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A HIGH TEMPERATURE SUPERCONDUCTING QUASI-ELLIPTIC NOTCH FILTER FOR RADIOASTRONOMY

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ABSTRACT: *A novel miniaturized high temperature superconductor (HTS) four-pole filter for the Radio Frequency Interference Mitigation of the 900 MHz cellular band in radiotelescopes is presented. This circuit incorporates a new type of capacitive cross-coupling between nonadjacent resonators to achieve a quasi-elliptic response. Design procedure along with simulation and experimental results are shown.* © 2009 Wiley Periodicals, Inc. Microwave Opt Technol Lett 52: 88–90, 2010; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.24863

Key words: *radio frequency interference; radioastronomy; filters; superconductors*

1. INTRODUCTION

Radio Frequency Interference Mitigation (RFIM) is becoming an ever-increasing challenge for modern radioastronomy observations as the frequency spectrum becomes more crowded [1]. One well-known option to blank unwanted frequencies is to insert very high Q HTS filter before the low noise amplifier (LNA) of the radio telescope front end [2–6]. In Ref. [2], several HTS bandpass filters are discussed with spiral resonators centered at the UHF radioastronomical band for pulsar observations. In Refs. [3–5], very wide bandpass filters operating from 1330 to 1730 MHz are presented, and in Ref. [6], a Chebyshev notch filter is proposed using zig-zag resonators to mitigate the 1389–1399 MHz digital television band. However, as it is well known, Chebyshev configurations can be improved if transmission zeroes are inserted at the edges of the reject band. In Ref. [7], a quasi-elliptic filter is proposed, for base stations in which zeroes are inserted by adjusting the coupling coefficients of each individual resonator to the main transmission line using an iterative method.

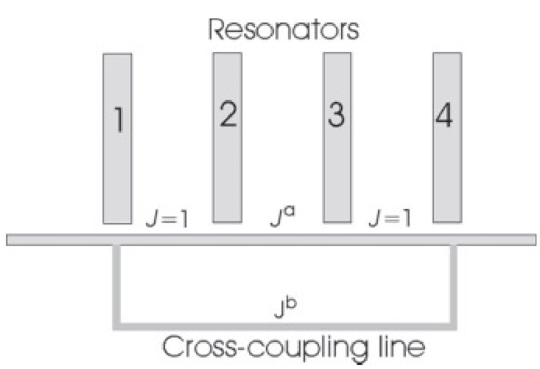

Figure 1 Schematic of quasi-elliptic notch filter

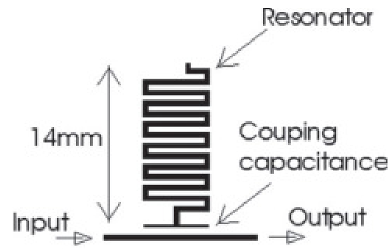

Figure 2 Single meandered resonator coupled to main transmission line

In this paper, a four-pole notch filter with quasi-elliptic response is presented for RFIM centered at 859 MHz with 8.1% fractional bandwidth (FBW) for the mitigation of the 900 MHz cellular band. The filter is designed with miniaturized half-wavelength resonators coupled to the main transmission line. A novel capacitive cross-coupling between the first and last resonators is added to introduce two transmission zeroes at the edges of the rejected band. The whole structure is meandered to fit on a 2-inch diameter sapphire substrate. Design procedure, simulated, and experimental results are presented.

2. DESIGN PROCEDURE

The bandstop design was started from the conventional quasi-elliptic low-pass prototype synthesis and transformation technique [8]. Figure 1 shows the schematic of the proposed notch filter. As it is explained in Ref. [8], a four-pole low-pass quasi-elliptic configuration can be achieved by interconnecting contiguous resonators 1–2 and 3–4 with admittance inverters of value $J = 1$. Resonators 2–3 are coupled using an admittance inverter $J_a \neq 1$ and nonadjacent resonators 1–4 are cross-coupled with $J_b \neq 1$. Moreover, J_a and J_b must have opposite signs. In the quasi-elliptic filter, the resonant frequency of each of the resonators is identical.

The low-pass filter prototype values are $g_1 = 0.9583$, $g_2 = 1.41$, $J_a = 1.1$, and $J_b = -0.1969$ for $\Omega_a = 1.85$ [8], where Ω_a represents the normalized frequency location of the transmission zero. A low pass to band reject transformation is realized, in which the $J = 1$ inverters are implemented with 50 Ω, 90° transmission lines. J_a and J_b have opposite signs hence they can be implemented with transmission lines providing a 180° phase shift (90° and 270°). The denormalized impedance value for $J_a = 1.1$ is 45 Ω and for $J_b = -0.169$ it is 253 Ω.

The normalized reactance slope parameter can be defined by the low-pass element values [8] (Eq. 1).

$$\frac{x_i}{Z_0} = \frac{1}{g_i \Delta \mathrm{FBW}}, \tag{1}$$

where Z_o is the characteristic impedance of the main transmission line, g_i corresponds to the quasi-elliptic low-pass element values, and ΔFBW is the 3 dB FBW of the filter.

The relation between x_i and the frequency response of the bandstop filter can be realized with Eq. (2) by varying the coupling-capacitance between a single resonator and the main line with the aid of a full-wave simulator (see Fig. 2) [9].

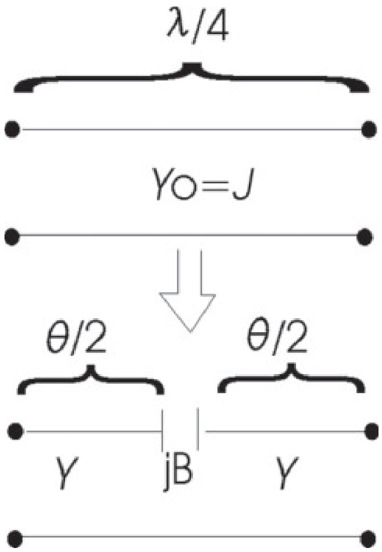

Figure 3 Admittance inverter transformation

$$\frac{x_i}{Z_0} = \frac{f_0}{2\Delta \mathrm{FBW}}. \quad (2)$$

The half-wavelength resonators are meandered to give 80% miniaturization, as shown in Figure 2.

The cross-coupling line between the first and last resonators of 253 Ω is not practical for fabrication as its width dimensions are below the achievable limit; therefore, the transformation shown in Figure 3 was implemented in which two transmission lines of length $\theta/2$ are connected in series with a capacitor C. The value of the capacitance required, and the length of the crossing lines are calculated using Eqs. (3)–(5).

$$Y = \frac{J}{\tan\left|\frac{\theta}{2}\right|}, \quad (3)$$

$$B = \frac{J}{1 - \left(\frac{J}{Y}\right)^2}, \quad (4)$$

$$C = \frac{\frac{B}{Z_0}}{2\pi f_0}, \quad (5)$$

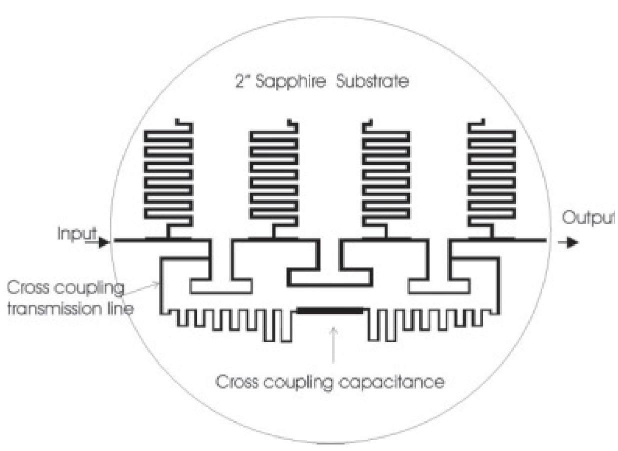

Figure 4 Final layout of the quasi-elliptic filter on 2″ sapphire substrate

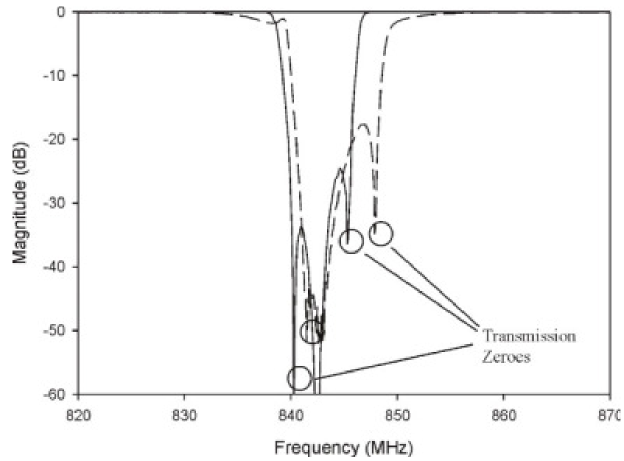

Figure 5 Simulated and measured S21 results of quasi-elliptic notch filter showing the transmission zeroes: ——— simulated, - - - - - measured

where $J = J_b = -0.1969$ is the admittance inverter line impedance value of the 253 Ω and θ is the electrical length of the crossing line. Here, θ is chosen such that the admittance Y of the new line corresponds to 50 Ω, Z_0 is 50 Ω, f_0 is the center frequency of the filter, and C is the value of the capacitance. The final values are $C = 0.08$ pF with a 50 Ω transmission line of $\theta = 365°$.

Figure 4 shows the final circuit layout on a 2-inch diameter sapphire substrate. The cross-coupling capacitor is implemented as two closely separated parallel lines. All transmission lines are meandered for miniaturization. The substrate is coated with YBCO on both sides and gold contact paths are used to connect to the K-type housing connectors.

3. SIMULATED AND EXPERIMENTAL RESULTS

The S21 simulated and measured responses of the four-pole notch filter are shown in Figure 5. The experimental bandwidth was measured to be 9%, about 1% wider than the simulated one. Moreover, there was a slight frequency shift of 15 MHz to a higher band. These discrepancies are thought to be due to substrate permittivity variation, patterning tolerances, and nonuniformities in the substrate thickness. From the Figure 5, the

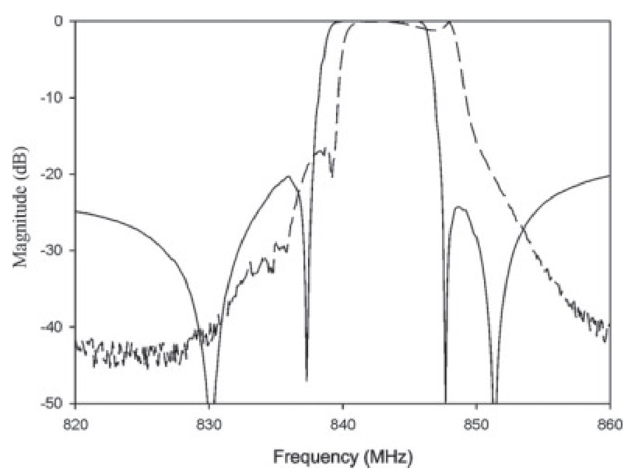

Figure 6 Simulated and measured S11 results of quasi-elliptic notch filter: ——— simulated, - - - - - - measured

transmission zeros at the stopband are clearly seen, for the simulation response they are at 840 and 845 MHz, whereas for the experimental they are at 843 and 848 MHz. The simulated rejection is better than −25 dB and the experimental one better than −20 dB.

For the S11 response, the return losses are below −17 dB throughout the passband and for the experimental better than −20 dB (Fig. 6).

4. CONCLUSIONS

A four-pole HTS microstrip notch filter to attenuate the 900 MHz cellular band has been presented for RFIM in radio telescopes. All the connecting lines and resonators are meandered for miniaturization. A capacitive cross-coupling line is used between nonadjacent resonators to achieve a quasi-elliptic response. The measured and simulated performances show good agreement.

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A NOVEL ULTRAWIDEBAND PLANAR ANTENNA WITH DUAL BAND-NOTCHED PERFORMANCE

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ABSTRACT: *In this article, a novel and compact ultrawideband (UWB) planar antenna is presented. This antenna consists of an annular patch and a 50 Ω microstrip feed line. The 3.5/5.5 GHz dual band-notched characteristic gets achieved by etching a C-shaped slot in the patch and complimentary split ring resonator (CSRR) structure in the ground. The measured results show that the proposed antenna can achieve the voltage standing wave ratio (VSWR) requirement of less than 2.0 in very UWB frequency range from 3.05 to 23.8 GHz with dual band-rejection performance of 3.4 to 4.0 GHz and 4.92 to 6.0 GHz. Measured VSWR of the proposed antenna is in good agreement with the simulated result.* Microwave Opt Technol Lett 52: 90–92, 2010; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.24856

Key words: *UWB antenna; microstrip-fed; dual band-notched performance; CSRR; C-shaped slot*

1. INTRODUCTION

Recently, the ultrawideband (UWB) planar antennas play an increasingly important role in current UWB systems due to its attractive merits, such as small size, low cost, and ease of fabrication. Over the commercial frequency band from 3.1 to 10.6 GHz approved by Federal Communication Commission (FCC) in 2002 [1], there are some other existing narrowband systems, such as WLAN (IEEE802.11a and HIPERLAN/2) system in 5.15–5.825 GHz and WiMAX system in 3.4–3.7 GHz. To avoid possible interference between UWB system and WLAN/WiMAX systems, it's desirable to design UWB antennas with dual notched bands in both 3.4–3.7 GHz and 5–6 GHz. In the last few years, band-notched UWB planar antennas based on various techniques have been proposed [2–6]. However, most of these antennas can generate only one notched frequency band.

In this article, a novel UWB planar antenna with dual band-notched performance is proposed. The dual band-notched response in both 3.4–3.7 GHz and 5–6 GHz can be achieved by etching a C-shaped slot in the annular patch and CSRR slots in the ground, respectively. The CSRR-based medium has the property of negative effective permittivity [7] and can be used to reject unwanted frequency band. Due to the relatively high insertion loss for CSRR slots, the method reported in Ref. [8] is applied to improve the out-off-band flatness of the CSRR band-stop filtering property. Moreover, the dual band-notched frequencies can be varied by adjusting the length of C-shaped slot and the dimensions of CSRR independently. Simulated and measured results of the proposed antenna, such as the VSWR, return loss, and radiation patterns, are provided.

2. ANTENNA CONFIGURATION

Figure 1 shows the geometry of the proposed dual band-notched UWB antenna. The antenna was fabricated on a 30 × 30 mm^2 FR4 substrate with relative dielectric constant of 4.4 and thickness of 1.6 mm. The radiation component is an annular patch with a C-shaped slot which is fed by a 50 Ω microstrip line with width of 3 mm. The length of the C-shaped slot is about 0.58λ for the desired frequency of 3.5 GHz. A rectangular notch with size of 3 mm × 2.2 mm on the top of the ground is proposed to improve impedance bandwidth of the presented antenna. CSRR slots in the ground plane, together with the open stub extended from feed line, can obtain band-stop filtering property at 5.5 GHz with good out-off-band flatness. The center of CSRR has been displaced by 1.5 mm from the center of the microstrip. Figure 2 shows a photo of the fabricated antenna.

3. RESULTS AND DISCUSSION

CST Microwave Studio 2006B is employed to perform the simulation and optimization process and Agilent 8722ES Vector Network Analyzer (10 MHz–40 GHz) is used to measure the VSWR. Figure 3 presents the characteristic of the simulated and measured VSWR of the fabricated dual band-notched UWB antenna. The obtained results show a good agreement between measurement and simulation results. It is seen that the proposed