A Novel Compact Dual Notch Band Filter Based on Meta-material Concept

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Abstract: Using composite right-left handed (CRLH) transmission line concept, a novel miniaturized dual notch band filter (DNBF) is proposed. The suggested DNBF consists of an interdigital transmission line (ITL), split ring resonators (SRRs) and complementary split ring resonators (CSRRs). Since the resonance frequency of the SRRs and CSRRs are quite independent of each other, the dual notch bands of the proposed filter can be separately controlled and shifted by changing the dimension of the SRRs and CSRRs. In this paper, the reject bands are designed for WLAN (2.4 GHz) and WiMAX (3.5 GHz) to reject these frequency bands from the ultra-wide band communication systems. The simulation results show that the transmission response has more than 32 dB rejections near each band. To validate the design concept, the proposed NBPF has been fabricated and tested. Experimental verification is provided and good agreement has been found between simulation and measurement. To the best of our knowledge, the proposed NBPF is more compact in comparison with other reported filters.

Keywords: Metamaterial, Interdigital Transmission Line (ITL), Split Ring Resonator (SRR), Complementary Split Ring Resonator (CSRR), Dual Notch Band Filter (DNBF), Compact Size.

1. Introduction

Band reject and narrowband notch filters are commonly used in microwave communication, and radar systems to cancel undesired signals and prevent interference of waves from spurious sources. In particular, one of their important applications is in the ultra-wide band (UWB) communication systems that use the frequency band from 3.1 GHz to 10.6 GHz [1]. The high power interferers of wireless LANs (WLANs) in the 5–6 GHz spectrum, Bluetooth and IEEE 802.11b/g systems in the 2.4–2.5 GHz frequency range, cellular phones in the 0.9–1.9 GHz band, and WiMAX services in the 2.5–2.9, 3.4–3.6 and 5.2–5.9 GHz bands, can have a detrimental impact on the received UWB signal, since their intermodulation products can fall in-band and also, they can cause the compression of the receiver gain [1-8]. Achieving attenuation in the interfering signals range without degrading the gain, require the utilization of many passive components with high-quality factors, which causes cost and complexity increase. Moreover, the parasitic resistances of the utilized passive components prevent a high attenuation.

Many various planar filter configurations have been proposed in this regard. Here, we have focused on the reported dual notch-band filters (DNBF). In [9], by utilizing frequency-variable transformation to the low-pass prototype, the dual stopbands are realized. In most recent publish researches, the size reduction of DNBF has investigated by using a dual-mode loop resonator [10], or implementing two and three-section stepped-impedance resonators [11, 12]. However, the above proposed filters have not the ability to adjust the center frequency of each stopband. To solve this problem, [13] has proposed a filter structure that implement the open-loop resonators with different length. Although, this configuration leads to a larger size of the filter. The defected ground and microstrip structure have also been utilized for the realization of compact DNBFs with adjustable stopbands [14-16]. Nevertheless, the design of conveniently adjustable stopbands of DNBFs in smaller sizes is still a challenging research area [17-19].

In recent years, metamaterials have attracted rising interest among microwave researches [20-22]. These structures with simultaneously negative permittivity and permeability are named left-handed (LH) materials. They are frequency-selective structures which makes them very appropriate candidates for implementing them in the designation of compact microwave filters and diplexers.

Since these devices need resonant elements, the ideal approach for the realization of filters and diplexers based on metamaterials is the resonant-type approach [23, 24]. It has been shown in [25] that SRRs and CSRRs are useful structures for the synthesis of narrowband and wide-band-pass filters [8].

In this paper, a compact Dual Notch Band (DNB) filter using an interdigital transmission line (ITL) with split ring resonators (SRRs) and complementary split ring resonators (CSRRs) is proposed. To make the design practical, the reject bands of the filter are considered for the WLAN band and WiMAX band, namely 2.4 GHz and 3.5 GHz. Consequently, one of the applications of such notch filters may be to reject the WLAN and WiMAX frequency bands from the ultra-wide band (UWB) communication systems. The proposed DNB filter realized two transmission zeros at 2.4 and 3.5 GHz with sharp attenuations near its stopbands. This filter has a compact size and its minimum strip and gap width are equaled to 0.2 mm, which keeps its fabrication cost low.

2. Metamaterial Ring Resonators (SRR and CSRR)

Metamaterial is an artificial effectively homogeneous electromagnetic structure with negative values of ε and μ [25]. As these structures are frequency-selective, they are very good candidates for the design of compact microwave devices. As an example, metamaterial transmission lines offer a controllability on their electrical characteristics that gives us the possibility to design microwave components with superior performance, such as enhanced bandwidth components, or microwave devices with small dimensions. It has been shown that SRRs and CSRRs are capable of realizing metamaterial structures [25]. Split ring resonators (SRRs) have initiated tremendous interest among microwave researchers due to their potential applications in the synthesis of metamaterials with negative effective permeability [26, 27]. From a duality argument, Falcone et al. in 2004 introduced complementary split-ring resonators (CSRRs) as new metamaterial resonators that exhibit negative permittivity [28]. The key property of these metamaterial resonators is that they are electrically small and exhibit an effective negative permittivity or permeability in a narrow band above their resonant frequency.

The circuit model for a SRR and CSRR is shown in Fig. 1. The CSRR circuit model can be obtained by replacing the metal parts of the SRR with apertures, and the apertures with metal plates, based on the duality theorem [25]. The resonance frequency of these resonators can be calculated as:

$$f_\circ = \frac{1}{2\pi\sqrt{L_c C_c}} \tag{1}$$

Another kind of metamaterial resonators are spirals resonators. These structures provides a strong magnetic dipole at their resonance so they can be very useful in the design of negative magnetic permeability and left-handed media [25]. In this paper, we use the two-turn spiral resonator (2-SR) shown in Fig. 2(a). A quasi static analysis of this configuration leads to the equivalent circuit Fig. 2(b). This equivalent circuit shows that the resonance frequency of the 2-SR is half of the resonance frequency of the conventional SRR of the same size and shape [25]:

$$f_\circ(2-SR) = \frac{1}{2} f_\circ(SSR) \tag{2}$$

This property is a clear advantage, as it implies a smaller electrical size at resonance. The electrical size can still be reduced by increasing the number of turns.
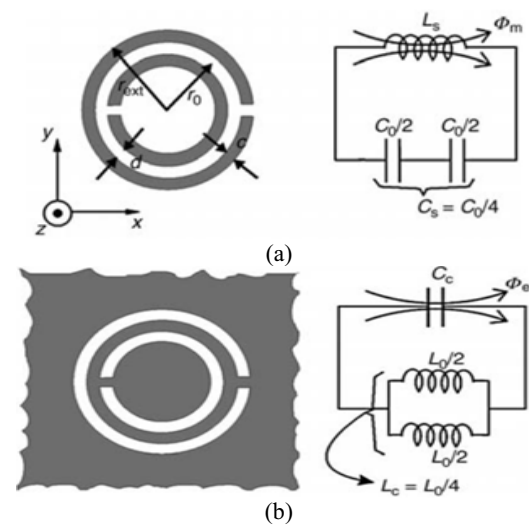

Fig. 1. Topologies of the SRR (a) and CSRR (b), and their equivalent circuit models [25].
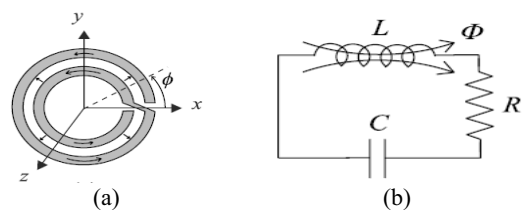

Fig. 2. (a) Topologies of the Spiral and (b) their equivalent circuit model [25].

3. Interdigital TL

An interdigital capacitor is a multi-finger periodic structure that can be utilized as a series capacitor in planar transmission lines [29]. In Fig. 3(a), an interdigital capacitor with its equivalent circuit model is depicted. As it can be seen, it is made of some gaps that meander back and forth, creating two sets of interdigital fingers. These gaps are designed very long and folded to create compact single-layer capacitors. Typically, the capacitance of these structures vary from 0.05 pF to about 0.5 pF and can be raised by increasing the number of fingers, or by adding a thin layer of high dielectric constant material between the conductors and the substrate [29]. Series capacitance per length of an interdigital capacitor, i.e. , with physical parameters that have been presented in Fig. 3(a), is equal to [30]:

$$C_{int} = \frac{\varepsilon_r + 1}{W_{int}} L_{int} \left[(N - 3) A_1 + A_2 \right] \quad (3)$$

where ε_r is the permittivity of the microstrip substrate, N is the number of structure fingers and the dimensions Wint and Lint are shown in fig. 3(a). Approximate expressions for   and   are obtained by doing curve fitting on the data

given in [31]. These expressions are as [29]:

$$A_1 = 4.409 \tanh \left[0.55 \left(\frac{h}{w} \right)^{0.45} \right].10^{-6} \, (pF/\mu m)$$

$$A_2 = 9.92 \tanh \left[0.52 \left(\frac{h}{w} \right)^{0.5} \right].10^{-6} \, (pF/\mu m) \quad (4)$$

where h is the substrate thickness and w is the width of one finger. In Fig. 3 (a), L_{int} and C_p are conventional series inductance and shunt capacitance in a microstrip TL and are considered as parasitic elements in an interdigital structure. Values of these elements can be calculated from TL theory using the length of the structure (l), as [29]:

$$L_{int} = \frac{\sqrt{\varepsilon_{re}} Z_0}{c} \ell$$

$$C_p = \frac{\sqrt{\varepsilon_{re}}}{Z_0 c} \ell \quad (5)$$

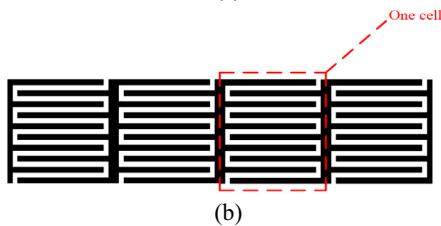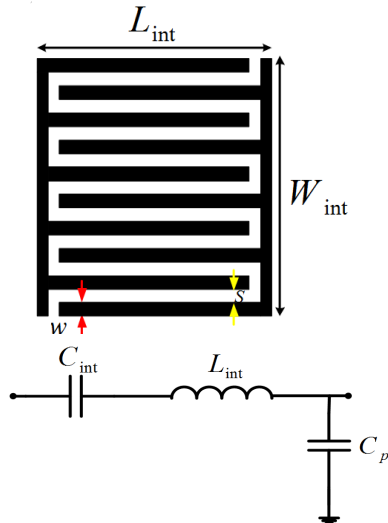where ε_{re} is the effective permittivity of a microstrip TL whose strip width is w, Z_0 is the characteristic impedance of a microstrip TL with strip width of W_{int} (where $W_{int}=(2N-1)s+Nw$), c is the velocity of light in Free space, and N is the number of fingers in interdigital capacitor. As shown in Fig. 3 (b), an interdigital TL (ITL) can be constructed by cascading some interdigital capacitors, which are unit cells of the interdigital TL [29]. If we define per-unit-length impedance (Z') and admittance (Y') as [29]:

$$Z' = j \left(\omega L' - \frac{1}{\omega C'_{int}} \right), \quad (6)$$

$$Y' = j \omega C',$$

Where

$$L_{int} = L'.\ell, \quad C_p = C'.\ell, \quad C_{int} = C'_{int} / \ell, \quad (7)$$

then, from transmission line theory, the propagation constant   and the characteristic impedance of a TL with series impedance and parallel admittance , are obtained as [29]:

Fig. 3. (a) Interdigital capacitor and its equivalent circuit mode and (b) Interdigital TL [29].

$$\gamma = \sqrt{Z'\,Y'}\,, \tag{8}$$

$$Z_c = \sqrt{\dfrac{Z'}{Y'}}\,. \tag{9}$$

Hence, for interdigital TL, the complex propagation constant and characteristic impedance are [29]:

$$\gamma_{\text{int}} = j\sqrt{\left(\omega L' - \dfrac{1}{\omega C'_{\text{int}}}\right)(\omega C')} =$$
$$j\sqrt{\omega^2 L'C' - \dfrac{C'}{C'_{\text{int}}}} = j\beta_{\text{int}} \tag{10}$$

$$Z_{c,\text{int}} = \sqrt{\dfrac{j\left(\omega L' - \dfrac{1}{\omega C'_{\text{int}}}\right)}{j\left(\omega C'\right)}} = \sqrt{\dfrac{L'}{C'} - \dfrac{1}{\omega^2 C'C'_{\text{int}}}} \tag{11}$$

It is clear from the above equations that γ_{int} and $Z_{0,int}$ are real for $\omega > \omega_{se} = \dfrac{1}{\sqrt{LC_{\text{int}}}}$.

It is worth mentioning that the ITLs are categorized as complex right/left handed materials (CRLH). These metamaterial structures include also the right-handed (RH) effects that occur naturally in practical LHMs. In fact, a purely LH transmission line cannot exist since parasitic shunt capacitance and series inductance effects are unavoidable in practice [29].

4. Design of DNB Filter

Configuration of the proposed DNB filter (Top and bottom layers) is shown in Fig. 4. The synthesis of the left-handed transmission lines is done by properly etching SRRs in the host line [32]. Compared to the dual transmission line approach, we can term this approach the resonant-type approach, since resonators (SRRs) are utilized as loading elements. Resonant type metamaterial transmission lines also exhibit a composite right/left-handed (CRLH) behavior. The natural host lines for the implementation of SRR-based left-handed metamaterials in one dimension are the microstrip line and the coplanar waveguide. SRRs provide the negative effective permeability and must necessarily be etched in the upper substrate side in close proximity to the conductor strip to achieve high line-to-SRRs magnetic coupling. In this case, the magnetic field generated by the current flowing through the line exhibits a significant component in the axial direction of the rings. Under these conditions, the rings will be properly excited and the overall structure will be expected to behave as an effective medium with negative-valued permeability in a certain band above the resonance frequency of the SRRs. Moreover, as long as the substrate is thin enough, the magnetic flux generated by the current flowing through the line can efficiently penetrate the SRRs and hence a high magnetic coupling between line and rings is expected. A stopband appears in the vicinity of f_0=3.5GHz, the resonance frequency of the SRRs. In fact, the forbidden band extends above and below f_0. This stopband can be interpreted as a consequence of the properties of the structure, which behaves as a one-dimensional effective medium with negative magnetic permeability in a narrow band above resonance and with high positive permeability in a narrow band below resonance, which causes a strong mismatch at the feeding port of the line. Indeed, outside the forbidden band, the signal does not see the presence of the rings and it is propagated between the input and output ports of the structure. On the other hand, the proper host transmission line for the implementation of one-dimensional CSRRs is the microstrip configuration. Etching CSRRs in the ground plane, a significant component of the electric field will be paralleled to the rings' axis. Similarly, Due to the negative effective permittivity in the vicinity of CSRR's resonance, the signal will be inhibited in a narrow band. So, a stopband appears in the vicinity of f_0=2.4GHz, the resonance frequency of the CSRR. Consequently, by combining these two methods to produce stopbands, we achieve a dual notch bands. The proposed filter is designed to operate at 2.4 and 3.5 GHz with good band-stop performance, high transmission of the sideband frequencies, and
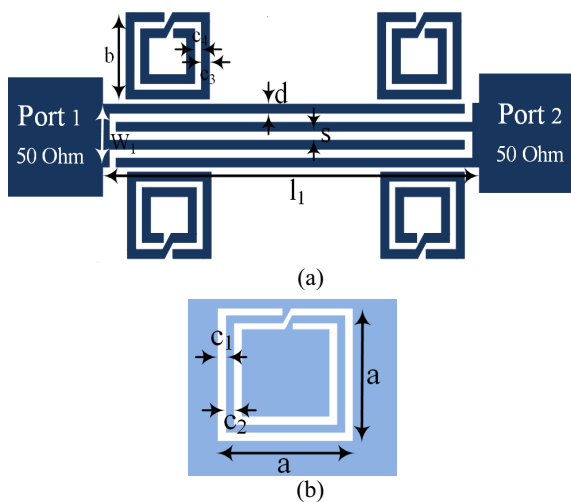
Fig. 4. Configurations of the proposed UWB filter. (a) top and, (b) bottom layers.
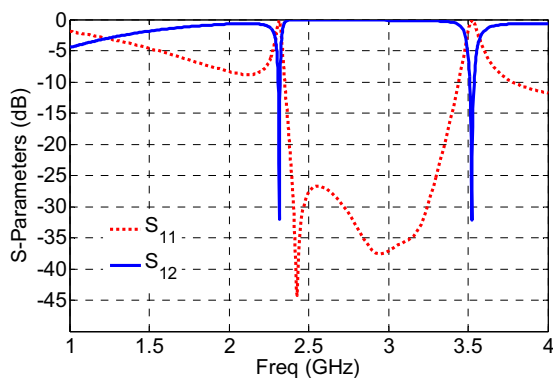
low sensitivity of its transmission response to the angle of incidence, to block the undesired frequency in UWB band. To the best of our knowledge, the size of proposed DNB filter is more compact in comparison with known similar filters. The optimized design dimensions of the suggested DNB filter, which are shown in Fig. 4, can be seen in Table 1. The substrate is chosen to be Rogers RO3210 with the relative permittivity 10.2 and the thickness of 2.54 mm. Also in the simulations, the metallic and dielectric loss have been taken into account by using the conductivity of copper $\sigma = 5.8 \times 10^7$ S/m and the loss tangent $\tan\delta = 0.0027$ of the substrate. The overall dimensions of the designed DNB filter are less than $0.074\ \lambda_0 \times 0.032\ \lambda_0$ (7.4 mm × 4.8 mm), where λ_0 is the free space wavelength at the first center notch-bands.
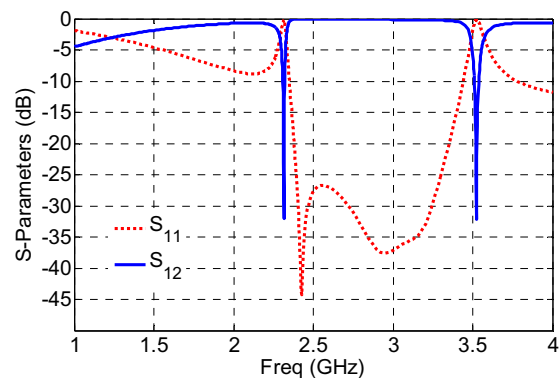
5. Results and Discussion

The frequency response of the proposed DNB filter is simulated by Agilent ADS simulator and presented in Fig. 5. As it can be seen, the stopband with low frequency of 2.40 GHz has less than 0.48 dB insertion loss and greater than 32 dB return loss and the high frequency stopband of 3.5 GHz has less than 0.32 dB insertion loss and greater than 33 dB return loss. The 3-dB bandwidth is 110MHz from 2.37GHz to 2.48GHz and 390MHz from 3.26GHz to 3.65GHz. Thus, in the dual notch bands, the fractional frequency bandwidths are about 4.7% and 6.8%. Moreover, a wide stopband between two notch bands with the insertion loss higher than 30 dB in the range of 2.43 to

Fig. 6. Frequency responses of the proposed dual stopband filter when the length of the interdigital TL is 5mm, 6mm, 7mm and 8mm, respectively.

3.47 GHz is achieved. In Fig. 6, the influence of resizing the length of the interdigital TL on the frequency response of the dual stopband is shown. As can be seen, by reducing the size of the interdigital capacitor (shown in fig. 4(a) by l1), the upper notch band can be easily shifted to the higher frequencies and also, by increasing the size of the interdigital capacitor, the upper notch band can be shifted to the lower frequencies. Moreover, we can shift the upper notch band to the higher frequencies and the lower notch band to the lower frequencies by resizing the size of the SRRs and CSRR, respectively. Consequently, we able to easy shift the frequency for lower and upper stop bands by resizing the size of the interdigital TL, SRRS and CSRR.

As for the group delay, this filter has a good performance too. Fig. 7 indicates that the group delay is less than 1.4ns in the lower band and 1.1ns in the upper stopband. The maximum variation of the group delay is also about 0.3ns.

Table 1. Design parameters of the proposed dual notch-band filter in millimeters.

l_1=7.4	w_1=1.4	a=3.2	b=1.6	d=0.2
s=0.2	c_1=0.2	c_2=0.2	c_3=0.2	c_4=0.2

Fig. 5. Simulated frequency responses of the proposed DNB filter.
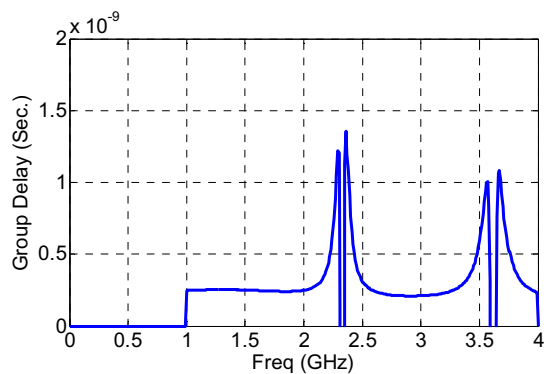
Fig. 7. Simulated Group delay of the proposed DNB filter (maximum group delay variation 0.3ns).
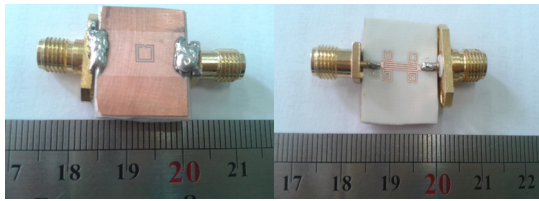
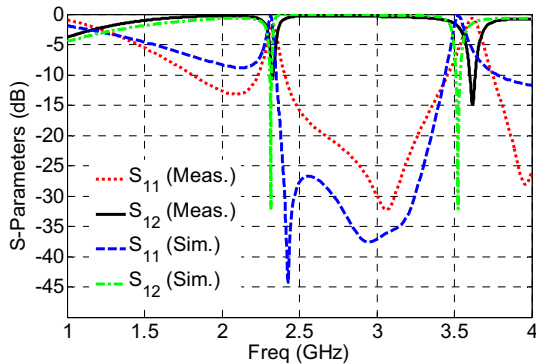**Fig. 8.** Top and bottom views of the fabricated NBP filter.

Fig. 9. Simulated and Measured frequency responses of the proposed DNB filter.

Table 2. Performance Comparisons of the Recent Filter with the Other.

Reference number	Insertion loss(dB) lower/upper	Return loss (dB) lower/upper	Size $\lambda_o^1 \times \lambda_o^1$ (lower passband)
[7]	1.93/1.89	18/15	0.44×0.59
[33]	1.45/1.74	20/24	0.27×0.39
[34]	2.5/2.8	21/29	1.16×0.36
[35]	1.5/1.7	16/18	0.20×0.13
[36]	0.8/1	15/16	0.19×0.19
This Work	0. 8/0.9	12/15	0.074×0.032 (7.4×4.8 mm²)

6. Conclusion

In this paper, a novel microstrip dual notch band filter (DNBF) is proposed by using an interdigital transmission line that is loaded with SRRs and CSRR. This filter can be used for the rejection of spurious signal frequencies and prevention of wave interference. The proposed filter is designed to operate at 2.4 and 3.5 GHz with good band-stop performance, high transmission of the sideband frequencies, and low sensitivity of its transmission response to the angle of incidence. The measurement results show that the transmission response has more than 12 dB rejection around the lower stop band and 15 dB rejection near the higher stop band. It has been also shown that the proposed microstrip dual notch-band filter can provide good rejection between two notch-bands and each of the notch-band's frequencies can be independently adjusted. Consequently, the proposed DNBF filter have many advantages in terms of compact size, low insertion losses, high return losses, high selectivity, high Q-factor, easy band-stop frequency shifting, easy fabrication and, easy integration with other planar microwave circuits.

After evaluating the proposed filter performance with ADS, the proposed filter has fabricated (Fig. 8) and measured using the network analyzer Rohde & Schwarz zvk. As it is shown in Fig. 9, there is a good agreement between the measurement and simulation results. However, there is some deviation especially small resonant frequency shift between the simulation and measurement that may be attributed to the fabrication related tolerances. The measured return losses are better than 17 dB, and the measured band insertion losses are smaller than 1.15 dB. The results show that the designed filter has a competitive attenuation slope in the lower and upper pass band transitions, respectively. At last, a comparison is done between the proposed filter and the other reported DNB filters (Table 2) which shows that the designed filter has both the better performance and smaller size in comparison with the others. In fact, as the elements of the basic cell including the SRRs, CSRR, and ITL are all electrically small, we could achieve a very compact filter. Totally, the proposed DNB filter have advantages of small size, low insertion loss, high return loss, and the ease of band-stop frequencies shifting.

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