A new compact and wideband quadrature feeding network with an integrated asymmetrical phase shifter

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Abstract

This article reports a new class of quadrature feeding network which integrates an asymmetrical phase shifter to obtain the compact size and wideband property simultaneously. By adjusting the impedance of two loading stubs, this asymmetrical phase shifter can realize 90° phase shift, impedance transformation, and phase slope alignment at the same time. Operation principle and design procedure for this introduced quadrature feeding network are provided with synthesis equations and design curves. By using the shape of Archimedes gradual-change lines, the area of this circuit layout is fully utilized, and the overall size of the proposed feeding network is only 0.05λ\(^2\). Compared with a traditional compact design, this technique can effectively enhance the amplitude balance bandwidth by 1.7 times (from 58% to 101%) and phase balance bandwidth by 5.9 times (from 14% to 83%) as well as keep the circuit size nearly unchanged under the same conditions. To verify the design concept, two prototypes of the quadrature feeding networks are designed, fabricated, and tested. Good agreement between the measured and simulated results is achieved and they well reveal the attractive wideband equal amplitudes and 90° phase balance performance for this proposed new compact quadrature feeding network.

1 | INTRODUCTION

Quadrature feeding networks are key circuit blocks in the multi-fed circularly polarized antennas, balanced power amplifiers, image rejection mixers and I/Q modulators [1–3]. Generally, the quadrature feeding networks are composed of a Wilkinson power divider and a 90° constant phase shifter in [4–9]. Among them, Wilkinson power divider provides the equal power splitting and wideband phase shifter offers 90° constant phase difference at two output ports. Thus, they are also named as Wilkinson quadrature splitters.

In [4], the Schiffman phase shifter was utilized to construct a quadrature feeding network with equal amplitudes of ±0.5 dB and constant 90° phase difference of ±10° over a wide bandwidth of 80%. In the similar manner, a variety of quadrature feeding networks have been carried out by using the loaded-line type phase shifter [5], metamaterial-line type phase shifter [6], coupled-line type phase shifter [7], and so forth. In [8,9], the high-pass network and phase correcting networks were developed to further increase the operation bandwidth. Although this kind of quadrature feeding network can achieve the wide bandwidth. One prominent drawback cannot be ignored that: the cascaded power divider and phase shifter occupy very large substrate area and result in the large insertion loss and high cost. For some of applications, such as the handheld device, the circuit size is very limited. Thus, it is demanded for the compact quadrature feeding network.

Till now, several works have been reported. In [10], a novel compact structure was presented. As the key part, the quarter-wave transmission line not only provides the desired 90° phase shift, but is also a part of power divider and acts as a quarter-wave impedance transformer to realize equal amplitude responses. This design is compact enough to be put in the middle of the annular-ring antenna. On the similar principle, several compact quadrature feeding networks have been proposed for the multi-fed circularly polarized antenna applications [11–13]. Recently, the possibility to construct reconfigurable feeding network has also been studied in [14].
Compared with the previous designs by cascading the power divider and 90° phase shifter [4–9], this kind of quadrature feeding network has advantages of compact size and low cost. Thus, it is very suitable for the size-limited and cost-sensitive applications. However, the bandwidth is much narrower than the first type. It is because the equal power splitting and 90° phase difference can only be satisfied at the centre frequency \( f_0 \). When the operation frequency deviates from the centre point, the performance will be quickly deteriorated.

In this article, we would like to focus our attention on developing a new class of quadrature feeding network which can achieve compact size and wideband property at the same time. To meet this requirement, an asymmetrical phase shifter is proposed and integrated in the quadrature feeding network, which can simultaneously provide 90° phase shift, impedance transformation, and phase slope alignment. By virtue of these useful functions in integrated phase shifter, desired wideband equal amplitudes and 90° phase difference responses can be both obtained. Compared with the traditional compact design in [10], the proposed quadrature feeding network effectively enhances the amplitude balance bandwidth by 1.7 times (from 58% to 101%) and phase balance bandwidth by 5.9 times (from 14% to 83%) as well as keep the circuit size nearly unchanged under the same conditions. In the following sections, the configuration, working principle, design method, experimental results and discussion will be given.

## 2 | DESIGN METHOD AND THEORETICAL ANALYSIS

The circuit model of the proposed quadrature feeding network is shown in Figure 1. The key element of described concept is the integrated asymmetrical phase shifter in the blue dashed box, which consists of a transmission line section with characteristic impedance \( Z_2 \) and two different shunt stubs with characteristic impedances \( Z_1 \) and \( Z_3 \). Electrical length of all these sections is 90° at the centre frequency \( f_0 \) that is, \( \theta_1(f_0) = 90° \). For the subnetwork, the source and load impedance are \( Z_S \) and \( Z_L \), respectively.

Figure 2 depicts the transmission poles \( f_{p1} \) and \( f_{p2} \) and insertion phase responses \( \angle S_{21} \) of the proposed asymmetrical phase shifter under \( Z_S = 100 \) \( \Omega \) and \( Z_L = 50 \) \( \Omega \) to reveal its basic working principle. Firstly, the phase shifter can provide magnitude response of \( |S_{21}| \leq 1 \) with good matching over the wide bandwidth. As seen from port 1 with \( Z_0 = 50 \) \( \Omega \), \( Z_S \) should be 100 \( \Omega \) for adaptation and power balance between ports 2 and 3, and the port impedance of port 2 and 3 is herein set as 50 \( \Omega \) \( (Z_L = 50 \) \( \Omega \)). In other words, it acts a good wideband impedance transformer from \( Z_S = 100 \) \( \Omega \) to \( Z_L = 50 \) \( \Omega \). Secondly, it provides a fixed \( \angle S_{21} = 90° \) at \( f_0 \), which can be considered as a modified 90° transmission line. Thirdly, the phase slope of \( \angle S_{31} \) can be adjusted by aligning the positions of two transmission poles \( f_{p1} \) and \( f_{p2} \), which is the critical parameter for this proposed quadrature feeding network. As shown in Figure 2, we can find the impedances of two shunt stubs, that is, \( Z_1 \) and \( Z_3 \), can control the phase slope of \( \angle S_{21} \) as required. Meanwhile, the impedance of middle line \( Z_2 \) controls the level of return loss. It can be also viewed as the asymmetrical case of the generalized multimode phase shifter under the mode = 0 in [15]. These two shunt stubs act as the K-inverters so as to simultaneously realize the impedance transformation and phase slope alignment by controlling coupling at two ports. In the following, synthesis method and theoretical analysis for the proposed feeding network will be given in detail.

Based on the microwave network theory [16], the S-parameter of the proposed feeding network can be derived by multiplying the ABCD matrices of the cascaded sections in Figure 1. Supposing the signal from port 1 evenly flows to each sub-network, the \( \angle S_{21} \) and \( \angle S_{31} \) can be obtained as

\[
\angle S_{21} = -\arctan \left( \frac{B + Ck}{j(A + Dk)} \right) = -\arctan \left( \frac{a_0\omega^2 + a_0}{b_1\omega} \right) \tag{1a}
\]

\[
\angle S_{31} = -\left( \frac{\pi}{2} + \theta_2 \right) \frac{f}{f_0} \tag{1b}
\]

where \( \omega = \tan(\theta) = \tan(\theta_0) f/f_0 = \tan(0.5\pi \times f/f_0) \) and coefficients \( a_0, a_2, \) and \( b_1 \) in Equation (1a) are

\[
\begin{align*}
  a_2 &= kZ_1 + kZ_2^2 + kZ_3 \\
  a_0 &= -kZ_3 (Z_1 + Z_2 + Z_3) \\
  b_1 &= kZ_2 (Z_2 Z_3 + Z_1 Z_2 + Z_1 Z_3 + kZ_1)
\end{align*}
\]

In Equation (2), \( z_1 = Z_1/Z_L, z_2 = Z_2/Z_L, z_3 = Z_3/Z_L, \) and \( k = Z_0/Z_L. \)
From Equations (1) and (2), we can find that the insertion phase $\angle S_{21}$ of the integrated asymmetrical phase shifter is a function of the normalized impedance $z_1, z_2,$ and $z_3,$ and the impedance ratio $k.$ Meanwhile, the $\angle S_{21}(f_0)$ is always equal to $90^\circ$ because $\Omega = \tan(\theta_0) = \tan(0.5\pi) = \infty$ in Equation (1a).

Then, the phase difference $\Delta\Phi$ of the proposed feeding network in Figure 1 can be derived as

$$\Delta\Phi = \angle S_{21} - \angle S_{31} = \left(\frac{\pi}{2} + \theta_2\right) \frac{f}{f_0} - \tan \left[\frac{z_2z_2^2 + z_3}{b_1\Omega}\right]$$  \hspace{1cm} (3)

To achieve a constant phase difference $\Delta\Phi$ in a certain bandwidth, the phase slope of $\angle S_{21}$ and $\angle S_{31}$ should be equal, that is

$$\left.\frac{d\angle S_{21}}{df}\right|_{f=f_0} = \left.\frac{d\angle S_{31}}{df}\right|_{f=f_0} = \left.\frac{d\angle S_{21}}{d\theta}\right|_{\theta=\theta_0} = \left.\frac{d\angle S_{31}}{d\theta}\right|_{\theta=\theta_0}$$  \hspace{1cm} (4)

Based on the relationship between two expressions of the phase slope $d\Phi/df$ and $d\Phi/d\theta,$ that is, $d\Phi/df = (\theta_0/f_0) d\Phi/d\theta,$ we can deduce that

$$\left.\frac{\angle S_{21}}{d\theta}\right|_{\theta=\theta_0} = \frac{z_2[kz_2z_3 + z_1(z_2 + z_3 + kz_3)]}{z_1(k + z_2^2)z_3}$$

$$= 1 + \Delta\Phi_0 \frac{\pi}{2}$$

where $\Delta\Phi_0$ is the phase difference of the proposed feeding network at the centre frequency $f_0,$ that is $\Delta\Phi_0 = \Delta\Phi(f_0).$

For the magnitude response, the squared magnitude of $|S_{21}|$ for the integrated phase shifter can be derived as

$$|S_{21}|^2 = \frac{1}{1 + |F_M|^2} \& F_M = \frac{S_{11}}{S_{21}}$$  \hspace{1cm} (6)

In Equation (6), $S_{11}$ and $S_{21}$ can be obtained by ABCD matrix [16] as

$$S_{11} = \frac{AZ_L + B - CZ_LZ_1 - DZ_S}{AZ_L + B + CZ_LZ_3 + DZ_S}$$

$$= \frac{A + B - (C + D)k}{A + B + (C + D)k}$$  \hspace{1cm} (7a)

$$S_{21} = \frac{2\sqrt{Z_LZ_1}}{AZ_L + B + CZ_LZ_3 + DZ_S}$$

$$= \frac{2\sqrt{k}}{A + B + (C + D)k}$$  \hspace{1cm} (7b)

where $Z_S/Z_L = k$ and the normalized $Z_L$ equals to 1.

Then, the characteristic function $F_M$ of the $|S_{21}|$ for this asymmetrical phase shifter can be derived as

$$F_M = \frac{A + B - (C + D)k}{2\sqrt{k}} = \frac{j(k_1 + k_2\Omega^2) + k_3\Omega}{k_4\Omega\sqrt{1 + \Omega^2}}$$  \hspace{1cm} (8)

where

$$\begin{align*}
k_1 &= k_2z_1(z_1 + z_2 + z_3) \\
k_2 &= z_1z_1(z_2^2 - k) \\
k_3 &= z_2(z_2 + z_1 - k(z_1 + z_2)z_3) \\
k_4 &= 2\sqrt{k_1^2z_2^2}z_3z_3
\end{align*}$$  \hspace{1cm} (9)

As discussed in [17], the Chebyshev function with $(n + q)$ order in-band equal-ripple behaviour can be expressed as

$$|S_{21}|^2 = \frac{1}{1 + |F_{bnc}|^2} = \frac{1}{1 + e^{\cos^2(n\phi + q\xi)}}$$  \hspace{1cm} (10)

where

$$\begin{align*}
\varepsilon &= \sqrt{\frac{10RL/10}{1 - 10RL/10}} & a &= \frac{1}{\cos \theta_c} \\
\cos \phi &= a\cos \theta \\
\cos \xi &= a\cos \theta \sqrt{\frac{\alpha^2 - 1}{\alpha^2\sin^2 \theta}}
\end{align*}$$  \hspace{1cm} (11a, b, c)

In Equation (11), $\varepsilon$ is the equal-ripple constant corresponding to the specified return loss RL (dB). $a$ is the quantity related to $\theta_c,$ and $\theta_c$ is the electrical length at the lower cut-off frequency $f_c$ of the passband, which can be mathematically expressed by the FBW of the passband, that is FBW = $(180^\circ - 2\theta_c)/90^\circ.$ The highest order term of $\cos(n\phi + q\xi)$ in Equation (10) is $\cos^\alpha + q^\alpha\sin^\alpha\theta.$ Hence, for $|F_{bnc}|^2$ in Equation (10) to have the same powers of $\cos\theta$ and $\sin\theta$ as Equation (8), both $n$ and $q$ in Equation (10) must be selected as $1.$ Then, $F_{bnc}$ can be rearranged as

$$F_{bnc} = \cos (\phi + \xi) = \frac{k_1 + k_2\Omega^2}{\Omega\sqrt{1 + \Omega^2}}$$  \hspace{1cm} (12)

where

$$\begin{align*}
k_1' &= \varepsilon \left[\alpha \left(\alpha + \sqrt{\alpha^2 - 1}\right) - 1\right] \\
k_2' &= -\varepsilon
\end{align*}$$  \hspace{1cm} (13)

To realize the specified magnitude and constant phase shift $\Delta\Phi$ responses, the $F_M$ in Equation (8) and the $F_{bnc}$ in Equation (12) must be exactly the same as each other. Meanwhile, the Equation (5) should be satisfied. Forcing $k_1/k_4 = k_1',$ $k_2/k_4 = k_2',$ $k_3 = 0$ and the relationship in Equation (5), the impedances $Z_1, Z_2,$ and $Z_3$ can be exactly determined with the prescribed $\Delta\Phi, RL, Z_S,$ and $Z_L.$
Figure 3 depicts synthesized curves of $Z_1 \sim Z_3$ for this integrated asymmetrical phase shifter under different cases. In Figure 3, the values of $Z_1$, $Z_2$, and $Z_3$ can be obtained directly. For example, when $\Delta \Phi = 90^\circ$, $RL = 15 \text{ dB}$, $Z_S = 100 \Omega$, and $Z_L = 50 \Omega$, we can derive that $Z_1 = 135.80 \Omega$, $Z_2 = 59.08 \Omega$, and $Z_3 = 31.59 \Omega$. Meanwhile, the unachievable phase shift region is denoted as a grey line box under the limitation of microstrip impedance $Z_m < 160 \Omega$ (for the used substrate). $\Delta \Phi = 90^\circ$ can be achieved in both cases. Figure 3 also reveals that the proposed quadrature feeding network shown in Figure 1 can be applied for the different terminated impedance ($Z_L$), such as $Z_L = 70 \Omega$ in Figure 3b.

Based on the above-described synthesis equations, we summarize the design procedure to determine all parameters for the proposed quadrature feeding network.

1. Electrical length $\theta_1$ is set as $90^\circ$ for the operation frequency $f_0$ and electrical length $\theta_2$ is obtained according to the desired phase difference ($\Delta \Phi_0$), that is, $\theta_2(f_0) = \Delta \Phi_0$. For the quadrature feeding network, $\theta_2(f_0) = 90^\circ$.
2. Three normalized impedances $z_1$, $z_2$, and $z_3$ are all determined by solving Equations (5), (8), and (12) with the prescribed phase difference $\Delta \Phi_0$ return loss $RL$ and load impedance $Z_L$. And then, the final characteristic impedance $Z_1$, $Z_2$, and $Z_3$ can be obtained by $Z_1 = z_1 Z_L$, $Z_2 = z_2 Z_L$, and $Z_3 = z_3 Z_L$.
3. Due to equal-amplitude response, the impedance $Z_4$ is set equal to the source impedance $Z_0$ that is, $Z_4 = Z_3 = 2Z_0$. Meanwhile, the impedance $Z_5$ is set as $(2Z_0 Z_L)^{1/2}$, which acts as an $90^\circ$ impedance transformer. For $Z_0 = Z_L = 50 \Omega$, we can derive that $Z_4 = 100 \Omega$ and $Z_5 = 70.71 \Omega$.
4. Choose the suitable substrate parameters (thickness $b$ and relative permittivity $\varepsilon_r$) and map all circuit parameters (electrical length and impedance) into the physical layout.

To examine the effectiveness of this synthesis method, the proposed feeding network with prescribed different phase difference $\Delta \Phi = 90^\circ$, $180^\circ$, $270^\circ$ and load impedance $Z_L = 50$ and $70 \Omega$ are designed. The simulated phase difference responses are shown in Figure 4 and corresponding synthesized values are given in Table 1. From Figure 4, we can find that the constant phase differences at $90^\circ$, $180^\circ$, and $270^\circ$ are achieved as desired. Meanwhile, the phase responses keep nearly unchanged when $Z_L$ takes different values for all three $\Delta \Phi$ cases. It is shown that the proposed feeding network has stable wideband impedance transformation ability and suitable for the different load impedances. What’s more, all impedance values of $Z_1 \sim Z_3$ are realizable. Thus, the proposed design concept and synthesis method are well validated theoretically.

![Figure 3](image3.png)

**FIGURE 3** Synthesized curves of the proposed asymmetrical phase shifter with the prescribed constant phase shift $\Delta \Phi$, return loss $RL$, impedance of source $Z_S$ and load $Z_L$. (a) $Z_S = 100 \Omega$, $Z_L = 50 \Omega$, (b) $Z_S = 100 \Omega$, $Z_L = 70 \Omega$

![Figure 4](image4.png)

**FIGURE 4** Simulated phase shift responses of the proposed feeding network under the different load impedance $Z_L$.

3 | EXPERIMENTAL RESULTS AND DISCUSSION

To experimental verification, the proposed quadrature feeding network is fabricated and tested in this section, and its performance will be investigated against the traditional counterpart in [10]. Both operate at the $1.0 \text{ GHz}$ and are fabricated on the same substrate. The configuration of proposed quadrature feeding network is shown in Figure 5. It is etched on a RO4003C substrate with $b = 0.813 \text{ mm}$, $\varepsilon_r = 3.55$, and $\tan \delta = 0.0027$. The feeding lines are designed as a set of Archimedes gradual-change lines and arranged in concentric circles. The overall size is $L_x \times L_y = 58 \times 58 \text{ mm} (0.32 \lambda_g \times 0.32 \lambda_g$ at centre frequency

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\( f_0 = 1 \text{ GHz} \) and \( \lambda_g \) is the wave-guided wavelength). For the prescribed RL = 15 dB, \( \Delta \Phi = 90^\circ \), and \( Z_L = Z_0 = 50 \, \Omega \), the circuit parameters of designed feeding network can be determined: \( Z_1 = 135.80 \, \Omega \), \( Z_2 = 59.08 \, \Omega \), \( Z_3 = 31.59 \, \Omega \), \( Z_4 = 100 \, \Omega \), \( Z_5 = 70.71 \, \Omega \), and \( \theta_1 = \theta_2 = 90^\circ \) by using synthesis method in Section 2. The final dimensions can be mapped from the calculated impedances and electrical lengths after a few tuning processes are executed by using Ansys HFSS to compensate for junction effects, such as \( l_{01} = 8 \, \text{mm}, \, r_1 = 14.9 \, \text{mm}, \, r_2 = 21 \, \text{mm}, \) \( r_3 = 16.4 \, \text{mm}, \, r_4 = 10 \, \text{mm}, \, w_1 = 0.15 \, \text{mm}, \, w_2 = 1.4 \, \text{mm}, \) \( w_3 = 3.43 \, \text{mm}, \, w_{01} = 1.8 \, \text{mm}, \, w_{02} = 1.0 \, \text{mm}, \, w_{03} = 0.45 \, \text{mm}, \) \( \gamma_1 = 255^\circ, \gamma_2 = 150^\circ, \) and \( \gamma_3 = 119^\circ \). Figure 6 gives photographs of these two fabricated prototypes.

Figure 7 shows the frequency response of the traditional design. As shown in Figure 7a, the measured \(|S_{11}|\) is lower than \(-10 \, \text{dB}\) within the frequency range from 0.1 to 2.0 GHz. However, the equal amplitude of \(|S_{21}|\) and \(|S_{31}|\) can only be obtained at one frequency point \( (f_0 = 1 \, \text{GHz}) \). The measured \(|S_{21}| = |S_{31}| = -3.7 \, \text{dB} \) at \( f_0 \) is slightly worse than the simulated one \((-3.2 \, \text{dB})\) due to the extra losses introduced by the SMA connectors, roughness of metal surface, unexpected radiation, and so forth. The bandwidth of \(|S_{21} - S_{31}| < 1.8 \, \text{dB}\) is 58% \((0.74–1.34 \, \text{GHz})\). Meanwhile, the desired 90° phase difference is only obtained at the centre frequency \( f_0 \) as shown in Figure 7b. The bandwidth of 90° \( \pm 10^\circ \) is only 14% \((0.94–1.08 \, \text{GHz})\).

For the proposed design, the measured bandwidth of \(|S_{11}| < -10 \, \text{dB}\) is 82% \((0.61–1.46 \, \text{GHz})\) which is smaller than previous \(|S_{11}|\) bandwidth in Figure 7a. However, the equal amplitude of \(|S_{21}|\) and \(|S_{31}|\) can be achieved at two frequency points, that is, \( f_1 = 0.64 \, \text{GHz}, \, f_2 = 1.24 \, \text{GHz} \), as shown in Figure 8a. The measured \(|S_{21}| = |S_{31}| = -3.7 \, \text{dB} \) at \( f_1 \) and \(|S_{21}| = |S_{31}| = -3.5 \, \text{dB} \) at \( f_2 \). The insertion loss is slightly better than the traditional one and the worst amplitude imbalance of 1.8 dB occurred around the \( f_0 \). We can find the

### TABLE 1 Synthesized values for the different phase difference \( \Delta \Phi \) and load \( Z_L \) of this proposed feeding network.

| Case 1 | \( Z_0, \, \Omega \) | \( Z_L, \, \Omega \) | \( Z_0, \, \Omega \) | \( Z_L, \, \Omega \) | \( Z_0, \, \Omega \) | \( Z_L, \, \Omega \) |
|--------|----------------|----------------|----------------|----------------|----------------|----------------|
| \( \Delta \Phi = 90^\circ \) | 135.80 | 59.08 | 31.59 | 100 | 70.71 |
| \( \Delta \Phi = 180^\circ \) | 51.25 | 59.08 | 17.87 | 100 | 70.71 |
| \( \Delta \Phi = 270^\circ \) | 31.50 | 59.08 | 12.43 | 100 | 70.71 |

| Case 2 | \( Z_0, \, \Omega \) | \( Z_L, \, \Omega \) | \( Z_0, \, \Omega \) | \( Z_L, \, \Omega \) | \( Z_0, \, \Omega \) | \( Z_L, \, \Omega \) |
|--------|----------------|----------------|----------------|----------------|----------------|----------------|
| \( \Delta \Phi = 90^\circ \) | 100.15 | 69.90 | 49.03 | 100 | 83.67 |
| \( \Delta \Phi = 180^\circ \) | 45.18 | 69.90 | 26.49 | 100 | 83.67 |
| \( \Delta \Phi = 270^\circ \) | 29.09 | 69.90 | 18.11 | 100 | 83.67 |

**FIGURE 5** Configuration of the proposed new quadrature feeding network. (a) Top view; (b) Side view

**FIGURE 6** Photographs of two fabricated prototypes. (a) Traditional one; (b) Proposed one
bandwidth of $|S_{31} - S_{21}| < 1.8$ dB is enhanced from 58% to 101% (0.48–1.48 GHz) by 1.7 times. Meanwhile, Figure 8b shows the predicted constant 90° phase difference is achieved as well. The bandwidth of $90° \pm 10°$ is enhanced from 14% to 83% (0.56–1.35 GHz) by 5.9 times. The simulated and measured results are found in good agreement with each other.

In order to assess the amplitude balance and phase balance performance simultaneously for quadrature feeding network, the parameter of axial-ratio (AR) is herein introduced, which represents the performance of circular polarization. Based on the antenna theory in [18,19], the relationship between AR, $|S_{31}|/|S_{21}|$, and $\Delta S_{31} - \Delta S_{21}$ is satisfied as below

$$
\gamma = \tan^{-1} (|S_{31}|/|S_{21}|) \quad (14a)
$$

$$
\delta = \varphi_{31} - \varphi_{21} \quad (14b)
$$

$$
\epsilon = 0.5 \sin^{-1} (\sin(2\gamma) \sin \delta) \quad (14c)
$$

$$
AR = 1/\tan \epsilon \quad (14d)\
$$

Using Equations (14a)–(14d), the theoretical AR responses can be synthesized based on the experimental results of the traditional and proposed quadrature feeding networks in Figures 7 and 8, and the relevant results are depicted in Figure 9. Compared with the traditional quadrature feeding network, the proposed design can enhance the $AR < 3$ dB bandwidth from 26% (0.88–1.14 GHz) to 98% (0.49–1.44 GHz) by more than 3.7 times. What's more, the overall circuit size ($\pi r_0^2$, $r_0 = 0.126 \lambda_g$) is nearly unchanged compared with the traditional counterpart ($\pi r_0^2$, $r_0 = 0.119 \lambda_g$) as shown in Figure 6. Table 2 tabulates the performance comparison

**FIGURE 7** Simulated and measured results of the traditional quadrature feeding network in Figure 6a. (a) S-parameter magnitude response, (b) Phase difference response

**FIGURE 8** Simulated and measured results of the proposed quadrature feeding network in Figure 6b. (a) S-parameter magnitude response, (b) Phase difference response

**FIGURE 9** Simulated and measured synthesized AR responses of the traditional and proposed quadrature feeding network
with several previous works. The proposed structure not only can realize wideband equal amplitude and 90° phase difference, but also maintain the compact size and simple geometry.

4 | CONCLUSION

A new class of quadrature feeding network with compact size and wide bandwidth is presented. The integrated asymmetrical phase shifter can provide 90° phase-shift value, phase slope alignment, and impedance transformation simultaneously in one component. Operation principle and synthesis method are demonstrated. Compared with a traditional compact counterpart, this proposed design can effectively enhance the amplitude balance bandwidth by 1.7 times and phase balance bandwidth by 5.9 times under the condition of the similar circuit size.

ACKNOWLEDGEMENTS

This work was supported in part by the National Natural Science Foundation of China (Grant no. 61,901,226), in part by the National Natural Science Foundation of Jiangsu Province (Grant no. BK20190727), in part by Nanjing University of Posts and Telecommunications Scientific Foundation (Grant no. NY219128), in part by Universities Natural Science Research General Project (Grant no. 19KJB510045), and in part by the Multi-Year Research Grant from the University of Macau (Grant no. MYRG2018-00,073-FST).

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How to cite this article: LyuYP, LiB, ZhuL, ChengCH. A new compact and wideband quadrature feeding network with an integrated asymmetrical phase shifter. IET Microw. Antennas Propag. 2021;15:397–403. https://doi.org/10.1049/mia2.12066