Ultra-Wideband, Differential-Mode Bandpass Filters with Four Coupled Lines Embedded in Self-Complementary Antennas

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SUMMARY A design method for an ultra-wideband bandpass filter (BPF) with four coupled lines has been developed. For demonstration purposes, Ω-matched self-complementary antennas integrated with the ultra-wideband, differential-mode BPF with four coupled lines, a notch filter, and a low-pass filter (LPF) were prepared and tested. An optimized structure for a single-stage, broadside-coupled and edge-coupled four-lines BPF was shown to exhibit up to 170% fractional bandwidth and an impedance transformation ratio of 1.2 with little bandwidth reduction, both analytically and experimentally. Using the optimized structure, 6-stage BPFs were designed to transform the self-complementary antenna’s constant input impedance (60Ω) to 50Ω without degrading bandwidth. In addition, two types of filter variations—a LPF-embedded BPF and a notch filter-embedded BPF—were designed and fabricated. The measured insertion loss of both filter systems was less than 2.6 dB over the ultra-wideband (UWB) band from 3.1 GHz to 10.6 GHz. The filter systems were embedded in the wideband self-complementary antennas to reject unnecessary radiation over the next pass band and 5-GHz wireless LAN band.

key words: differential mode, antennas, band-pass filters, notch filters, low-pass filters, impedance transformer

1. Introduction

The UWB radio system is very promising, since transmission data rates greater than those of other wireless LAN systems are possible with less power dissipation [1]. Since a BPF covering 3.1 GHz to 10.6 GHz is one of the key components of the UWB systems, several attempts have been made to realize such filters [2]–[5]. These filters have been designed for use in single mode operation for driving coaxial systems.

However, there is a need for filters designed for differential mode operation to realize compact RF modules. There are two reasons why. First of all, most high-speed logic circuits in LSIs operate in differential mode. Secondly, miniaturized self-complementary antennas have shown to exhibit excellent ultra-wideband characteristics covering UWB applications [6], where differential mode signals must be fed according to their principles of operation. Therefore, designing other RF passive components, such as filters and impedance transformers, for use in differential mode, would eliminate the need for mode transfer circuits. This would also result in a simple RF module structure without baluns.

While a few microwave differential mode filters have been reported [7], [8], the bandwidth was too narrow for UWB applications. Even though the impedance transformer composed of two coupled lines shows modest differential bandpass performance [6], its maximum rejection out of the pass-band is limited because the rejection is determined by the impedance ratio of the transformer. To achieve fully differential mode operation as well as deep rejection demanded by UWB applications, the authors have proposed a concept of ultra-wideband antenna systems using the BPFs with four coupled lines embedded in self-complementary antennas [9].

This paper describes the design considerations, details of the differential-mode BPFs with four coupled lines fabricated on low-cost resin circuit boards, and their application to a UWB antenna system.

2. Simplified Expressions for Symmetric Four Coupled Lines

To understand the basic characteristics of the four coupled lines, two kinds of broadside-coupled and edge-coupled four-lines structures shown in Fig. 1 were investigated. In both structures, lines were laid out horizontally symmetric for a horizontally odd mode to be used as the characteristic mode, which is important for differential mode operation.

The telegrapher’s equations for the four coupled lines are expressed as follows:

\[
\frac{dV_1}{dz} = j\omega L_{11} I_1 + j\omega M_{12} I_2 + j\omega M_{13} I_3 + j\omega M_{14} I_4
\]

(1)

\[
\frac{dV_2}{dz} = j\omega M_{12} I_1 + j\omega L_{22} I_2 + j\omega M_{23} I_3 + j\omega M_{24} I_4
\]

(2)

Fig. 1 Schematic cross-sections of the four coupled lines.
(a) Vertically and horizontally symmetric structure.
(b) Horizontally symmetric structure.
W and S indicate the width and space of the lines, respectively.
When the four coupled lines are horizontally symmetric, we mode operation, and to another set of four equations for differential mode. Moreover, vertical characteristic modes for the horizontally odd mode are expressed by equations (9)–(12) above.

These telegrapher’s equations can be reduced to equations (13)–(16) below that are identical in terms of form to the equations for the asymmetric coupled two lines [10].

This implies that the characteristic modes for the horizontally symmetric four coupled lines consist of a horizontally even mode and a horizontally odd mode. Moreover, vertical characteristic modes for the horizontally odd mode are expressed by equations (9)–(12) above.

\[
\begin{align*}
\frac{dV_1}{dz} &= j\omega M_{11} I_1 + j\omega M_{12} I_2 + j\omega M_{31} V_3 + j\omega M_{41} V_4 \\
\frac{dV_2}{dz} &= j\omega M_{12} I_1 + j\omega M_{22} I_2 + j\omega M_{32} V_3 + j\omega M_{42} V_4 \\
\frac{dI_1}{dz} &= j\omega (C_{11} + C_{21} + C_{22} + C_{12}) V_1 - j\omega C_{12} V_2 \\
\frac{dI_2}{dz} &= j\omega (C_{12} + C_{22} + C_{23} + C_{24}) V_2 - j\omega C_{24} V_4 \\
\frac{dI_3}{dz} &= -j\omega C_{13} V_1 - j\omega C_{23} V_2 \\
\frac{dI_4}{dz} &= -j\omega C_{14} V_1 - j\omega C_{24} V_3 + j\omega (C_{14} + C_{24} + C_{34} + C_{44}) V_4 \\
\end{align*}
\]

where, 
\[L_{ii} = 1, 2, 3, 4; \text{inductance of line } i\]
\[M_{ij} = 1, 2, 3, 4; \text{mutual inductance between lines } i \text{ and } j\]
\[C_{ii} = 1, 2, 3, 4; \text{capacitance between line } i \text{ and ground}\]
\[C_{ij} = 1, 2, 3, 4; \text{capacitance between lines } i \text{ and } j\]

When the four coupled lines are horizontally symmetric, we obtain the following relations:

\[L_{11} = L_{22}, L_{33} = L_{44}, C_{11} = C_{22}, C_{33} = C_{44}, \]
\[M_{14} = M_{23}, M_{13} = M_{24}, C_{14} = C_{23}, C_{13} = C_{24}\]

In this case, the set of equations (1)–(8) above is reduced to the following set of four equations for differential (odd) mode operation, and to another set of four equations for common (even) mode operation:

\[
\begin{align*}
\frac{dV_1}{dz} &= j\omega (L_{11} - M_{12}) I_1 + j\omega (M_{13} - M_{14}) I_2 \\
\frac{dV_2}{dz} &= j\omega (M_{12} - L_{22}) I_2 + j\omega (L_{33} - M_{34}) I_3 \\
\frac{dI_1}{dz} &= j\omega (C_{11} + 2C_{12} + C_{13} + C_{14}) V_1 \\
\frac{dI_2}{dz} &= j\omega (C_{12} + 2C_{24} + C_{13} + C_{14}) V_2 \\
\end{align*}
\]

By replacing the elements of the matrices, we can derive the following vertically even and odd mode characteristic impedances of the vertically symmetric four coupled lines:

\[
\begin{align*}
Z_e &= \frac{L'_{11} + M'_{12}}{C'_{11}} \\
Z_o &= \frac{L'_{11} - M'_{12}}{C'_{11} + 2C_{12}} \\
\end{align*}
\]

By replacing the elements of the matrices, we can derive the following vertically even and odd mode characteristic impedances of the vertically symmetric four coupled lines:

\[
\begin{align*}
Z_e &= \sqrt{\frac{L_{11} - M_{12} + M_{13} - M_{14}}{C_{11} + 2(C_{12} + C_{14})}} \\
Z_o &= \sqrt{\frac{L_{11} - M_{12} - M_{13} + M_{14}}{C_{11} + 2(C_{12} + C_{14})}} \\
\end{align*}
\]

When the ground is far from the filter, \(C_{11}\) is very small compared to the other capacitances. Since \(C_{13}\) is the broadside capacitance, \(C_{13}\) can be much larger than \(C_{12}\) and \(C_{14}\) for the structure with a very thin substrate. Therefore, \(Z_o\) can be much larger than \(Z_e\), as confirmed below for practical structures through simulation and measurement as follows:

Parameters of the more common vertically asymmetric structure shown in Fig. 1(b), were determined through simulation using SONNET™, A thickness as thin as 0.073 mm
was selected for upper material ($\varepsilon_r = 3.7\tan\delta = 0.002$) and 0.2 mm for lower material ($\varepsilon_r = 3.5\tan\delta = 0.002$). Fig. 2(a) shows the 4-port configuration. Table 1 lists the vertically $c$- and $\pi$-mode parameters determined by fitting the simulated $Z$-parameters, and using the analytical expression that is identical in terms of form to the equations used for the symmetric coupled two lines. For practical line widths and line spaces, $Z_c$ is more than 10 times greater than $Z_e$. The effective relative permittivities of the structures in $\pi$-mode are almost the same (3.4) and roughly equivalent to the relative permittivity of the substrate (3.7), which is advantageous in reducing the $\lambda/4$ line length [12], [13]. In the vertically symmetric structures shown in Fig. 1(a), the $c$-mode becomes the even-mode, the $\pi$-mode becomes the odd-mode, $R_c$ is 1, and $R_e$ is $-1$. In this case, $Z_e$ is much higher than $Z_c$.

### 3. Basic Characteristics of Single-Stage BPF with Four Coupled Lines

Since the characteristics of these four coupled lines in differential mode can be expressed using the same equations as for the coupled two lines in single mode, these four coupled lines may be utilized for differential mode interdigital BPFs [11]. In the configuration where the opposite ends of lines for the ports are set open as shown in Fig. 2(b) for the vertically symmetric structure, $Z$-parameters for the horizontally odd mode can be expressed using the same equations as for the symmetric coupled two lines, where $\beta_c$ and $\beta_o$ denote even and odd mode phase constants as follows:

$$Z_{11} = Z_{22} = -jZ_c \cot(\beta_c L) + Z_o \cot(\beta_o L)$$  \hspace{1cm} (21)$$

$$Z_{12} = Z_{21} = -jZ_c \csc(\beta_c L) - Z_o \csc(\beta_o L)$$  \hspace{1cm} (22)$$

Supposing that port impedances are set to be $Z_{in}$ (port 1) and $Z_{out}$ (port 2), S-parameters for the differential signal can be derived as shown in equations (23)–(25) below. With the analytical expressions above, the characteristics can be approximated for the following three regions, under the assumption that $Z_e/Z_c \ll 1$ and both $Z_{in}$ and $Z_{out}$ are equal to $Z_c/2$ as follows:

[1] Region I: $\beta_c L \ll Z_o/Z_c$ and $\beta_o L \ll Z_o/Z_c$

Considering that $\csc(\beta_c L) \gg Z_e/Z_o$, $\csc(\beta_o L) \gg Z_e/Z_o$, $S_{21}$ approaches 0 and the input is almost totally rejected in this region. Moreover, this region is a narrow band because

$$S_{11} = \left\{ \frac{Z_o}{Z_c} \right\}^2 + 4 \left( \frac{Z_{in} Z_{out}}{Z_c^2} \right) - 2 \frac{Z_o}{Z_c} \csc(\beta_c L) \csc(\beta_o L) [\cos(\beta_c L) \cos(\beta_o L) + 1] + 2j \left( \frac{Z_{in} - Z_{out}}{Z_c} \right) \left\{ \frac{Z_o}{Z_c} \cot(\beta_o L) + \cot(\beta_c L) \right\}$$  \hspace{1cm} (23)$$

$$S_{22} = \left\{ \frac{Z_o}{Z_c} \right\}^2 + 4 \left( \frac{Z_{in} Z_{out}}{Z_c^2} \right) - 2 \frac{Z_o}{Z_c} \csc(\beta_c L) \csc(\beta_o L) [\cos(\beta_c L) \cos(\beta_o L) + 1] - 2j \left( \frac{Z_{in} + Z_{out}}{Z_c} \right) \left\{ \frac{Z_o}{Z_c} \cot(\beta_o L) + \cot(\beta_c L) \right\}$$  \hspace{1cm} (24)$$

$$S_{21} = S_{12} = -4j \frac{Z_{in} Z_{out}}{Z_c^2} \left\{ \csc(\beta_c L) - \frac{Z_o}{Z_c} \csc(\beta_o L) \right\}$$  \hspace{1cm} (25)$$
Z_o/Z_e is very small.

[2] Region II: |β_o L − π| < Z_o/Z_e

Since the effective relative permittivity of the odd mode is larger than that of the even mode, β_o L is larger than β_e L. Thus, the frequency where β_o L reaches π is lower than that where β_e L reaches π. Where β_o L is approximately π, S_{21} can be approximated as follows, where ε_{ee} and ε_{eo} denote the effective relative permittivity for both even and odd modes:

\[ S_{21} \approx \cos \left( \frac{\beta_e L}{2} \right) \exp \left( -\frac{\beta_e L}{2} \right) \]  
\[ \beta_e L = \sqrt{\frac{\varepsilon_{ee}}{\varepsilon_{eo}}} \pi \tag{27} \]

Since ε_{ee} is somewhat smaller than ε_{eo} as shown Table 1, |S_{21}| becomes small but does not reach 0. This implies that even maximum rejection remains finite at the upper rejection band. Moreover, the region also is a narrow band because Z_o/Z_e is small.


Where neither csc(β_o L) nor csc(β_e L) is very large, S_{11} can be expressed as follows:

\[ S_{11} \approx 0 \left( \frac{Z_o}{Z_e} \right) \tag{28} \]

0 \left( \frac{Z_o}{Z_e} \right) is a small with (Z_o/Z_e)

Equation (28) implies that all of region III becomes a pass band. Since the bandwidths for region I and region II are narrow, this BPF with four coupled lines can be expected to exhibit very wideband band-pass characteristics.

Given the above analytical results, wideband single-stage BPFs of the slightly asymmetric and symmetric structures shown in Fig. 3, were designed and fabricated. To accommodate the multi-layer printed circuit board (PCB) process as well as to reduce dielectric loss, Panasonic Megtron6™, a resin PCB material, was selected for the substrate [14]. For the slightly asymmetric structure shown in Fig. 1(b), thickness of the upper material (\(\varepsilon_r = 3.7 \tan \delta = 0.002\)) was 0.073 mm and thickness of the lower material (\(\varepsilon_r = 3.5 \tan \delta = 0.002\)) was 0.2 mm, which is the same substrate structure shown in Sect.2. For the symmetric structure shown in Fig. 1(a), thickness of the substrate (\(\varepsilon_r = 3.7 \tan \delta = 0.002\)) was 0.1 mm. Fig. 4 shows the simulated and measured fractional bandwidths and port impedances with which return loss is minimized at the center frequency. Ultrawide bandpass characteristics were obtained not only for single-stage BPFs of the symmetric structure but also for those of the slightly asymmetric structure. Both the wider line width and wider line space cause a larger fractional bandwidth. The maximum fractional bandwidth was 170% for the slightly asymmetric BPF for which line width was 0.7 mm and line space was 0.8 mm. In terms of fractional bandwidth, both simulated values and measured values are in good agreement. However, in terms of port impedance, the measured values are lower than simulated values. This is due to the infinitesimal metal thickness employed in the moment method simulation. Fig. 5 shows the band-pass characteristics of the slightly asymmetric single-stage BPF using four coupled lines that are 6 mm long, 0.6 mm wide, and spaced 0.4 mm apart. The measured bandwidth with 1-dB insertion loss was 10.3 GHz between 1.5 GHz to 11.8 GHz, and covered the entire UWB band.

When the four coupled lines are used for an impedance transformer, the input and output port impedances become different. From Eq. (23) the product of Z_{in} and Z_{out} was kept constant (Z_o^2/4) to minimize return loss, with Z_o varied to estimate the effect of port impedance on a fractional bandwidth. In this case, the measured fractional bandwidth was mainly reduced due to the increase in insertion loss in the lower band, as shown in Fig. 6.

This can be considered consistent with Eq. (23), which
can be approximated as follows, where neither $\csc(\beta_e L)$ nor $\csc(\beta_o L)$ is very large, and $Z_o/Z_e$ is very small:

$$|S_{11}| \approx \frac{|j(Z_{in} - Z_{out})\cot(\beta_e L)/Z_e|}{1 + j(Z_{in} + Z_{out})\cot(\beta_e L)/Z_e} \approx \frac{|Z_{in} - Z_{out}|}{Z_e} |\cos(\beta_e L)|$$  \hspace{1cm} (29)

$\cos (\beta_e L)$ is larger at the lower-band edge than that at the higher-band edge because $\cos(\beta_e L)$ approaches $\cos(\pi \sqrt{\varepsilon_e/\varepsilon_o})$ at the higher-band edge. This value is not large when considering the effective relative permittivity shown in Table 1. After all, with less than a $\pm 20\%$ change in $Z_{out}/Z_{in}$, the fractional bandwidth is kept almost identical as shown in Fig. 6(b). Therefore, an impedance ratio ($Z_{out}/Z_{in}$) of less than 1.2 per stage could be achieved without affecting the fractional bandwidth.

![Fig. 5 Measured and simulated band-pass characteristics. The length of the four coupled lines is 6 mm. $Z_{in} = Z_{out} = 123\, \Omega$.](image)

![Fig. 6 Measured effect of port impedance on fractional bandwidth.](image)

### 4. Multi-Stage BPF for the UWB System

Vertically symmetric 6-stage BPFs with impedance transformation from 50 $\Omega$ to 172 $\Omega$ were designed and fabricated for impedance-matching the constant input impedance (172 $\Omega$) of the self-complementary antenna laid out on a substrate of high permittivity. The 0.1-mm thick resin material of high permittivity ($\varepsilon_r = 9.8,\tan\delta = 0.002$) was used to miniaturize the filter as well as to realize four coupled lines filters with lower matched port impedances. To minimize the size, 6-stage BPF was adopted, of which impedance ratio was a little bit higher than 1.2. Moreover, matched port impedances of the filters were increased to some extent, because bandwidth was affected less when the matched port

<table>
<thead>
<tr>
<th>Stage</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
</tr>
</thead>
<tbody>
<tr>
<td>$W$ (mm)</td>
<td>1.0</td>
<td>1.0</td>
<td>1.0</td>
<td>1.0</td>
<td>0.6</td>
<td>0.6</td>
</tr>
<tr>
<td>$S$ (mm)</td>
<td>0.1</td>
<td>0.2</td>
<td>0.4</td>
<td>0.5</td>
<td>0.6</td>
<td>0.8</td>
</tr>
<tr>
<td>$L$ (mm)</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
<td>4.0</td>
</tr>
<tr>
<td>$Z$ ($\Omega$)</td>
<td>62</td>
<td>78</td>
<td>100</td>
<td>120</td>
<td>140</td>
<td>160</td>
</tr>
</tbody>
</table>

$W$: line width, $S$: line space, $L$: line length, $Z$: matched port impedance

Table 2 4-coupled lines parameters of the 6-stage BPF.
impedance is higher than the port impedance. Since the matched port impedance is dependent on both line width and line space, wider line width up to 1 mm was selected to obtain wider bandwidth as shown in Fig. 4. Selected four coupled lines’ parameters are shown in Table 2.

In addition, differential-mode low-pass filters (LPFs) included with the BPF for rejecting the upper band beyond the UWB band, and differential-mode notch filters included with the BPF for rejecting the 5-GHz wireless LAN band, were also designed and fabricated as shown in Fig. 7. Figure 8 shows the schematic of the LPF composed of semi-lumped inductor and capacitors. Two coupled lines of low characteristic impedance with a united backside plate were used for a differential shunt capacitor, and two coupled lines of high impedance were used for a differential inductor. The capacitance and inductance values were determined according to the 3-stage Chebyshev LPF formulae for 0.3-dB ripple and 12-GHz cutoff frequency.

Figure 9 shows the measured characteristics of the fabricated filters. Port impedances of the filters were set to 50 Ω and 172 Ω. Subtracting the pad T-parameters determined through electromagnetic simulation eliminated the parasitic contributions of measuring pads used for the air-coplanar probes.

Insertion loss of the BPF was less than 1 dB from 3.5 GHz to 11.0 GHz, and less than 2.2 dB over the UWB band (3.1 GHz to 10.6 GHz). The 3-stage notch filter eliminated the 5 GHz to 6.5 GHz pass band, and affected the pass band characteristics by less than 1 dB over the other UWB band. The LPF eliminated the next pass band beyond 12 GHz. The maximum insertion loss was 2.4 dB for the BPF with a notch filter, and 2.6 dB for the BPF with a LPF over the UWB band except the intentionally rejected 5-GHz band.

5. Embedded BPF Performance in the Antenna System

The three filters were embedded in self-complementary antennas fabricated on the same 0.1-mm thick material of high relative permittivity. Fig. 10 shows a fabricated self-complementary antenna embedded with the notch filter and the 6-stage BPF. Figure 11 shows the measured return loss of three integrated antennas and an unmatched unit antenna. Return loss was clearly improved by the BPF as compared with an unmatched unit antenna.

These antennas are designed to operate in the horizontally differential or odd mode, and suppress the horizontally even mode. To eliminate even mode excitation of the antenna, the filter’s even mode impedance is set very high with matching odd mode impedance also set. In this case, the filter can be expected to function as a balun. To quantitatively estimate balun performance, said performance was simulated with the 6-stage BPF as shown in Fig. 12, where one of the two differentially 50 Ω ports was shorted and the other port terminated at 50 Ω.

Mixed mode balun characteristics can be obtained us-

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**Fig. 7** Schematic of the differential-mode LPF.

**Fig. 8** Measured characteristics of the fabricated filters.

**Fig. 10** Photograph of the antenna with notch filters and the BPF.
S_{ss11}: return loss of unbalanced port (port 1)
S_{dd22}: differential mode return loss of balanced ports (port 2, port 3), stimulated by differential mode signal
S_{ds21}: differential mode insertion loss, where unbalanced port is stimulated by single mode signal
S_{cs21}: common mode insertion loss, where unbalanced port is stimulated by single mode signal

Despite the single mode stimulus of port 1, the differential output is dominant at the balanced port as shown in Fig. 12. The common mode output \( (S_{cs21}) \) is less than \(-9.2\) dB from 3.1 GHz to 10.6 GHz, and the differential mode output \( (S_{ds21}) \) is more than \(-2.0\) dB. Figure 13 compares the return loss and insertion loss between balanced and unbalanced stimulus at port 1. Since the insertion loss and return loss are almost equivalent, the four coupled lines can be considered to work properly as a balun.

Since the four coupled lines were confirmed to work properly for single mode stimulus, radiation patterns were estimated using the antenna system connected via a coaxial cable. Radiation patterns at 3.1 GHz, 5.7 GHz, and 10.6 GHz were estimated with the antenna system as shown in Fig. 14. On the xz-plane, radiation patterns were similar to the dipole antenna’s E-plane patterns, and on the yz-plane, radiation patterns were similar to the dipole antenna’s H-plane patterns. Maximum gain was 1.4 dBi at 3.1 GHz, 1.6 dBi at 5.7 GHz, and \(-2.1\) dBi at 10.6 GHz. Figure 15 compares \( S_{21} \) of the antenna with and without notch filters. \( S_{21} \) of the antenna with notch filters was suppressed from 5 GHz to 6.5 GHz, which agreed with the return loss characteristics shown in Fig. 9 and Fig. 11.

6. Conclusion

Design considerations for an ultra-wideband BPF with four coupled lines were described. Impedance-matched self-complementary antennas integrated with an ultra-wideband differential mode BPF with four coupled lines, a notch filter, and a LPF were demonstrated. Broadside-coupled and edge-coupled four-lines BPFs were shown analytically and experimentally to exhibit broadband performance and an impedance transformation ratio of 1.2 per stage with little
bandwidth reduction. Thus, 6-stage BPFs were utilized to transform the self-complementary antenna’s constant input impedance \((60\pi \varepsilon^{-1/2}(\Omega))\) to 50 \(\Omega\). In addition, low-pass filters included with the BPF and notch filters included with the BPF were designed and fabricated. The measured insertion loss of the filter systems was less than 2.6 dB from 3.1 GHz to 10.6 GHz. The filters were successfully embedded in wideband self-complementary antennas to reject unnecessary radiation over the next pass band and 5-GHz wireless LAN band.

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References


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