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HIGH Q PRINTED HELICAL RESONATORS AND FILTERS

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Abstract: **High Q compact printed helical resonators are described operating around 1.8 to 2GHz. These consist of a multilayer PCB incorporating a printed helical transmission line. Losses in the via holes are reduced by ensuring that the standing wave current at this point is near zero. This ensures a significant increase in Q (around 4x). Further increased energy storage per unit volume is achieved due to the 3D helical nature of the resonator. Unloaded Qs of 235 and 195 have been obtained on low loss PCBs with dielectric constants of 2.2 and 10.5 respectively. As a further demonstrator a three-section filter, designed to offer the response required by the front end filter of a modern GSM phone, has been developed.**

***Keywords* - Printed Resonators, Filters**

I INTRODUCTION

Mobile radios, telecommunications and other small systems often have tight specifications on front end filtering and oscillator resonator Q for low phase noise. The specifications are often set by the tight requirements on adjacent channel noise performance and therefore require high Q compact resonator structures that can be coupled together to perform the required filtering as well as the high Q resonator . Such resonators are also required to produce the filter sections in diplexers used to separate the transmit and receive sections within hand held cellular phones.

Spiral inductors usually provide low Q due to low flux linkage in the inner turns as well as dielectric and resistive losses. Helical resonator structures [1] offer high Q transmission line resonators, however they usually require precise mounting for correct operation. Filters made from these resonators are usually produced from a cascade of quarter-wave coaxial dielectric resonators with one end grounded and capacitive coupling between the resonators. These filters can be complex to assemble, require considerable production time and are often bulky and costly.

This paper describes a high Q compact fully printed helical resonator structure produced in triplate. This allows cascade integration into fully printed filter structures and dramatically reduces the complexity, build time and cost associated with coupled coaxial dielectric resonator filters. These fully printed filter structures also lend themselves to production within high frequency integrated circuits.

II RESONATORS

These printed resonators have an equivalent circuit as shown in Figure 1 and are a combination of a curved printed half-wave transmission line resonator similar to the straight structure described in [3] as shown in Figure 2 and an inductively coupled λ/2 helical resonator [2] as shown in figure 3.

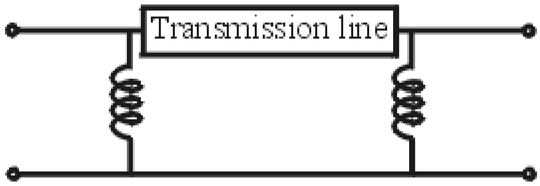

Figure 1 Equivalent circuit

Figure 2 Printed Inductively Coupled Transmission Line Resonator

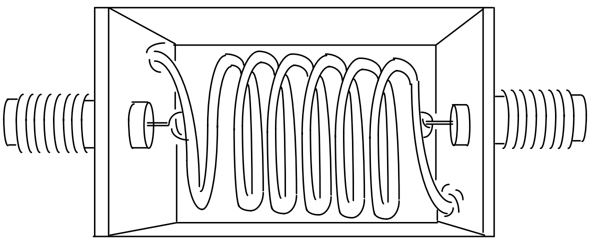

Figure 3 Shunt Inductively Coupled Helical Resonator

The new resonator is shown in Figure 4 and consists of a half wave helical transmission line resonator with shunt inductors at either end. The helix is 1½ turns made up of a ¾ turn on one side of the PCB connected using a via hole to the ¾ turn on the other side. There is a ground plane above and below the helix. Special grounding slots have been used to produce the low loss inductors.

The equivalent circuit is again as shown in Figure 2. This structure allows the resonator to be printed in a fraction of the

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space of that of a straight transmission line resonator. Losses in the via holes are reduced by ensuring that the standing wave current at this point is near zero. This ensures a significant increase in Q (around 4x). Further increased energy storage per unit volume is achieved due to the 3D helical nature of the resonator. Unloaded Qs of 235 and 195 have been obtained on low loss PCBs with dielectric constants of 2.2 and 10.5 respectively.

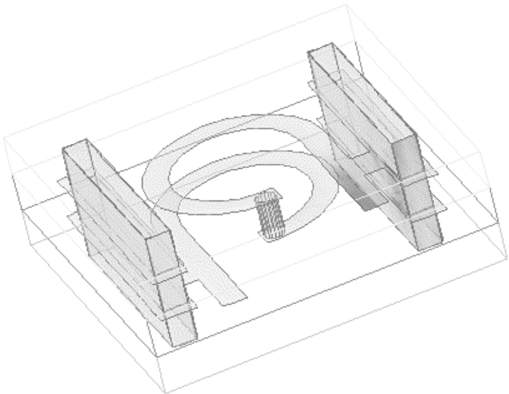

Figure 4 Shunt inductively coupled resonator

III CONSTRUCTED RESONATORS

Several 1900MHz printed helical resonators have been realised on a number of different permitivity boards, using a triplate topology and through hole plating technology. A low dielectic constant board (RT 5880) was chosen to increase the size of the structure, increase alignment tolerance and reduce dielectric losses (Figure 5). A high dielectric constant board (RT 6010) was also chosen to reduce the size of the structure at the cost of higher dielectric losses and reduced alignment tolerances.

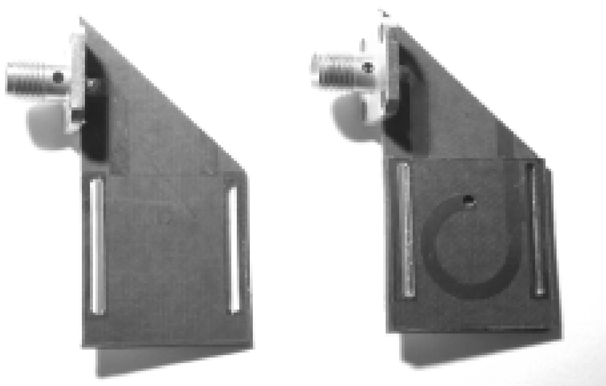

Figure 5. A photograph of a printed helical resonator

The overall resonator dimensions when printed on Rogers RT 5880 (ε_r=2.2, H=0.78mm, T=54um after through hole plating) are 18mm^2 by 2.5mm. The overall resonator dimensions when printed on Rogers RT 6010 (ε_r=10.5, H=1.27mm, T=54um after through hole plating) are 8mm^2 by 4mm.

IV MEASUREMENTS

Simulation results predicted a Q_0 of 270 and a resonance frequency of 1890MHz for the RT5880 resonator and a Q_0 of 215 and a resonance frequency of 1765MHz for the RT6010 resonator. The measured Q_0s were 235 and 195 respectively, while the fundamental resonances were 1960MHz and 1990MHz (Figures 6,7).

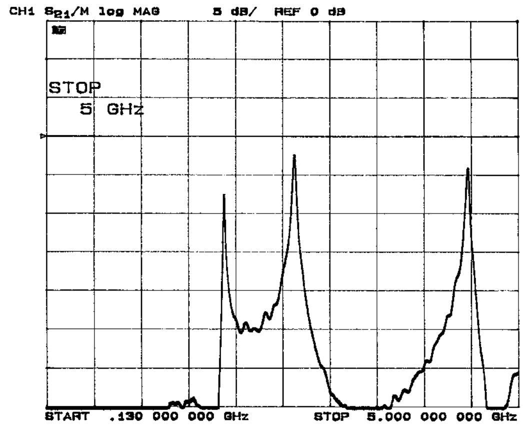

Figure 6. S_{21} frequency response for a printed helical resonator

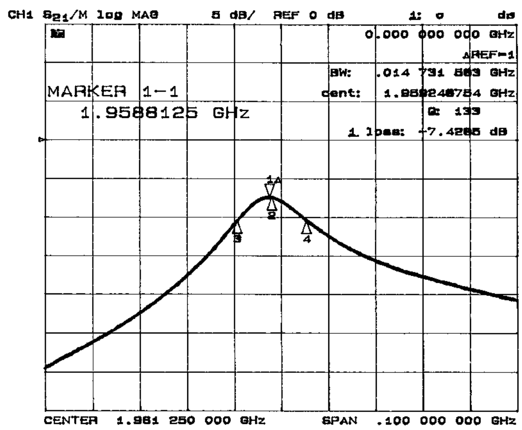

Figure 7 Measured loaded Q of the RT5880 printed helical resonator

It is foreseen that this structure could be implemented on extremely high dielectric constant substrates such as barium titanate, allowing a further size reduction to half the dimensions of the RT6010 resonator (4mm^2 by 2mm is feasible). Several of these resonators could then be appropriately coupled to produce compact fully printed filter sections for use in such systems as mobile phone transmit/receive diplexers.

V THEORY

The theory for transmission line resonators using this structure [4] is shown below where the transmission coefficient of the resonator, S_{21}, can be shown to be:

$$S_{21} = \frac{4z\Gamma}{\left(2+\Phi^2\right) - \Phi^2\Gamma^2} \tag{1}$$

where:

$$\Gamma = \exp\left[-(\alpha + j\beta)L\right] \tag{2}$$

$$\Phi = z + zX - 1 \tag{3}$$

$$z = \frac{Z_T}{Z_0} \tag{4}$$

Z_T is the resonator transmission line impedance, Z_0 is the terminating impedance, α and β are the attenuation coefficient and phase constant of the transmission line respectively. The resonant frequency can be shown to be:

$$f_0 = \frac{V_{EFF}}{2L}\left[1 + \left(\frac{1}{\pi}\right)\tan^{-1}\left(\frac{2Xz}{X^2z^2 + z^2 - 1}\right)\right] \tag{5}$$

The insertion loss at resonance is therefore:

$$\left|S_{21}(0)\right| = \frac{4z}{4z + 2\alpha L\left[(z-1)^2 + X^2z^2\right]} \tag{6}$$

If $Q_L >> \pi$ then:

$$Q_L = \frac{\pi}{4z}\left|S_{21}(0)\right|\left((z-1)^2 + X^2z^2\right) \tag{7}$$

When the phase shift of the resonator is neglected:

$$S_{21}(\Delta f) = \frac{\left|S_{21}(0)\right|}{1 + 2jQ_L(\Delta f / f_0)} \tag{8}$$

$$\left|S_{21}(0)\right| = 1 - Q_L / Q_0 \tag{9}$$

$$Q_0 = \frac{\pi}{2\alpha L} \tag{10}$$

It is interesting to note that If $Z_T = Z_0$, where Z_T is the resonator line impedance and Z_0 is the terminating impedance and α is the voltage attenuation coefficient of the line β is the phase constant of the line. For small αL (< 0.05) and $\Delta f/f_0$ << 1, the following properties can be derived for the first resonant peak (f_0) of the resonator where $\Delta f = f - f_0$ then equation (5) simplifies to:

$$f_0 = \left(\frac{V_{eff}}{2L}\right)\left(1 + \left(\frac{1}{\pi}\right)\tan^{-1}\left(\frac{2}{X}\right)\right) \tag{11}$$

Equation (7) simplifies to:

$$Q_L = \pi S_{21}(0)\left(\frac{X^2}{4}\right) \tag{12}$$

And:

$$S_{21}(0) = (1 - Q_L / Q_0) = \frac{1}{\left(1 + \left(\frac{\alpha L}{2}\right)X^2\right)} \tag{13}$$

From these equations it can be seen that the insertion loss and the loaded Q factor of the resonator are interrelated. In fact as the shunt capacitors (assumed to be lossless) are increased the insertion loss approaches infinity and Q_L increases to a limiting value of $\pi/2\alpha L$ which we have defined as Q_0. It is interesting to note that when $S_{21} = 1/2$, $Q_L = Q_0/2$.

VI FILTERS

As a further demonstrator a three section filter has been designed using these resonators. This has been arranged to offer the response required by the front-end filter of a modern GSM phone.

The filter has been designed using the design techniques described in Mathaei, Young and Jones (P.528). The design is based on a Chebishev filter with 0.2dB ripple with ω_0 = 1.2315.1010 and $\Delta\omega$ = 0.03061. It includes immitance invertors as illustrated in Figure 8. Two difficulties with the design are: correct knowledge of the impedance of the helix which is often high and the correct propagation velocity within the helix. TDR measurements on the helix used in [2] demonstrated impedances around 350Ω. The final filter and frequency response are shown in Figures 9 and 10.

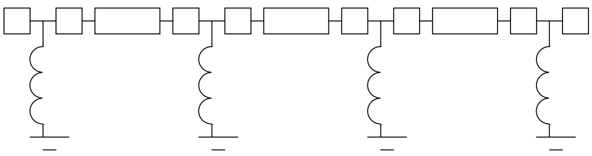

Figure 8 Prototype Filter Structure

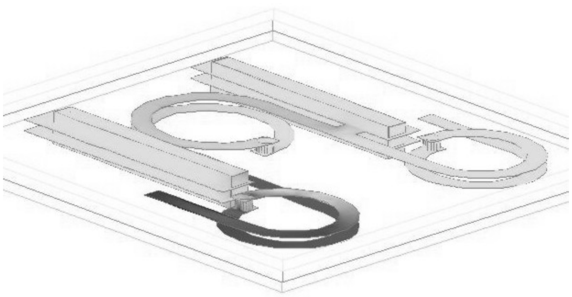

Figure 9 3D plot of filter

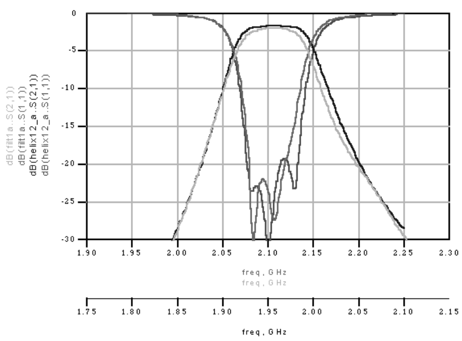

Figure 10 Frequency response of Filter

VII CONCLUSIONS

A compact fully printed three layer helical resonator and a GSM band three resonator structure has been designed. Theoretical equations have been derived to describe the operation of the resonator and the structure has been implemented on several different dielectric constant boards. Measurements have been taken that agree closely with the theory and simulation. It is foreseen that this structure lends itself to direct coupling to form planar printed GSM band filters / diplexers on a wide variety of substrates and relatively easy integration in to high frequency integrated circuits where the substrate is left or removed.

VIII ACKNOWLEDGEMENTS

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