Concurrent tri-band power amplifier based on novel tri-band impedance transformer

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Abstract: In this paper, we present a concurrent tri-band power amplifier by using a tri-band matching network synthesis method. At the package plane, a cross-type impedance transformer and susceptance matching network are placed in parallel to realize the desired admittance at three independent frequencies. Design equations of the different possible topologies are present based on ABCD-parameters. Then, the proposed synthesis approach is validated with the design of a tri-band power amplifier at frequency bands 0.90, 1.85 and 2.40 GHz. For a continuous-wave measurement, the peak drain efficiency is 70.0%, 67.1% and 59.8% at 0.90, 1.85 and 2.40 GHz, respectively.

Keywords: tri-band power amplifier, tri-band matching networks, power amplifier (PA)

Classification: Microwave and millimeter-wave devices, circuits, and modules

References

1 Introduction

Modern wireless communication systems use sophisticated modulation schemes and architecture to achieve high data rates, which would cause large signal bandwidth, increasing frequency bands and modes. Power amplifier (PA), as one of the critical components in the transmitter, tend to support multi-band/multi-mode operation. As such, transmitter calls for wideband/multi-band PA with high efficiency.

Therefore, multi-band PA, which could offer optimized performance for those desired bands, is also an attractive candidate for the communication system. Some multi-band designs are realized by using several narrow-band power amplifiers in previous works. In this case, switching is required when the operation frequency changes. This method can not amplifier multi-band signal simultaneously when the carrier aggregation is employed. As a result, concurrent dual-band/tri-band PAs are preferred. Several multi-band PAs consider the load matching network as a whole [1, 2, 3, 4, 5, 6, 7]. In these works, the performances at the desired frequency bands can be optimized to state-of-the-art by using fundamental or harmonic tuning. Nonetheless, in some of these works, one of the drawbacks is that the supersession of out-of-band output power is not good enough, due to the continuity of passive network and transistor they use.

Other works [8, 9] utilize dual-characteristic impedance transformers and dual-band susceptance T-type stub to construct dual-band admittance matching. The dual-band impedance transformer act as a quart-wave line with different characteristic impedance to transform 50 Ohm to two different real impedances at two irrelevant frequency bands. Consequently, equivalent quart-wave line would bring high load impedance between two frequency bands. In this way, the out-of-band output power could be suppressed, while avoiding the detrimental spur owing to the nonlinearity in concurrent mode.

Based on the previously published paper [8, 9], this paper presents a concurrent tri-band power amplifier, which utilizes the modified tri-band impedance transformer approach [10]. The modified impedance transformer represents three differ-
ent conductances when terminated by a 50 Ohm load. And a tri-band susceptance matching with stubs is also involved in this work.

2 Tri-band matching method

To construct a matching network for three bands, the topology as shown in Fig. 1(a) is proposed. With the same bias condition, the transistor could normally have different optimum impedance for each band. [1, 2, 3, 4, 5, 6, 7] Directly transform 50 Ohm to three complex impedances (i.e., $Z_1$, $Z_2$ and $Z_3$) simultaneously, with a high order of numerical optimisations. In this work, we turn the impedance matching to admittance matching, and two parallel networks are employed to meeting the requirements for each part of admittance (i.e., $Y_1$, $Y_2$ and $Y_3$) (show in Fig. 1(b)). The tri-band impedance transformer is utilized to represent the desired conductances (i.e., $G_1$, $G_2$ and $G_3$) (red arrows in the smith chart) for each band while the open/short shunt stubs are used for realizing the remaining susceptances (i.e., $B_1$, $B_2$ and $B_3$) (blue arrows in the smith chart).

The key element of this matching network is the tri-band impedance transformer. The “equivalent quarter-wave line” type transformer, which would generate the only real part contained impedance, can be used for the conductance matching. However, it is too complex to construct a network that acts as the quarter-wave line at three bands. Then, we employ a modified dual-band quarter-wave impedance transformer for the first two bands ($f_1$ and $f_2$), and the network can be tuned for meeting the admittance at the last band ($f_3$). Meanwhile, the open/short shunt stubs, which can only generate the imaginary part of impedance, are utilized for the susceptance matching.

2.1 Tri-band impedance transformer

For a quarter-wave transmission line with characteristic impedance $Z_0$ loaded by $Z_{load}$, the input impedance is give by

$$Z_{in} = \frac{Z_0^2}{Z_{load}}.$$  \hspace{1cm} (1)

At the first ($f_1$) and second ($f_2$) frequencies, the desired impedances are $Z_1$ and $Z_2$. Then, the characteristic impedances of the equivalent quarter-wave line should be

$$Z_{01} = \sqrt{\left(\frac{Z_{load}}{G_1}\right)} \text{ where } G_1 = real\left(\frac{1}{Z_1}\right)$$ \hspace{1cm} (2)

$$Z_{02} = \sqrt{\left(\frac{Z_{load}}{G_2}\right)} \text{ where } G_2 = real\left(\frac{1}{Z_2}\right)$$

The proposed impedance transformer is shown in Fig. 2. This is a modified T-type structure with two shunt stubs with the total susceptance of $jB$, connected to two same transmission lines with electrical length of $\theta$, and characteristic impedance of $Z_0$. The $ABCD$-parameters of the proposed structure shown in Fig. 2 can thus be derived as
which leads to

\[
\begin{bmatrix}
\cos \theta_s & jZ_s \sin \theta_s \\
\frac{1}{Z_s} \sin \theta_s & \cos \theta_s
\end{bmatrix}
\begin{bmatrix}
1 & 0 \\
jB & 1
\end{bmatrix}
\begin{bmatrix}
\cos \theta_s & jZ_s \sin \theta_s \\
\frac{1}{Z_s} \sin \theta_s & \cos \theta_s
\end{bmatrix} = \begin{bmatrix}
\cos \theta_s & jZ_s \sin \theta_s \\
\frac{1}{Z_s} \sin \theta_s & \cos \theta_s
\end{bmatrix},
\]

which leads to
\[
\begin{bmatrix}
\cos 2\theta_s - \frac{1}{2}BZ_s \sin 2\theta_s & jZ_s (\sin 2\theta_s - BZ_s \sin^2 \theta_s) \\
\frac{1}{Z_s} (\sin 2\theta_s + BZ_s \cos^2 \theta_s) & \cos 2\theta_s - \frac{1}{2}BZ_s \sin 2\theta_s
\end{bmatrix}.
\]

Meanwhile, the \(ABCD\)-parameters of quarter-wave or the equivalent can be written as
\[
\begin{bmatrix}
0 & \pm jZ_0 \\
\pm j\frac{1}{Z_0} & 0
\end{bmatrix}.
\]

Comparing (4) to (5), the necessary condition for dual-band impedance transformer can be stated as
\[B = \frac{2}{Z_0 \tan 2\theta_s},\]
\[Z_{01} = \pm Z_s \tan \theta_{s1} \quad \text{for} \quad f_1,
\[Z_{02} = \pm Z_s \tan \theta_{s2} \quad \text{for} \quad f_2.\]

Equation (7) can be solved simultaneously to get the parameters \(Z_s\) and \(\theta_s\). Once the values of \(Z_s\) and \(\theta_s\) are known, using equation (6), the corresponding parameters \((B_1\) and \(B_2\)) for shunt stubs can be obtained.

For a transmission line with electrical length of \(\theta\) and characteristic impedance of \(Z\), the input impedance is
\[Z_{\text{in}} = Z_0 \frac{Z_{\text{load}} + jZ \tan \theta}{Z + jZ_{\text{load}} \tan \theta}.\]

The impedance transformer network can also be seen as a transmission line, and its electrical length \(\theta\) and characteristic impedance \(Z\) are given as
\[\cos \theta = \cos 2\theta_s - \frac{1}{2}BZ_s \sin 2\theta_s,
\[Z = Z_s \sqrt{\frac{\sin 2\theta_s - BZ_s \sin^2 \theta_s}{\sin 2\theta_s + BZ_s \cos^2 \theta_s}}.\]

Substituting (9) into (8), then we get the input impedance of the third band \((f_3)\). Moreover, with the requirement of triple band matching, the real part \((G_{\text{in}})\) of the input admittance \((Y_{\text{in}})\) should be equal to \(G_3\). Subsequently, for the third band, the parameter of shunt stubs \(B_3\) can be calculated via numerical tools.

Finally, all the parameters of shunt stubs for three bands can be get from above. As it is shown in Fig. 2, there are three types of shunt stubs based on weather the stub is realized by open/short transmission line. For each case, the condition of design parameters can be written as

- In the case of that all the stubs are short
\[
\begin{bmatrix}
\tan \theta_{p11} & \tan \theta_{p12} \\
\tan \theta_{p21} & \tan \theta_{p22} \\
\tan \theta_{p31} & \tan \theta_{p32}
\end{bmatrix}
\begin{bmatrix}
Z_{p1} \\
Z_{p2}
\end{bmatrix} =
\begin{bmatrix}
1/B_1 \\
1/B_2 \\
1/B_3
\end{bmatrix}
\]
In the case of that all the stubs are open

\[
\begin{bmatrix}
\tan \theta_{p11} & \tan \theta_{p12} \\
\tan \theta_{p21} & \tan \theta_{p22} \\
\tan \theta_{p31} & \tan \theta_{p32}
\end{bmatrix}
\begin{bmatrix}
1/Z_{p1} \\
1/Z_{p2}
\end{bmatrix} =
\begin{bmatrix}
B1 \\
B2 \\
B3'
\end{bmatrix}
\] (11)

In the case of one short and one open

\[
\begin{bmatrix}
\tan \theta_{p11} & 1/\tan \theta_{p12} \\
\tan \theta_{p21} & 1/\tan \theta_{p22} \\
\tan \theta_{p31} & 1/\tan \theta_{p32}
\end{bmatrix}
\begin{bmatrix}
Z_{p1} \\
Z_{p2}
\end{bmatrix} =
\begin{bmatrix}
1/B1 \\
1/B2 \\
1/B3'
\end{bmatrix}
\] (12)

where \( Z_{p1} \) and \( Z_{p2} \) are the characteristic impedances of two stubs, \( \theta_{p11} \) and \( \theta_{p22} \) are the electrical length of them at the frequency \( f_x \). There are four variables and three equations, the characteristic impedances and electrical length can be also calculated via numerical tools.

It is worthy to mention that the designed impedance transformer also generate the imaginary part of impedance at \( f_3 \). So the desired value (\( jB_0 \)) of susceptance matching should be the result of subtracting between \( \text{imag}(1/Z_3) \) and \( jB_3' \) generated by the transformer.

The \( \pi \)-type topology can also be employed for tri-band impedance transformer. However, to realize a triple band matching, the \( \pi \)-type network need to be modified by adding two additional stubs. The schematic draw shown in Fig. 1 shows that it has to put three or more stubs at one junction, which is not a convenient for circuit design. So the the \( \pi \)-type topology would not be discussed here.

### 2.2 Tri-band susceptance matching

As it mentioned above, the short/open shunt stubs could only generate the imaginary part of impedance when the transmission line is lossless. The tri-band susceptance matching network could be constructed by two or three stubs. The topology of matching network is shown in Fig. 3. In this paper, we denote that \( f_1 \) is the reference frequency, and \( n_2 = f_2/f_1 \), \( n_3 = f_3/f_1 \).

Table I. Design parameters for triple band matching network.

<table>
<thead>
<tr>
<th>Characteristic impedances ( Z ) and electrical length ( \theta ) (Ohm/degree@0.9 GHz)</th>
<th>( Z_s )</th>
<th>( Z_{p1} )</th>
<th>( Z_{p2} )</th>
<th>( Z_{T1} )</th>
<th>( Z_{T2} )</th>
<th>( Z_{T3} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>30.0</td>
<td>109.6</td>
<td>109.6</td>
<td>70.0</td>
<td>42.4</td>
<td>30.0</td>
<td></td>
</tr>
<tr>
<td>( \theta_s )</td>
<td>( \theta_{p1} )</td>
<td>( \theta_{p2} )</td>
<td>( \theta_{T1} )</td>
<td>( \theta_{T2} )</td>
<td>( \theta_{T3} )</td>
<td></td>
</tr>
<tr>
<td>89.6</td>
<td>52.8</td>
<td>52.8</td>
<td>68.5</td>
<td>78.6</td>
<td>90</td>
<td></td>
</tr>
</tbody>
</table>

In Fig. 3, it is not hard to do tri-band susceptance matching since there are six variables and only three equations for designing. To make the procedure more brief and clear, we denote that \( Y_a = j0@f_2 \). It could be realized by a quarter-wave open transmission line or a half-wave short transmission line of \( f_2 \). Normally, the quarter-wave transmission line is preferred for the smaller size of circuits.
Then, after a rational characteristic impedance of Line 1 ($Z_{T1}$) is selected and fixed, the electrical length of Line 1 ($\theta_{T1}$) can be expressed as

$$\theta_{T1} = \frac{1}{n_2} \tan^{-1}\left(\frac{-1}{Z_{T1}B_2}\right). \tag{13}$$

Once $Z_{T1}$ and $\theta_{T1}$ is fixed, the desired susceptance of $f_1$ at the junction should be

$$Y_a + Y_b = -j(B_1 Z_{T1} + \tan \theta_{T1}) \quad \frac{Z_{T1} + Z_{T1}^* B_1 \tan \theta_{T1}}{Z_{T1}^2 + B_1^* \tan \theta_{T1}}. \tag{14}$$

For $f_3$, the equation (14) is rewritten as
Ya + Yb = \frac{-j(B'_3Z_{T1} + \tan(n_3\theta_{T1}))}{Z_{T1} + Z'_{T1}B'_3\tan(n_3\theta_{T1})}.

(15)

The susceptance stubs could be reused as the bias line in power amplifier designing. Moreover, it is more convenient and effective to de-coupling RF signal by capacitance rather than inductor. So the design parameters of Line 2 and Line 3 can be obtained from

\[
Y_a + Y_b = \frac{j\tan(\theta_{T2})}{Z_{T2}} + \frac{-1}{Z_{T3}\tan(\theta_{T3})} \text{ for } f_1,
\]

or

\[
\frac{j\tan(n_3\theta_{T2})}{Z_{T2}} + \frac{-1}{Z_{T3}\tan(n_3\theta_{T3})} \text{ for } f_3,
\]

(16)

where \(Z_{T2}\), \(Z_{T3}\) and \(\theta_{T3}\) are the unknown variables, and \(Z_{T2}\) can be tuned for the optimum size of circuits without the influence on the matching at \(f_2\).

Above all, to do the tri-band matching for three arbitrary frequencies with different impedances, the steps are summed here:

- Covert the impedance matching to susceptance matching by employing two networks,
- Design the dual-band impedance transformer with (7) and then complete tri-band conductance matching by using (10)–(12),
- Update the desired susceptance, select one frequency for the first susceptance matching and fix the first transmission line, then complete the susceptance matching with (14)–(16).

Fig. 5. Load impedance of the output matching network.

\[
Y_a + Y_b = \frac{-j(B'_3Z_{T1} + \tan(n_3\theta_{T1}))}{Z_{T1} + Z'_{T1}B'_3\tan(n_3\theta_{T1})}.
\]

Fig. 6. Fabricated circuit of the designed tri-band PA.
3 Typographical style

To verify this matching method, a design example of tri-band PA at 0.90 \( (f_1) \), 1.85 \( (f_2) \) and 2.40 GHz \( (f_3) \) is given here. The optimal load impedances are obtained from the load-pull simulations with the GaN HEMT CGH40010F, with the bias voltage and current of 28 V and 150 mA.

Then, the method is validated by designing the output matching networks (Fig. 4). Moreover, the load impedances are too sensitive to the frequency if we directly utilize the tri-band matching method to match transistor to 50 Ohm (see Fig. 5(a)). As such, an additional matching network is inserted between the tri-band matching network and the load (50 Ohm). Thus, the load impedances \( (f_1 \) and \( f_2 \) of the tri-band matching network get smaller (36 Ohm) while the load impedance at \( f_3 \) is changed to 66.5 + j * 11.2 Ohm (green arrows in Fig. 4(b)). We choose 0.90 GHz and 1.85 GHz as the first two frequencies for impedance transformer design, and the two shunt stubs are all set to be open-circuit line. For the susceptance matching, the quarter-wave open-circuit line is set at 0.90 GHz, and the left stub is designed as shorted. The detailed design parameters are all obtained via numerical calculation by Matlab, as it listed in Table I. The load impedances with the additional matching network, which are less sensitive, are shown in Fig. 5(b). The high resistances can be used for out-of-band power suppression.

The proposed PA is realized on the 20 mil ROGERS 4350B substrate with \( \varepsilon = 3.66 \). Fig. 6 shows the fabricated circuit of the proposed dual-band PA. Continuous wave (CW) signal measurements are firstly performed to test the proposed tri-band PA’s performance (shown in Fig. 7). Maximum drain efficiency (DE) of 70.0%, 67.1% and 59.8% are measured with output power of 40.0, 41.5 and 40.3 dBm at 0.90, 1.85 and 2.40 GHz, respectively. As it shown in Fig. 7(a), the out-of-band power suppression is above 20 dB. Moreover, the measured CW performance in balance concurrent mode [2] is also shown in Fig. 7.

To evaluate the PA’s performance for modulated signal applications, a 20 MHz 64-QAM signal with 6.8 dB Peak–to–Average–Power–Ratio (PAPR) is used to test the PA in the three bands. Digital pre-distortion (DPD) technique is performed to linearize the PA [11]. Fig. 8 shows the measured output spectrums and constellation. For the three bands, the measured adjacent channel leakage ratios (ACLRs)
are improved from −35 dBc to better than −50 dBc with DPD performed while the EVM are all better than −30 dB. The average output powers are 35 dBm and the average DE are 47.2%, 40.8% and 43.1% in the three bands. Table II lists the overall measured performance of the proposed PA and compares it to other similar published works.

![Fig. 8. Measured output spectrums with and without DPD at (a) 0.90, (c) 1.85 and (d) 2.40 GHz, and (b) output constellation at 0.90 GHz](image)

Table II. Comparison table with other multi-band PAs.

<table>
<thead>
<tr>
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<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Ratio(b)</td>
<td>1.37</td>
<td>2.37</td>
<td>1.53</td>
<td>1.36</td>
<td>2.5</td>
<td>1.33</td>
<td>2.66</td>
</tr>
<tr>
<td>Pout, max (dBm)</td>
<td>41.5/41.8</td>
<td>46/45</td>
<td>41/41</td>
<td>41.1/40.8</td>
<td>39.8/40.8/39.2</td>
<td>43/43</td>
<td>40.0/41.5/40.3</td>
</tr>
<tr>
<td>Out-of-Band Suppression, max (dB)</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>N/A</td>
<td>35(a)</td>
<td>38</td>
</tr>
<tr>
<td>DE, max (%)</td>
<td>76/70</td>
<td>65/68</td>
<td>70/64</td>
<td>80/74</td>
<td>56.4/58.3/43.4</td>
<td>69/61</td>
<td>70.0/67.1/59.8</td>
</tr>
</tbody>
</table>

Notes:
(a) Obtained from publication figures.
(b) Frequency Ratio = (Maximum Frequency)/(Minimum Frequency).
4 Conclusion

A novel matching method has been proposed for matching three unrelated bands and impedances simultaneously. A tri-band impedance transformer and a susceptance matching stub are introduced for transforming ohmic terminal load to the desired impedance. The CW measurement results show that the proposed PA archives 70.0%, 67.1% and 59.8% efficiencies at 0.90, 1.85, 2.40 GHz, respectively. With a 20 MHz 64-QAM signal stimulation, the tri-band PA has 47.2%, 40.8% and 43.1% average DE at the three bands, with ACLRs lower than −50 dBc using DPD techniques.

Acknowledgments

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