A New Method for Designing Modified Compact Microstrip LPF with Sharp Roll-Off and Wide Stopband

Abbas HOSEINABADI¹, Mohammad Bagher TAVAKOLI², Mohammad Jalal RASTEGAR FATEMI¹, Farbod SETOUDEH¹

¹ Dept. of Electronics, Coll. of Engineering, Saveh Branch, Islamic Azad University, Saveh, Iran
² Dept. of Electronics, Coll. of Engineering, Arak Branch, Islamic Azad University, Arak, Iran
³ Dept. of Electrical Engineering, Arak University of Technology, Arak, Iran

{a-hosseinabadi, m-tavakoli}@iau-arak.ac.ir, j_rastegar@iau-saveh.ac.ir, f.setoudeh@arakut.ac.ir

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Abstract. A new method for designing a compact microstrip lowpass filter (LPF) with wide stopband width (SBW) and sharp roll off (ROF) is presented. In the proposed designing procedure, high impedance microstrip lines are bent to achieve an LPF with compact size. Then, to compensate for the effect of bending microstrip lines, the lengths of the lines are mathematically modified. Moreover, adding a suppressing cell composed of a radial stub resonator and a butterfly stub resonator increases the SBW. Also, an elliptic filter structure is used to obtain sharp ROF. In this work, an LPF with 1.12 GHz cutoff frequency, 0.147 λg × 0.133 λg filter size; where λg is the guided wavelength at cutoff frequency, the SBW equal to 13.4 GHz, and the ROF more than 201 dB / GHz, is designed, simulated and fabricated to demonstrate efficiency of the proposed method. Also, the other conventional characteristics for the fabricated LPF such as 0.3 dB insertion loss, 14.4 dB return loss, and suppression factor equal to 2.2, are in the appropriate range of their amounts.

Keywords

Lowpass filter, microstrip, elliptic filter, stopband width, roll off, compact size

1. Introduction

In recent decades, the communication systems, especially the microwave applications, benefited from the microstrip filters. The lowpass filters (LPFs) are widely used in most of the communication systems to reject unwanted high frequency signals from system answers [1, 2]. Newly, researchers tried to improve microstrip LPF characteristics such as increasing relative stopband bandwidth (RSB), sharpening roll off (ROF), compacting LPF normalized circuit size (NCS), obtaining low insertion loss (IL) and high return loss (RL) in the passband region, and enhancing suppression factor (SF) in the stopband zone; whereas the ROF, the NCS and the RSB were more considered [3–21]. It should be noted that the RSB, NCS, ROF and SF parameters are commonly used for the LPF and are defined in article [3]. To obtain the high ROF in an LPF that is proposed in [4], stepped impedance modified T-type with stepped impedance stub is used, but it does not have high RSB and the IL is not in the appropriate range. In [5] an LPF using T-shaped resonators is presented to achieve a sharp transition skirt; however its RSB is very small. In [6], the proposed LPF is comprised of three coupled stepped impedance resonators to create the high ROF, while one of disadvantages of this work is the low SF. In some of the reviewed articles, the NCS is more considered. In [7], a folded three-element LPF is proposed to achieve compact size, but its ROF is very low. In other two works [8], [9] the feeding lines are folded to reduce filter size, nevertheless in [9] the RSB is not high and in [8] the ROF is middle and the RSB is also low. Considering the importance of the RSB in the LPF many of the recent papers try to improve the RSB [10–18]. In [10–12] by using stepped impedance resonators the LPFs are designed, and although those have good RSB, their ROFs are very low. In another article [13], the LPF structure is composed of a defected rectangular-shaped resonator and a rectangular shape resonator as a suppressing cell, and amount of its RSB is high, also its NCS is low, nevertheless its ROF is not appropriate. An LPF with wide stopband and low NCS using triangular-shaped resonator is presented in [14], but its ROF is low. Other structures which are used for LPF designing include defected ground structure (DGS) to improve the RSB [15–18]. It is true that using the DGS improves the stopband width, but it complicates filter designing. Also the LPFs that are presented in [15–18] have not high ROF. In another effort an LPF with ultra-wide stopband is proposed [19], while its ROF is moderate and its IL is not acceptable and it also has a complex design. In article [20], using a meandered semi-hairpin resonator which is combined with a suppressing unit, an LPF with wide stopband and small NCS is presented and the disadvantages of it are small ROF and complexity in filter design. In [3], an LPF with good ROF and RSB consisting of two resonators with different triangular patches and four
high-low impedance resonators as suppressing cells has been designed. Finally, in [21], another LPF with good ROF and RSB consisting of three cascaded resonators with different semi-circle patches and four suppressors employing radial stubs has been proposed. But in the last two filters, the NCS is relatively large and those have complex design, too.

In this paper, the use of an elliptic filter structure causes a high ROF. To achieve miniature filter size, the high impedance microstrip lines are bent and their lengths are modified by mathematical analysis. Then as a result of adding suppressing cell, which consists of a combination of a radial stub resonator and a butterfly stub resonator, the RSB increases. At last, to show efficiency of the designed LPF, it was fabricated and measured.

2. LPF Design Process

2.1 Conventional Elliptic LPF Design

Due to the desired characteristics of an elliptic filter such as high ROF, high SF, and low IL, a six-pole elliptic LPF with 0.1 dB passband ripples and 1.1 GHz cutoff frequency is introduced. Figure 1a shows the LPF prototype circuit with lumped elements. In Fig. 1b layout of the six-pole microstrip LPF is depicted. All inductors and capacitors are realized using high-impedance and low-impedance lines respectively. Characteristic impedance for high-impedance lines \( Z_{0L} = 131 \, \Omega \), and characteristic impedance for low-impedance lines \( Z_{0C} = 20 \, \Omega \) are selected. In addition, to calculate high-impedance line lengths \( l_i \) for \( i = 1, 2, \ldots, 5 \) and low-impedance line lengths \( l_j \) for \( j = 2, 4, 6 \), substituting the corresponding parameters from Tab. 1 in (1) is used [2]:

\[
\begin{align*}
  l_i &= \frac{\lambda_{bl}}{2\pi} \sin^{-1}\left(\frac{2\pi f_c L_i}{Z_{0L}}\right), \\
  l_j &= \frac{\lambda_{bc}}{2\pi} \sin^{-1}\left(2\pi f_c Z_{0C} C_j\right)
\end{align*}
\]

where \( f_c = 1.1 \, GHz \) is the desired cutoff frequency and \( \lambda_{bl} \) and \( \lambda_{bc} \) are guided wavelengths for high-impedance and low-impedance lines at \( f_c \) respectively.

For the LPF with lumped elements two finite-frequency attenuation poles occur at \( f_{p1} \) and \( f_{p2} \) that are presented by (2). The values of \( f_{p1} \) and \( f_{p2} \) are also given in Tab. 1 along with other related values in Fig. 1.

\[
\begin{align*}
  f_{p1} &= \frac{1}{2\pi \sqrt{L_i C_4}}, \\
  f_{p2} &= \frac{1}{2\pi \sqrt{L_i C_2}}.
\end{align*}
\]

The compensating for the effect of connecting inductor \( L_2 \) to capacitor \( C_4 \) which is a discontinuity with different widths, the simultaneous compensation for the unwanted susceptance at the junction of the inductive lines for \( L_1, L_2, \) and \( L_3 \), is achieved by correcting \( h_2 \) and \( h_3 \), while \( h_1 \) and \( h_3 \) are kept unchanged. For this purpose, equation (3) is solved by substituting the cutoff frequency \( f_c \) and the attenuation pole frequency \( f_{p2} \) instead of \( f_{p1} \):

\[
1 = \frac{1}{2\pi f_c L_2} \left(2\pi f_c Z_{0C} C_4\right)^2 = B_{b2}(f) + B_{T}(f)
\]

where according to Fig. 1b, \( B_{b2}(f) \) specifies the susceptance of \( L_2 \) and \( C_4 \) branch which was implemented by microstrip lines and will be obtained by (4):

\[
B_{b2}(f) = \frac{1}{X_{L_2}(f) - \frac{1}{B_{c2}(f)}},
\]

\[
X_{L_2}(f) = Z_{0L} \sin\left(\frac{2\pi f_c L_2}{\lambda_{bl}(f)}\right) + Z_{0C} \tan\left(\frac{\pi f_c L_2}{\lambda_{bc}(f)}\right),
\]

\[
B_{c2}(f) = \frac{1}{Z_{0C}} \left(\frac{1}{\lambda_{bc}(f)} + \frac{1}{\lambda_{bl}(f)}\right) \tan\left(\frac{\pi f_c L_2}{\lambda_{bl}(f)}\right).
\]

Note that the second term on the right in (5) shows the effect of low impedance line in the step connection related to \( L_2 \) and \( C_4 \) branch. Similarly, the second term on the right in (6) shows the effect of high impedance line in the step connection related to \( L_2 \) and \( C_4 \) branch. In (3) \( B_{T}(f) \) is unwanted susceptance resulting T-junction between three high impedance lines \( h_1, h_2 \) and \( h_3 \) and it can be written as (7):

![Fig. 1. The conventional elliptic LPF: (a) The prototype circuit with lumped elements. (b) Layout.](image)
To modify the deviation caused by bending of $l_{l2}$ and $l_{l3}$, susceptance $B_{2\text{new}}(f)$ that is presented in (8) is added to the right side of (3). Then (3) can be rewritten in form of (9):

$$B_{2\text{new}} = \frac{1}{Z_{ol}}(\tan(\pi l_{l21}/\lambda g_l(f)) + 2 \tan(\pi l_{l22}/\lambda g_l(f)) +$$

$$\tan(\pi l_{l23}/\lambda g_l(f)) + 2 \tan(\pi l_{l32}/\lambda g_l(f)) +$$

$$\tan(\pi l_{l33}/\lambda g_l(f)),$$

$$\frac{1}{2\pi l_2 - (2\pi C_2)^{-1}} = B_{s2}(f) + B_{s}(f) + B_{2\text{new}}(f).$$

The values of $l_{l31} = 3.2$ mm, $l_{l32} = 8.1$ mm and $l_{l33} = 0.9$ mm are considered but the values of $l_{l21} = 1.3$ mm, $l_{l22} = 1.2$ mm and $l_{l23} = 1.5$ mm are first estimated. After solving by substituting the cutoff frequency $f_c$ and the attenuation pole frequency $f_{p2}$ instead of $f$ in (9) new values for $l_{l2} = 3.5$ mm and $l_{l2} = 11.7$ mm are obtained. After determining new values for $l_{l2}$ and $l_{l2}$, new values $l_{l2} = 1.1$ mm, $l_{l2} = 1.2$ mm and $l_{l2} = 1.4$ mm can be selected accordingly. Correction the lengths of $l_{l4}$ and $l_{l4}$ are similar to the method described in this section and could be done by replacing the parameters proportional to the $L_4 C_4$ branch. After the correction process of the $L_4 C_4$ branch, values of $l_{l4} = 8.5$ mm and $l_{l4} = 8.3$ mm are obtained.

Figure 4 illustrates EM simulated $S_{21}$ parameters for modified bend layout by (9) for $L_2 C_2$ branch which is shown in Fig. 3b (and $L_2 C_4$ branch but not shown), beside $S_{21}$ parameters for the lumped elements circuit corresponding to connection of $L_1$, $L_2$, $L_3$ and $C_2$ (or $L_1$, $L_3$, $L_4$ and $C_4$ for $L_4 C_4$ branch). Also, Figure 4 shows EM simulated $S_{21}$ parameters for unmodified bend layout of $L_2 C_2$ and $L_4 C_4$ branches.

### 2.2 Compacting LPF Size

The dimensions of the filter can be reduced by using the technique of bending high impedance microstrip lines. Of course, it should be noted that bending the microstrip lines causes unwanted effects on the values of inductors and capacitors implemented with microstrip lines. To correct the deviation in the inductor and the capacitor values, which means deviation in electrical length of microstrip lines, the physical length of the lines must be modified. For example, Figure 3 shows how $l_{l2}$ and $l_{l3}$ are bent to reduce the filter size.

**Fig. 2.** The simulated $S_{21}$ parameters for the LPF with lumped elements and for the LPF with microstrip layout before and after correcting by (3).

Also for $L_4$ and $C_4$ branch an equation similar to (3) should be written by replacing the parameters corresponding to it then solved at $f_c$ and $f_{p2}$. After correcting, the lengths of the corrected lines are obtained as $l_{l2} = 4$ mm, $l_{l3} = 10.7$ mm, $l_{l4} = 9.2$ mm and $l_{l3} = 7.7$ mm. In Fig. 2 $S_{21}$ parameter responses for the LPF with lumped elements and microstrip layout before and after correcting by (3) are illustrated. As shown in Fig. 2 locations of $f_{p1}$ and $f_{p2}$ in the LPF with microstrip layout after correcting by (3) and the LPF with lumped elements are almost matched.

**Fig. 3.** Bending microstrip lines to reduce filter size: (a) Before bending, (b) after bending, (c) detailed.

**Fig. 4.** The $S_{21}$ parameters of the $L_2 C_2$ or $L_4 C_4$ branches: solid line (or bold solid line) for $L_2 C_2$ (or $L_4 C_4$) branch with lumped elements, dot line (or bold dot line) for $L_2 C_2$ (or $L_4 C_4$) branch with microstrip lines which bend then corrected with (9), and finally dash line (or bold dash line) for $L_2 C_2$ (or $L_4 C_4$) branch with microstrip lines which bend but not corrected with (9).
branches for comparing. As shown in Fig. 4 the location of \( f_{g1} \) and \( f_{g2} \) deviate by bending microstrip high impedance lines from values predicted with (2) however, the correction of the microstrip line lengths by (9) can match the values of \( f_{g1} \) and \( f_{g2} \) to values which obtained by (2). The \( S_2 \) parameters for the corrected bended microstrip layouts and the lumped element circuits are well matched. This match causes the modified LPF cutoff frequency closes to cutoff frequency of the LPF with lumped elements.

### 2.3 Adding Suppression Resonators

It can be seen by looking at Fig. 2 that the microstrip conventional elliptic LPF has a narrow SBW. In order to increase the SBW, a suppression cell consisting of two resonators has been used. Generating transmission zeros (TZs) at frequencies higher than the cutoff frequency can increase the SBW but this action should not interfere with the filter frequency response at frequencies below the cutoff frequency. As shown in Fig. 5, a combination of a radial stub resonator (RSR) and a butterfly stub resonator (BSR) in the middle of \( l_{1T} \) forms a suppression cell. Now \( l_{1T} \) is no longer valid and is replaced by \( l_{1T,1}, l_{1T,2} \) and \( l_{1T,3} \). The lengths of \( l_{1T,1}, l_{1T,2} \) and \( l_{1T,3} \) are selected by using ADS software tuning application. The values \( l_{1T,1} = 6.5 \) mm, \( l_{1T,2} = 6.4 \) mm and \( l_{1T,3} = 6.1 \) mm is obtained. The RSR structure is shown in Fig. 6. The values of parameters corresponding to Fig. 6a are \( R_{SR} = 7.4 \) mm, \( r_{RS} = 0.58 \) mm, \( \theta_{RS} = 120^\circ \). The equivalent lossless circuit with lumped elements is illustrated in Fig. 6b [22]. Assuming \( R_{SR} < \lambda / 8 \) and \( r_{RS} = R_{RS}/10 \), the values of \( C_{RS} \) and \( L_{RS} \) are extracted from (10) and (11) [23]:

\[
C_{RS} = \frac{\theta_{RS}R_{RS}^2\varepsilon_{eff}}{240\pi\hbar c},
\]

\[
L_{RS} = \frac{120\pi(2.8 - 10R_{RS}/r_{RS})}{c\theta_{RS}}
\]

where \( \varepsilon_{eff} \) is effective dielectric constant of microstrip line with width equal to \( w_{eq} = (R_{RS} + r_{RS})\sin(\theta_{RS}/2) \), also \( h = 0.813 \) mm is dielectric thickness and \( c \) is speed of light.

As a result, the transmission zero for the RSR (TZ1) can be written as follows:

\[
f_{TZ1} = \frac{1}{2\pi\sqrt{L_{RS}C_{RS}}}. \tag{12}
\]

A T-junction is used to connect the RSR to the high impedance microstrip line. Adding the T-junction causes the TZ1 position deviates from the value obtained by (12).

To correct adding T-junction effect the LC equivalent circuit of T-junction is appended to the LC equivalent circuit of the RSR. The appending of the LC equivalent circuit of the RSR and the LC equivalent circuit of the T-junction are shown in Fig. 6d.

The values for T-junction are \( w_{LT} = 0.2 \) mm and \( w_{LT} = 1 \) mm. The values of \( C_T \) [pF], \( L_{LT} \) and \( L_{LT} \) are extracted from (13), (14) and (15) with (16) [24], [25]:

\[
C_T = \frac{100}{\tanh(0.0072Z_0)} + 0.64Z_0 - 261.0, \tag{13}
\]

\[
L_{LT} = \frac{-w_{LT}}{h}\left[\frac{w_{LT}^2}{h} - 0.016\frac{w_{LT}}{h} + 0.064 + \frac{0.016h}{w_{LT}}L_{w,1}\right], \tag{14}
\]

\[
L_{LT} = \left[\frac{0.12w_{LT} - 0.47}{h}\frac{w_{LT}}{h} + 0.195\frac{w_{LT}}{h} - \frac{0.357 + 0.0283\sin}\left(\frac{\pi w_{LT}}{h} - 0.75\pi\right)\right]L_{w,2}, \tag{15}
\]

\[
L_{w,\sigma} = \frac{Z(w_{LT})\varepsilon_{eff}(w_{LT})}{c} \quad \text{[nH]} \tag{16}
\]

where \( Z(w_{LT}) \) and \( \varepsilon_{eff}(w_{LT}) \) are characteristic impedance and effective dielectric constant of \( w_{LT} \) or \( w_{LT} \) respectively.

The values of capacitors and inductors which related to Fig. 6 are brought in Tab. 2. Now the transmission zero for the RSR can be rewritten as follows:

\[
f_{TZ1} = \frac{1}{2\pi\sqrt{(L_{RS} + L_{LT})C_{RS}}}. \tag{17}
\]
And \( f_{TZ1} \) is obtained from (17) with value 4.3 GHz. As shown in Fig. 7 the TZ1 in both the \( S_{21} \) parameter curves of the RSR resonator with T-junction and its LC equivalent circuit is matched. In Fig. 8 the BSR structure is depicted. As shown in Fig. 8a the BSR structure is created by combining two RSR with one cross junction. Also the LC equivalent circuit of the BSR is depicted in Fig. 8c. The values of the BSR are \( R_{BS} = 4 \text{ mm}, \quad r_{BS} = 0.47 \text{ mm}, \quad \theta_{BS} = 65^\circ. \) The values of cross junction are \( w_{1CJ} = 0.2 \text{ mm} \) and \( w_{2CJ} = 0.51 \text{ mm}. \) The value of the parameters of \( C_{BS} \) and \( L_{BS} \) are extracted from (10) and (11) by substituting the RSR parameters with the related BSR parameters. In addition, the calculation of parameters \( C_{CJ}, L_{1CJ}, L_{2CJ} \) and \( L_{3CJ} \) is performed using (A1), (A2), (A3) and (A4) [24], [25] which are given in Appendix A. Also, the values of \( C_{CJ}, L_{1CJ}, L_{2CJ} \) and \( L_{3CJ} \) are brought in Tab. 2. In next, ignoring the parameter \( C_{CJ}, \) the transmission zero of the BSR (TZ2) can be written as:

\[
f_{TZ2} = \frac{1}{2\pi \sqrt{(L_{BS} + L_{2CJ} + 2L_{3CJ})C_{BS}}}. \tag{18}
\]

The \( f_{TZ2} \) is obtained from (18) with value 10.38 GHz. This value is near to the \( f_{TZ2} \) which is obtained from EM simulation of the BSR (12 GHz). As shown in Fig. 9 the TZ2 in both the \( S_{21} \) parameter curves of the BSR resonator and its LC equivalent circuit is almost matched.

Also, Figure 10 shows the \( S_{21} \) parameters of the RSR and the BSR resonators separately, combined and those LC equivalent circuits. As shown in Fig. 10 both resonators and combination of them have low IL in the passband.
These resonators add two transmission zeros at $f_{TZ1} = 4.3$ GHz and $f_{TZ2} = 12$ GHz, but due to their appropriate frequency responses at low frequencies do not appear significant change in the $S_{21}$ parameter in the passband region. Also attenuation of better than 20 dB of the suppression cell in the frequency range of 3.5 GHz to 14.4 GHz causes the stop bandwidth to be increased appropriately.

3. Comparing Results of Simulation and Measurement

Figure 11 shows the final design of the proposed LPF. Most of the values of the parameters related to Fig. 11 were determined in the previous sections however, the newly introduced parameters are $l_{5.11} = 3$ mm, $l_{5.12} = 3.5$ mm, $l_{5.31} = 0.9$ mm, $l_{5.32} = 1.5$ mm, $l_{5.33} = 1$ mm, $l_{5.34} = 0.8$ mm, $l_{5.35} = 1.1$ mm, $l_{41} = 6.7$ mm and $l_{42} = 1.8$ mm. The proposed LPF is simulated with the Agilent ADS software. Then the LPF is fabricated on RO4003C substrate with 0.813 mm thickness, dielectric constant ($\varepsilon_r$) equal to 3.38 and loss tangent ($\delta$) equal to 0.0021. The fabricated LPF is measured with the Agilent E8362B network analyzer. The actual photograph of the fabricated LPF is shown in Fig. 12. The results of $S_2$ and $S_1$ parameters simulation and measurement are illustrated in Fig. 13. Based on measurement results, the cutoff frequency is set at 1.12 GHz. The 20 dB attenuation SBW is obtained 13.4 GHz (from 1.2 to 14.6 GHz) which it causes the RSB to be equal to 1.7 and the ratio of the SBW to cutoff frequency ($SBW/f_c$) to be equal to 12. However the IL is less than 0.3 dB (0–816 MHz) and the RL is at least 14.4 dB, in addition the SF is 2.2.

For six-pole elliptic LPF with lumped elements which is shown in Fig. 1a the ROF from 3 dB to 40 dB is a maximum of about 269 dB/GHz. For the proposed LPF a good ROF more than 201 dB/GHz is obtained which means 75% of the LPF with lumped elements. In order to achieve a higher ROF, the degree of filter must be larger, which makes the IL unsuitable in addition to filter complexity. Table 3 compares the important parameters of filter with the results of the recent reviewed research. Note that some of compared articles in Tab. 3 have high ROF [4–6] however, compared to the present work in [4] the IL is not appropriate and in [4], [5] the RSB is very small. Also, in [6] the SF is small and other important parameters are not very appropriate. Using the microstrip line bending method the value of the NCS was reduced from 0.255 $\lambda_g \times 0.125 \lambda_g$ (for not bended LPF) to 0.147 $\lambda_g \times 0.133 \lambda_g$, that means 39% reduction. As can be deduced from Tab. 3, the method that presented in this paper works well for increasing the SBW and decreasing the NCS, while high ROF is obtained and the other LPF important parameters are in the proper range. It is necessary to mention that ROF (ζ), SF, RSB and NCS parameters for the LPF are very common and widely used parameters. The ROF is defined as:

$$\zeta = \frac{\alpha_{\text{max}} - \alpha_{\text{min}}}{f_s - f_c}$$

where in this paper $\alpha_{\text{max}} = 40$ dB, $\alpha_{\text{min}} = 3$ dB and $f_c$ is the first frequency whose attenuation is 40 dB. The SF, the RSB and the NCS are given by (20), (21) and (22) respectively:
4. Conclusion

In this paper by using the microstrip lines bending technique the LPF size reduced 39%. In order to match $S_{21}$ responses between the bended LPF and the lumped elements circuit, the microstrip line lengths were modified mathematically. Adding surplus suppression resonators led to increase the SBW up to 12 times the cutoff frequency. Taking the advantage of the elliptic structure for the LPF caused a high ROF up to 201 dB / GHz. Low IL, high RL, high SF, simple architecture and low cost are other benefits of the proposed LPF. Finally, close matching results between simulation and measurement represents advantages of using the method proposed in this article.

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Appendix A: Computational Equations of $C_{CJ}$, $L_{1CJ}$, $L_{2CJ}$ and $L_{3CJ}$

In this appendix, the computational equations for the LC equivalent circuit of cross junction (A1), (A2), (A3) and (A4) are presented. All necessary parameters and the results of calculation have been given in Sec. 2.3.

\[
C_{CJ} = \left[ \frac{86.6w_{CJ}}{h} - 30.9 \sqrt{\frac{w_{CJ}}{h} + 367.0} \right] \log \left( \frac{w_{CJ}}{h} \right) + \left( \frac{w_{CJ}}{h} \right)^3 + \frac{74.0w_{CJ}}{h} + 130.0 \left( \frac{w_{CJ}}{h} \right)^{3/2} - 240.0 + \frac{2.0}{w_{CJ}} - 1.5 \frac{w_{CJ}}{h} \left( 1.0 - \frac{w_{CJ}}{h} \right), \tag{A1}
\]

\[
L_{1CJ} = \left[ \frac{165.6w_{CJ}}{h} + 31.2 \sqrt{\frac{w_{CJ}}{h} + 11.8} \right] \frac{w_{CJ}}{h} - 32.0 \frac{w_{CJ}}{h} + 3.0 \left( \frac{w_{CJ}}{h} \right)^{3/2}, \tag{A2}
\]

\[
L_{2CJ} = \left[ \frac{165.6w_{CJ}}{h} + 31.2 \sqrt{\frac{w_{CJ}}{h} + 11.8} \right] \frac{w_{CJ}}{h} - 32.0 \frac{w_{CJ}}{h} + 3.0 \left( \frac{w_{CJ}}{h} \right)^{3/2}, \tag{A3}
\]

\[
L_{3CJ} = 337.5 + \left( 1 + \frac{7h}{w_{CJ}} \right) \frac{h}{w_{CJ}} - \frac{5.0w_{CJ}}{h} \cos \left( \frac{\pi (1.5h - w_{CJ})}{2h} \right). \tag{A4}
\]