A Dual-Passband Filter Architecture for Dual-Band Systems

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***Abstract*— This paper proposes a novel dual-passband filter based on a Chebyshev filter architecture. The two passbands can be independently tuned to desired frequencies. Also, the filter has the ability to provide matching to impedances with high real values at both frequencies which was impossible in previous design methodologies. This novel feature eliminates the need for further impedance matching through using quarter-wave transformers. The architecture is specifically attractive when matching to dual-resonating loads in dual-band systems. The measured results approve that the proposed method can achieve matching at both desired frequencies.**

Keywords- Bandpass filters; Microstrip; Transmission lines; RLC circuits;Resonance

I. INTRODUCTION

During recent years, wireless standards have been developed in previously unused frequency bands. WLAN, for example, used the 2 .4GHz band in its earlier versions then switched to 5GHz for improved speed [1]. Obviously, due to backward-compatibility issues, a system that can operate at both bands is desired. There has been significant interest in developing an RF front end that can cover different bands [2]. Fig. 1 shows the two possible choices for a dual-band receiver front end. In the first approach all system components are duplicated. This approach is neither cost-effective nor area-efficient. A smarter approach designs the elements such that they can operate at both frequencies of interest. This method, shown in Fig 1b, eliminates the need to duplicate the power-hungry RF components and also saves space by removing bulky passives.

Few design methodologies have been proposed in literature for designing the dual-passband matching networks [3, 4]. In [3] a complex design methodology for design of microstrip bandpass filters is presented that requires characterizing the technology by building a 2-dimentional design space. However, there is a possibility that the optimal point does not fall in the valid design space. Reference [4] proposed introducing three zeros to the transfer function by adding three stubs. However, this results in a 5-way microstrip junction that is difficult to model and thus cannot be optimized easily. In addition, the aforementioned approaches assume a 50Ω effective load resistance at the desired frequencies. However, in many applications such as inductively-powered biomedical implants the antenna is replaced with an inductor that is resonating in an LC structure [5]. Thus, the antenna input impedance will be a large real number. Conventionally, bulky transformers or quarter-wave transformers have been used in these cases. The proposed dual-passband network has the advantage that it can match to arbitrarily large impedances without additional circuits.

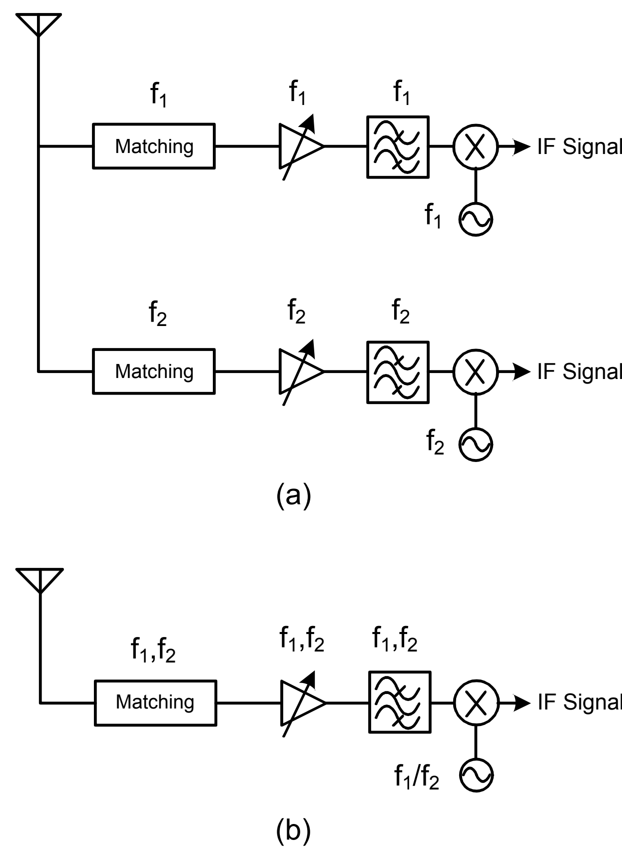

Figure 1. (a) Conventional dual-band receiver front-end. (b) Inherently dual-band receiver front-end.

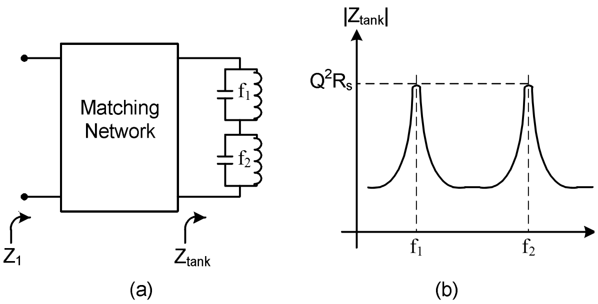

Figure 2. (a) Matching network for a dual-resonator tank. (b) Impedance of a dual-resonator tank.

In this paper, we introduced a matching network for dual-band systems shown in Fig. 1b. The approach proposes a novel method to match high real-valued impedance to standard LNA input impedance of 50Ω.

This paper is organized as follows: Section II provides a qualitative description of the approach. Section III derives the formulas for design of dual-passband filters. Section IV covers the simulation and measurement results. Finally, Section V draws the conclusion of the paper.

II. QUALITATIVE DESCRIPTION OF THE APPROACH

Assume two LC resonators are connected in series as shown in Fig. 2a. Our goal is to design a matching network that can provide 50Ω real impedance at both resonance frequencies.

Fig. 2b depicts the absolute value of the dual-resonator tank impedance [6] where Rs is the series resistance of the tank inductors and Q is the quality factor. For a typical microstrip coil with a quality factor of approximately 100 and series resistance of 0.1-1Ω, the impedance seen will be more than 1kΩ. Thus, designing a matching network becomes very challenging. Note that we expect the final circuit to be very narrowband according to the Bode-Fano criterion [7].

In order to tackle the high ratio of load over source impedance, consider the well-known Chebyshev equal-ripple filter illustrated in Fig. 3a. It is well known that for an even number of stages, N, the load impedance required to satisfy matching condition becomes larger than the source impedance [8]. We define this ratio as

$$\alpha = \frac{R_L}{R_S}. \quad (1)$$

However, note that α increases with the amount of ripple in the desired filter. Fig. 3a shows this dependency. Increased ripple can be considered as the formation of better "stop-bands".

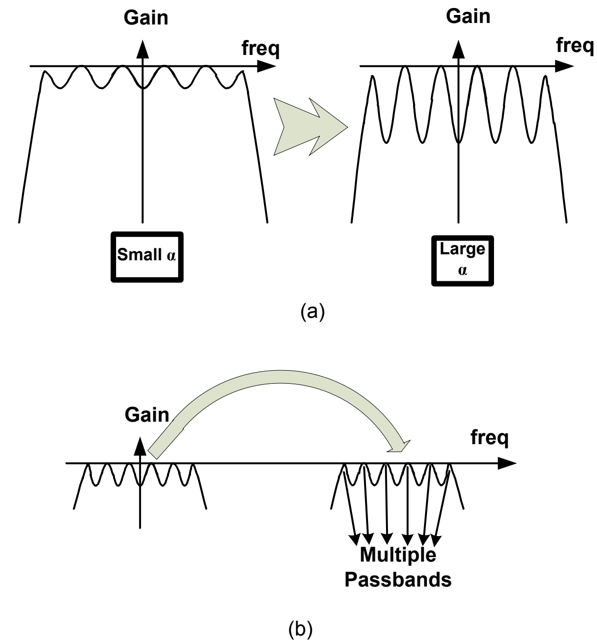

Figure 3. (a) For a Chebyshev filter, ripple increases as we increase α. (b) Transforming a high-ripple low-pass Chebyshev filter to bandpass filter enables the design of filters with multiple passbands.

To this point we have built a high-ripple low-pass filter. If we now apply a lowpass-to-bandpass transformation on the filter, a bandpass filter with multiple passbands will be formed as shown in Fig. 3b. This way, the high α challenge has been overcome while simultaneously achieving a multiple-passband response.

III. MODEL DERIVATION

Based on the described approach we can determine appropriate L and C values that provide matching for a given α, ω1, and ω2. Specifically, the input impedance and reflection coefficient of the second order Chebyshev filter can be derived as

$$\begin{aligned} Z_{in} &= j\omega L + \frac{R(1 - j\omega RC)}{1 + \omega^2 R^2 C^2} \\ \Gamma_{in} &= \frac{Z_{in} - 1}{Z_{in} + 1}. \end{aligned} \quad (2)$$

Here R is the load resistance normalized to Rs=50Ω and other impedance values are also normalized to the same value.

The power loss ratio can be written as

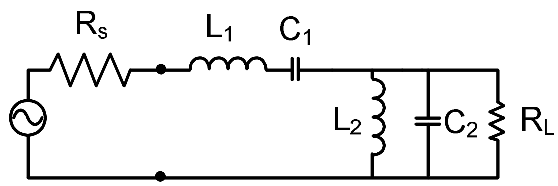

Figure 4. The final matching circuit consisting of two inductors and two capacitors.

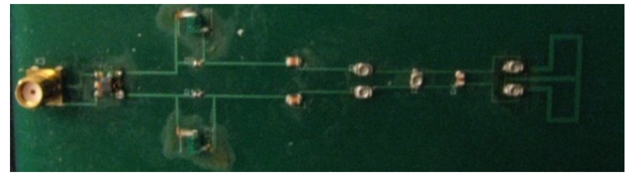

Figure 5. The dual-resonator tank and its matching network. The two loops on the right act as coils for tanks.

$$
\begin{aligned}
P_{LR} &= \frac{1}{1-\Gamma^2} = \\
& 1+\frac{1}{4R}[(1-R)^2 \\
& +(R^2C^2+L^2-2LCR^2)\omega^2+L^2C^2R^2\omega^4].
\end{aligned} \tag{3}
$$

We need to equate that with Chebyshev polynomial of 2nd order:

$$P_{LR} = 1+k^2T^2_{2(\omega)} = 1+k^2(4\omega^4-4\omega^2+1) \tag{4}$$

This results in:

$$
\begin{cases}
k^2 = \dfrac{(1-R)^2}{4R} \\
4k^2 = \dfrac{1}{4R}L^2R^2C^2 \\
4k^2 = \dfrac{1}{4R}(R^2C^2+L^2-2LCR^2).
\end{cases} \tag{5}
$$

Note that the amount of ripple required, k, is only dependent on R. After solving the first equation for k, L and C will be determined using the two other equations. L and C correspond to the values for the low-pass equivalent of the filter. As a result, we must perform a low-pass to band-pass transformation. Let us choose the transformed filter to be centered at the geometric mean of the two desired frequencies with bandwidth 2(ω2 - ω1). Therefore, the values for the final circuit elements shown in Fig. 4 are:

$$
\begin{aligned}
L_1 &= \frac{LZ_0}{\omega_0\Delta}, C_1 = \frac{\Delta}{\omega_0 L Z_0} \\
L_2 &= \frac{\Delta Z_0}{\omega_0 C}, C_2 = \frac{C}{\omega_0 \Delta Z_0}.
\end{aligned} \tag{6}
$$

where:

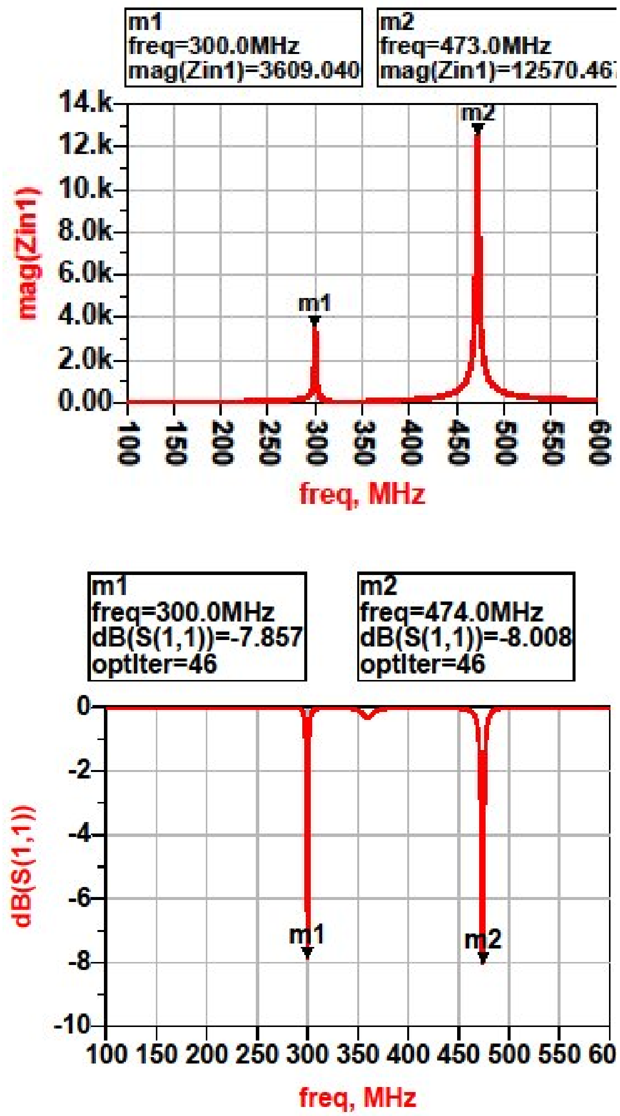

Figure 6. (a) EM-simulated impedance of the dual-resonator tank. (b) Simulated reflection coefficient of the final circuit.

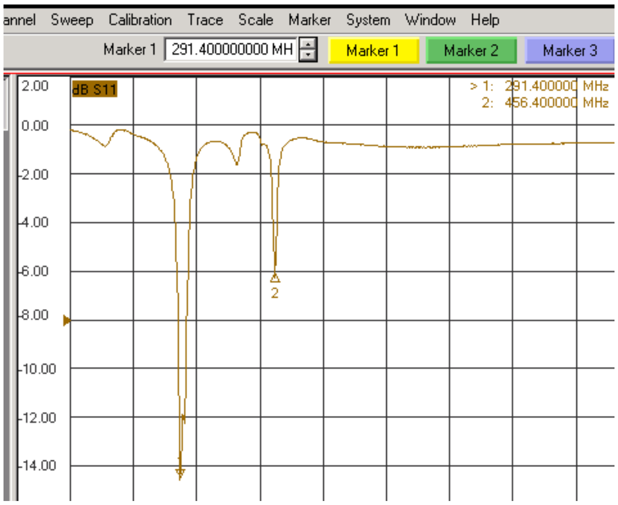

Figure 7. (a) The final matching circuit consisting of two inductors and two capacitors.

$$\Delta = \frac{\omega_2 - \omega_1}{\omega_0}$$
$$\omega_0 = \sqrt{\omega_2 \omega_1}. \tag{7}$$

IV. SIMULATION AND MEASUREMENT RESULTS

Based on the proposed method, a dual-passband filter was designed to match a dual-resonator load to 50Ω. Fig. 5 shows a dual-resonator tank and its matching network. The tank inductors are realized using PCB traces as is common in inductive power transfer schemes [9]. Fig.6a shows the simulated impedance of the tank using the Method of Moments. Thus, the load impedances at the resonance frequency is

$$RL = 4k\Omega \tag{8}$$

Also, the two center frequencies are:

$$\omega_1 = 2\pi \times 300\text{MHz}$$
$$\omega_2 = 2\pi \times 473\text{MHz}. \tag{9}$$

Using our derived formulas to derive component values and simulating the resulting circuit results in the reflection coefficient plot shown in Fig. 6b. The two matching frequencies are clearly visible in the response. The matching network circuit was made using discrete components as seen in Fig. 5. An Agilent N9020 network signal analyzer was used for the reflection coefficient measurements shown in Fig. 7.

Comparing Figs. 6b and 7 demonstrates the effectiveness of our proposed method. The slight shift in frequencies is largely due to extra parasitics introduced by board elements.

V. CONCLUSION

The paper presented a novel dual-band matching network based on a Chebyshev filter. It was shown that the proposed approach is able to match the dual-resonator tank to a known source resistance at desired frequencies. Simulation and measurement results showed that the matching network provides the required performance.

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