$\varepsilon = \sqrt{10^{\frac{\alpha}{10}} - 1}. \tag{1}$$

The characteristic function $F_N(\omega)$ for the general Chebyshev function is given by:

$$F_N(\omega) = \cosh\left[\sum_{n-1}^{N} \cosh^{-1}(x_n)\right],$$

where

$$x_n = \frac{\left(\omega - \frac{1}{\omega_n}\right)}{\left(1 - \frac{\omega}{\omega_n}\right)}, \tag{2}$$

ω_n being the position of the nth transmission zero in the complex plane where $s = j\omega$. For a symmetric low-pass filtering function all the transmission zeros are at infinity. $F_N(\omega)$ can be expressed as a quotient of two polynomials given by:

$$F_N(\omega) = \frac{P_N(\omega)}{H_N(\omega)},$$

where

$$H_N = \prod_{n=1}^{N} (s - \omega_n). \tag{3}$$

The numerator, $P_N(\omega)$ can be generated using the recursive technique detailed in [1]. The denominator, $H_N(\omega)$, being formed from the transmission zero positions.

A transfer function with an even number of poles, N, and $N-2$ finite-transmission zeros placed arbitrarily on the imaginary axis in the s-plane can be realized using a low-pass prototype with diagonal and vertical cross-couplings. The network can be synthesised extracting each element by pre-multiplying by its inverse, as detailed in [2].

The network can then be transformed to a narrow-band bandpass response. In the general case, the coupling values will be of mixed sign. Positive admittance inverters can be approximated by $\prod$ capacitor networks, and negative inverters by $\prod$ inductor circuits. To realize a practical layout the sign of some of the couplings can be reversed, scaling each of

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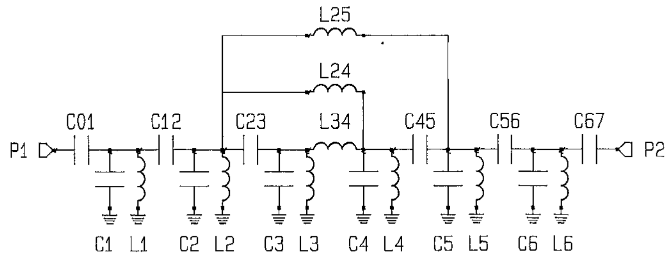

Fig. 1. Bandpass lumped element filter.

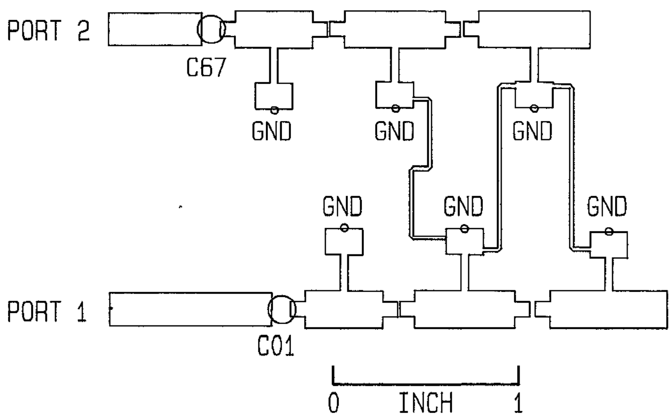

Fig. 2. Asymmetric microstrip filter.

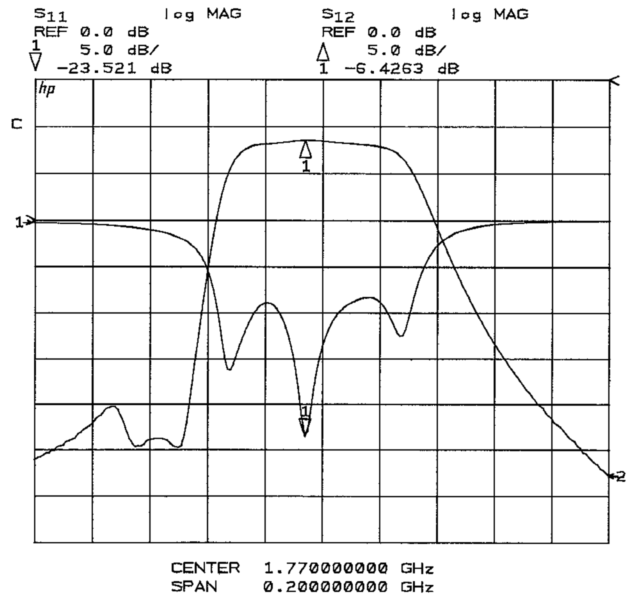

Fig. 3. Measured frequency response.

the inverters adjoining a node by -1. Shunt resonant elements can be scaled to practical values by introducing impedance transformers into the inverters.

III. Microstrip Filter

To test the operation of the synthesis method a novel form of microstrip filter employing semi-lumped elements has been devised. Although the quality factor (Q) attainable is much lower than that required for a practical duplexer the electrical design could be applied to dielectric resonator structures with potentially low-loss capability.

A sixth-order 0.1-dB insertion loss ripple function with two transmission zeros close to the transmit band-edge (low-pass values $-j1.51, -j1.81$) was found to provide >60-dB stop-band attenuation. After transforming from low-pass the bandpass circuit shown in Fig. 1 was obtained. Shunt inductor values of approximately 4 nH and inductive cross couplings have been used to yield a practical layout.

The microstrip filter, shown in Fig. 2, was constructed using an unshielded Duriod substrate relative permittivity $\epsilon_r = 2.55$. A relatively thick substrate (0.060 inch) was used to maximize the attainable Q. Resonators have been realized using electrically short lengths of low-impedance line to form shunt capacitors, with a high-impedance short circuited stub joined at a *T*-junction to form shunt inductors. Approximate dimensions were calculated using formulae given by Edwards [5]. More exact values were determined by optimizing dimensions using an RF CAD system, taking discontinuities into account, using the resonant frequency (ω_0) and susceptance slope parameter $(b_n = 1/(\omega_0 L_n))$ [6] calculated from the lumped element model. Capacitive couplings ($\approx 0.03 - 0.06$ pF) were realized using gaps between the low-impedance lines. The dimensions for the series capacitors were obtained by converting the microstrip gap CAD model to its equivalent $\prod$-circuit from the simulated *Y*-parameter measurements. Lumped element trimmer capacitors were used to realize the relatively high values (≈ 0.3 pF) for the input and output capacitances. High-series inductance values ($\approx 250 - 1100$ nH) were required for negative couplings. The series inductance (L_{se}) of a quarter wavelength a high-impedance (Z_{oh}) line can be calculated from [5] $(Z_{oh} = 133\Omega, L_{se} = 11.5$ nH). To decrease the coupling the line can be tapped close to the ground point on the distributed shunt inductor. The shunt inductor acts as a potential divider increasing the effective series inductance as seen at the input of the resonator. The position of the inductive couplings were optimized using a $\prod$-section model consisting of a tapped high-impedance line and adjoining shunt resonators. To minimize the effect of additional coupling between the parallel shunt inductor lines (which was not included in the simulation) the distance between resonators was maximized by using relatively small inductor values (≈ 4 nH) and large capacitor values ($\approx 1.4 - 1.9$ pF). As the high-impedance inductive lines are more lossy than the capacitive sections, this will maximize the attainable resonator Q. The complete filter was simulated using the dimensions determined for each section. The position of the tap points were adjusted to optimize the transmission zero positions.

IV. Results

The measured performance of the filter is shown in Fig. 3. A center frequency of 1770 MHz was measured that is slightly lower than the design value of 1842.5 MHz. This may be accounted for by additional stray inductance at the ground connection points and inaccuracies of the etching process. A mid-band insertion loss of 6.4 dB, and 3-dB bandwidth of 62 MHz were measured. A stopband attenuation of 40 dB was achieved at the transmission zero positions of 1705 MHz and 1720 MHz.

V. CONCLUSION

An asymmetric synthesis technique has been applied to bandpass filters to improve the skirt selectivity on one side of the passband. This approach permits greater flexibility in placing transmission zeros. The use of a novel microstrip structure has demonstrated the validity of introducing cross-couplings to realize stopband transmission zeros. Due to the low-Q of the microstrip structure the mid-band insertion loss (6.4 dB) and the attenuation achieved at the stopband notch (40 dB) were not adequate for the mobile radio terminal. However, the design approach is currently being applied to a dielectric resonator structure to realize a compact low-loss filter.

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