Design of a wideband parallel-coupled filtering antenna based on dual-frequency matching condition

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Abstract In this paper, a wideband filtering antenna is designed based on the dual-frequency matching condition. Instead of matching at only one radiating frequency in the conventional filtering antenna design, an antenna element with two radiating frequencies are considered and equivalently modeled as two series-connected parallel RLC resonators, a series inductor, as well as a shunt capacitance. To obtain a wideband impedance matching, last two stages of the parallel coupled-line sections in a filter are synthesized for matching the antenna at two frequencies. Finally, a prototype of a four-pole filtering antenna is designed and fabricated. The measured results achieve a wide bandwidth of 27.8% over 2.12–2.81 GHz with a flat antenna gain of 3.41 dBi over all the filtering band.

Keywords: coupled line, filtering antenna, wideband
Classification: Microwave and millimeter-wave devices, circuits, and modules

1. Introduction

With the rapid development of wireless communication technology, the multi-functional microwave devices have become more attractive because of the size reduction and overall performance improvement. As an integration of the filter and antenna, the filtering antenna due to its high radiation efficiency, sharp frequency selectivity and good harmonic suppression has been widely studied in recent years.

Traditionally, there are two ways to achieve antennas with filtering characteristics. One approach is that the bandpass filter and antenna are connected by the matching networks [1, 2]. This design method always suffers from higher loss induced by the extra impedance transformer. Another approach is based on the synthesis technique of the filter. In [3, 4, 5, 6, 7, 8, 9, 10, 11, 12], the filtering antennas were fabricated with the antenna regarded as the last resonator of the filter, which can reduce the size and realize a direct impedance matching. A kind of wideband circularly polarized filtering antennas were demonstrated with the good in-band responses and flat gains [13, 14, 15, 16]. Since these works were synthesized and designed only at the center radiating frequency, the fractional bandwidth was limited. Therefore, a planar filtering ultra-wideband (UWB) antenna was proposed by combining the shorted stubs and the UWB monopole antenna. However, the synthesis method was not valid [17]. Recently, filtering antennas in different forms have been proposed based on a new design concept [18, 19, 20, 21, 22, 23, 24, 25]. The parasitic elements were integrated in the radiating element without extra filtering circuits. For example, shorting vias, stacked patch and U-slot are applied for filtering function. The diplexer which consists of two channel filters can be also integrated with antennas [26, 27, 28, 29, 30].

In this paper, a synthesis method for wideband filtering antenna is proposed. To achieve a good in-band performance, the last two parallel coupled line sections in the filter design are utilized together and modified to match the antenna at two radiating frequencies. To verify this design method, a forked monopole antenna and five parallel coupled microstrip line sections are integrated to be a wideband filtering antenna. The measured results demonstrate a good filtering performance and a flat gain of 3.41 dBi over a wide frequency band from 2.12 GHz to 2.81 GHz.

2. Synthesis method

![Fig. 1. Configuration of the wideband filtering antenna with \(w_{st} = 2\) mm, \(l_{1} = 21\) mm, \(w_{a1} = 1\) mm, \(l_{2} = 20.8\) mm, \(l_{3} = 6\) mm, and \(l_{4} = 11\) mm.](Image)

![Fig. 2. The equivalent circuit of \(N^{th}\) order BPF.](Image)

The proposed wideband filtering antenna is shown in Fig. 1, which consists of \(N + 1\) coupled line sections and a forked monopole antenna. The parallel coupled-line sections are directly matched to the antenna without additional networks. An \(N^{th}\) order bandpass filter (BPF) which has similar bandwidth with the forked antenna is firstly considered. Its equivalent circuit is shown in Fig. 2. According to the traditional synthesis method of the filters
[31], the even and odd-mode characteristic impedances, \( Z_{0e} \) and \( Z_{0o} \), can be obtained easily. Similarly, Fig. 3 demonstrates the equivalent circuit of the filtering antenna, \( J_1 \ldots J_{N-1} \) are same with the BPF. Then, the forked monopole antenna is modeled as two series-connected parallel RLC resonators (\( R_1, L_1, C_1 \) and \( R_2, L_2, C_2 \)), a series inductor \( L_3 \) and shunt capacitance \( C_3 \). As shown in Fig. 4, the input admittance of the antenna calculated by the equivalent circuit model agrees well with the full-wave simulated results over the frequency range from 2.0 GHz to 2.8 GHz.

![Fig. 3](image-url)

**Fig. 3.** The equivalent circuit of the proposed filtering antenna with \( R_1 = 353 \, \text{Ω}, L_1 = 8 \, \text{nH}, C_1 = 0.62 \, \text{pF}, R_2 = 8.95 \, \text{Ω}, L_2 = 0.028 \, \text{nH}, C_2 = 150 \, \text{pF}, L_3 = 8 \, \text{nH}, \) and \( C_3 = 0.56 \, \text{pF} \).

To obtain a good matching between the filters and antennas, the conventional method [3, 4, 5, 6] optimized the last coupled line section to make the input admittance of the filter equal to the filtering antenna at the center operating frequency. In order to match the antenna over a wide band, the last two coupled line sections of the filter are considered and optimized together in this work. Therefore, two unknown variables \( J_N \) and \( J_{N+1} \) are further introduced in the filtering antenna equivalent circuit. As shown in Fig. 3, the input admittance \( Y_{in1}, Y_{in2}, \) and \( Y_{in3} \) can be derived as

\[
Y_{in1} = Y_{N+1} \pm j \frac{Y_{N+1} \tan(\beta L_{N+1})}{Y_{in1}} + j Y_{in1} \tan(\beta L_{N+1}) \tag{1}
\]

\[
Y_{in2} = \frac{j Y_{N+1}^2}{Y_{in1}} \tag{2}
\]

\[
Y_{in3} = Y_{N+1} \pm j Y_{N+1} \tan(\beta L_{N+1}) + j Y_{in2} \tan(\beta L_{N+1}) \tag{3}
\]

then the \( Y_{in}^0 \) can be obtained as

\[
Y_{in}^0 = \frac{J_N^2(Y_{in1} + j Y_{in1} \tan(\beta L_{N+1})) + j Y_{in1} \tan(\beta L_{N+1})}{Y_{in1} + j Y_{in1} \tan(\beta L_{N+1})} \tag{4}
\]

On the other hand, \( Y_{in1}^0 \) can be easily calculated according to the well-synthesized filter. Let \( Y_{in1}^0 = Y_{in}^0 \) at two different in-band frequencies, the unknown variables \( J_N \) and \( J_{N+1} \) can be solved. Finally, the even- and odd-mode characteristic impedance can be obtained [10].

3. Design example

In order to verify the proposed synthesis method, a wideband filtering antenna is designed and implemented, which consists of five coupled line sections and a forked monopole antenna. The proposed filtering antenna is fabricated on Rogers 4350B substrate with a relative permittivity of \( \varepsilon_r = 3.66 \) and a thickness of 0.508 mm. According to the synthesis method, a 4\(^{th}\) order BPF operating at 2.45 GHz is firstly considered with the bandwidth range from 2.14 GHz to 2.83 GHz (27.8% fractional bandwidth), and the even- and odd-mode characteristic impedances of each couple line can be obtained as

\[
(Z_{0e1}, Z_{0o1}) = (Z_{0e5}, Z_{0o5}) = (107.77 \, \Omega, 48.3 \, \Omega) \tag{5}
\]

\[
(Z_{0e2}, Z_{0o2}) = (Z_{0e4}, Z_{0o4}) = (72.03 \, \Omega, 40.29 \, \Omega) \tag{6}
\]

\[
(Z_{0e3}, Z_{0o3}) = (65 \, \Omega, 39.41 \, \Omega) \tag{7}
\]

The parameters of the couple lines can be extracted as \( w_1 = w_5 = 0.44 \, \text{mm}, w_2 = w_4 = 0.82 \, \text{mm}, w_3 = 0.97 \, \text{mm}, s_1 = s_5 = 0.1 \, \text{mm}, s_2 = s_4 = 0.13 \, \text{mm}, s_3 = 0.13 \, \text{mm}, l_1 = l_5 = 18.6 \, \text{mm}, l_2 = l_4 = 17.66 \, \text{mm}, \) and \( l_3 = 18 \, \text{mm} \). Then, the input admittance \( Y_{in}^0 \) in Fig. 2 can be easily obtained [32]. Finally, the inverter \( J_4 \) of the fourth couple line and \( J_5 \) of fifth coupled-line can be calculated according to proposed method, as shown in Fig. 3.

To simplify the calculation, \( f_1 = 2.49 \, \text{GHz} \) and \( f_2 = 2.57 \, \text{GHz} \) are chosen here to make \( \beta L_4 = \pi/4 \) at \( f_1 \) and \( \beta L_4 = \pi/4 \) at \( f_2 \). And the input admittance \( Y_{in}^0 \) can be calculated as \( Y_{in1}^0 = 0.0668 + j 0.0073 \) at \( f_1 \) and \( Y_{in2}^0 = 0.0716 + j 0.0144 \) at \( f_2 \). In addition, \( Y_{in}^0 = 0.016 \, \text{S} \) is selected for the easy fabrication. After calculation, the corresponding even- and odd-mode characteristic impedance of last two coupled lines of the filtering antenna are obtained as

\[
(Z_{0e4}, Z_{0o4}) = (65.25 \, \Omega, 40.75 \, \Omega) \tag{8}
\]

\[
(Z_{0e5}, Z_{0o5}) = (94.04 \, \Omega, 47.95 \, \Omega) \tag{9}
\]

The parameters of last two couple lines are optimized to be \( w_4 = 0.92 \, \text{mm}, s_4 = 0.17 \, \text{mm}, l_4 = 17.66 \, \text{mm}, w_5 = 0.53 \, \text{mm}, s_5 = 0.11 \, \text{mm}, \) and \( l_5 = 18.11 \, \text{mm} \).

4. Simulated and measured results

For the comparison, a forked monopole antenna directly connected with 4\(^{th}\) order BPF are also designed respectively. According to the proposed synthesis method, the size of last two coupled lines in filtering antenna are recalculated and modified when compared with the config-
impedance matching), proposed filtering antenna and forked monopole antenna.

Fig. 5. Simulated results of the 4th order BPF, 4th order BPF + antenna (the last two parallel coupled lines of the BPF are not adjusted for impedance matching), proposed filtering antenna and forked monopole antenna.

Fig. 6. Simulated and measured $|S_{11}|$ and gain of the proposed filtering antenna.

Fig. 7. Simulated and measured radiation patterns of the proposed filtering antenna. (a) x-y plane at 2.3 GHz. (b) x-y plane at 2.7 GHz.

uration of 4th order BPF connected directly with the forked monopole antenna. As a result, a wideband impedance matching and a better band selectivity performance can be achieved. As shown in Fig. 5, the forked monopole antenna shows a wideband performance without band selectivity function, while the proposed filtering antenna shows a good selectivity at the band edge as the 4th order BPF. The simulated 10-dB bandwidth of the proposed filtering antenna is achieved from 2.12 GHz to 2.81 GHz, which is much better than the BPF directly connected with the antenna (2.34 – 2.83 GHz), and only a slightly frequency deviation of 0.02 GHz exist between the BPF (2.14 – 2.83 GHz) and the proposed filtering antenna. The simulated total radiated power here represents the power radiated by antenna in all directions, which is suitable to be compared with $|S_{11}|$ of filter. The total radiated power is normalized to the input power. It can be found in the passband that the normalized total radiated power is as flat as the $|S_{11}|$ of the BPF, while the skirt selectivity is even better than the BPF.

Finally, the filtering antenna is fabricated and measured. The photograph of the fabricated prototype and measured results are shown in Fig. 6. The measured 10-dB bandwidth is almost the same with the simulated one over the frequency band from 2.12 GHz to 2.81 GHz. The measured maximum antenna gain is 3.41 dBi at 2.65 GHz, while the simulated one is 3.83 dBi at 2.7 GHz. Fig. 7 shows the simulated and measured far field radiation patterns at 2.3 GHz and 2.7 GHz in x-y plane. It can be notice that the simulated and measured results achieve a reasonable agreement from each other, which effectively validate our design approach.

5. Conclusion

A wideband filtering antenna design with a dual-frequency matching has been presented. The impedance matching condition at two different frequencies have been derived and successfully applied in the filtering antenna design with two radiating frequencies. This method could be further extended to model a wider band by involving more coupled-line sections.

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References


