

Article

Co-Design of Integrated Microwave Amplifier and Phase Shifter Using Reflection-Type Input Matching Networks for Compact MIMO Systems

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Abstract

This paper presents a co-design approach for a microwave amplifier–phase shifter that integrates an arbitrary termination impedance reflection-type phase shifter as the input matching network of a microwave transistor. Since the proposed reflection-type phase shifter input matching network is capable of transforming both real and/or complex impedances to a system impedance of 50 Ω , the co-design approach can directly match the optimum source impedance of the microwave transistor to 50 Ω through a reflection-type phase shifter input matching network. To validate the proposed method, prototypes of microwave amplifier–phase shifters with different input matching networks configurations are designed, fabricated, and measured with a center frequency of 2.45 GHz. The experimental results demonstrate that the proposed co-design microwave amplifier–phase shifter achieves improved electrical performances compared to the conventional approach, where a 50-to-50 Ω termination impedance phase shifter is cascaded with a 50-to-50 Ω termination impedance conventional microwave amplifier. Measurement results demonstrate that the gains of a standalone conventional microwave amplifier, a cascaded phase shifter with a conventional microwave amplifier, and the proposed co-design microwave amplifier–phase shifter are 14.13 dB, 13.28 dB, and 13.74 dB, while the 1 dB compression points are 25.72 dBm, 24.77 dBm, and 25.26 dBm, respectively. Within the 200 MHz bandwidth, the proposed co-design microwave amplifier–phase shifter exhibits a maximum phase shift range of 185.62° and a phase deviation error of $\pm 4.3^\circ$. The circuit size of the co-designed microwave amplifier–phase shifter is 38.5% smaller than the conventional cascaded phase shifter with a conventional microwave amplifier.

Keywords: arbitrary termination impedance; co-design; microwave amplifier; phase deviation error; phase shift range; reflection-type phase shifter



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1. Introduction

Modern communication systems require compact, efficient, lightweight, low-loss, and high-performance microwave circuits. Microwave circuits with multiple functionalities are essential for meeting these requirements [1–6]. Microwave amplifiers and phase shifters

are important circuits in radio frequency (RF) front-end systems. Microwave amplifiers are typically used to amplify the power level of a signal for long-distance transmission. Conversely, a phase shifter is an essential circuit used for controlling the phase of a signal without significantly affecting its amplitude. By adjusting the phase, the propagating signal can be controlled in order to steer and transmit in specific direction, which is crucial for applications such as phased array antennas, multiple input and multiple output (MIMO) systems, and beamforming networks [7–10].

The microwave frequency typically ranges from 300 MHz to 300 GHz, which is divided into several sub-bands, such as L-band (1~2 GHz), S-band (2~4 GHz), C-band (4~8 GHz), X-band (8~12 GHz), Ku-band (12~18 GHz), K-band (18~27 GHz), Ka-band (27~40 GHz), and millimeter bands above 30 GHz. This work chooses a center frequency (f_0) of 2.45 GHz, which is in the S-band. More specifically, this frequency is part of the industrial, science, and medical (ISM) radio bands. The 2.45 GHz ISM band is license-free and has been standardized around the world, making it popular for a wide range of uses, including Bluetooth, Wi-Fi, microwave heating, and biomedical telemetry.

Figure 1a shows the typical structure of an RF front-end transmitting system where phase shifters and microwave amplifiers are individually designed with termination impedances equal to a system impedance (Z_0) of 50 Ω . In this conventional approach, if the return loss (RL) characteristics of individual circuits are not good, then cascading two circuits may cause signal reflection due to the impedance mismatch. Furthermore, the conventional cascading topology may lead to increases in circuit size, cost, and degrade overall electrical performances, resulting from the increasing insertion loss (IL) by mismatching and additional devices such as adaptor and cable. To mitigate these disadvantages, Figure 1b illustrates a co-design approach, where the phase shifter is matched directly to the optimum source impedance (Z_S) of the microwave transistor, eliminating the need for additional input matching networks. Therefore, the co-designed approach can allow the circuit miniaturization and performances improvement of the RF front-end by eliminating a number of components (such as additional matching networks, connecting cables, and adapters) and their corresponding ILs.

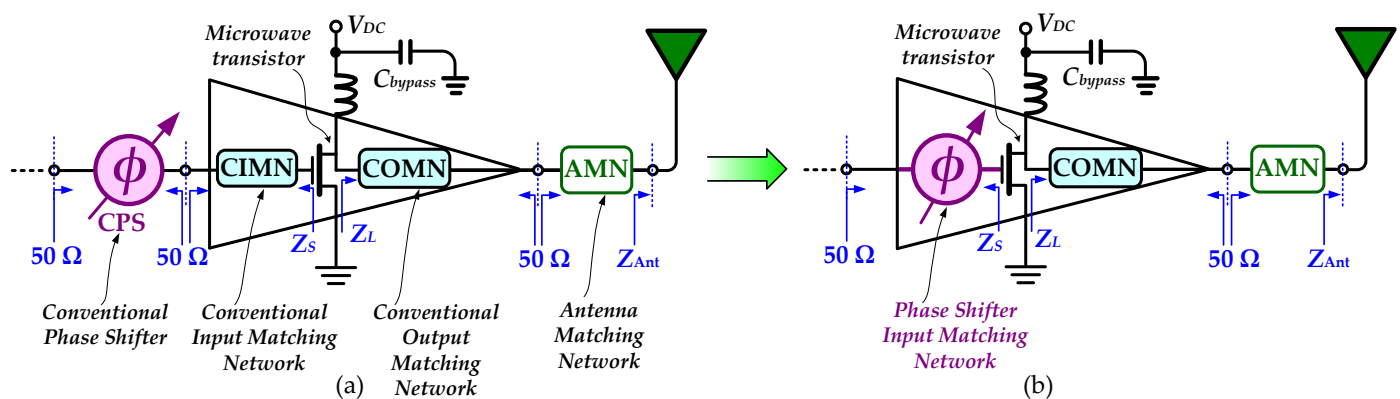


Figure 1. RF front-end transmitter of a wireless communication system: (a) conventional cascading design topology and (b) proposed co-designed microwave amplifier phase shifter (CPS: conventional phase shifter; CIMN: conventional input matching network; COMN: conventional output matching network; AMN: antenna matching network).

In recent years, various configurations of phase shifters have been investigated. Based on their circuit topologies, these phase shifters can be classified as transmission types [11,12] and reflection types [13–20]. Compared to transmission-type configurations, reflection-type phase shifters can provide good RL by using identical reflection loads connected at the through and coupled ports of the 3 dB branch-line hybrid. Recently, numerous

techniques have been explored for designing single-band/dual-band reflection-type phase shifters [13–15] and wideband reflection-type phase shifters [16]. In [17–19], in-band phase deviation error minimization techniques have been explored using coupled lines. Meanwhile, a highly compact single-unit reflection-type phase shifter with large phase shifter range was presented in [20]. Additionally, the tunable reflection-type phase shifter was described in [21–23]. Although there are significant advancements in reflection-type phase shifter design, the conventional reflection-type phase shifter is designed for a fixed $50\ \Omega$ termination impedance, which limits design flexibility, especially when integrating reflection-type phase shifter with other circuits.

To overcome the limitations of conventional reflection-type phase shifters, this work proposes a new arbitrary termination impedance reflection-type phase shifter. This arbitrary termination impedance reflection-type phase shifter is then used as an input matching network of a microwave transistor in the co-design of the integrated microwave amplifier–phase shifter. The proposed phase shifter can provide not only phase control functionality but also offer impedance matching functionalities so that the optimum source impedance of the microwave transistor is directly matched with the arbitrary termination impedance reflection-type phase shifter. This co-design approach leads to improved gain, output power, reduced circuit size, and simplified design complexity compared to the traditional cascading approach. To validate the effectiveness of the proposed co-design methodology of the microwave amplifier–phase shifter, three prototypes of microwave amplifiers with different input matching networks were designed, fabricated, and measured.

2. Design Method

Figure 2 shows the co-design of the microwave amplifier–phase shifter, which is composed of the proposed reflection-type phase shifter as the input matching network, microwave transistor, and output matching network. The phase of the input signal can be controlled by using reflection-type phase shifter input matching network realized with arbitrary termination impedance 3 dB branch–line hybrid and reflection loads.

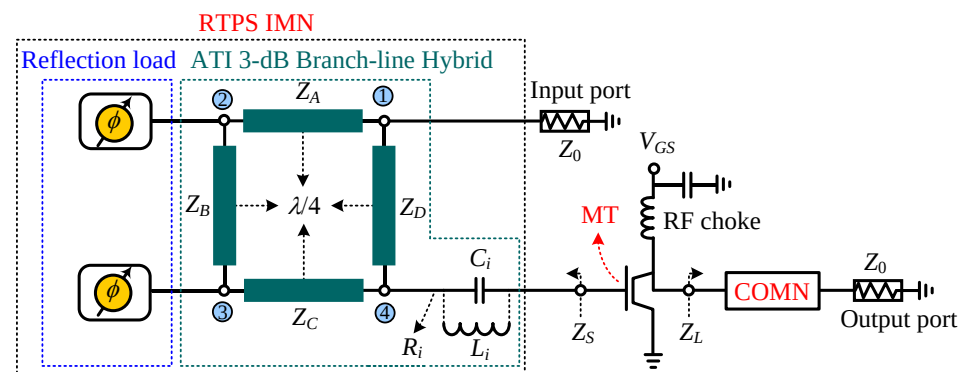


Figure 2. Structure of the proposed co-design of the microwave amplifier–phase shifter (RTPS: reflection-type phase shifter; IMN: input matching network; ATI: arbitrary termination impedance; MT: microwave transistor; COMN: conventional output matching network).

2.1. Arbitrary Termination Impedance 3 dB Branch–Line Hybrid

Figure 3a illustrates the proposed structure of the arbitrary termination impedance 3 dB branch–line hybrid, in which the input, through, coupled, and isolation ports are terminated with $Z_i = R_i \pm jX_i$, where $i = 1, 2, 3$, and 4, respectively. For convenience, the intersection points of each branch are denoted by nodes 1, 2, 3, and 4. To match the complex impedance to real impedance (R_i), the imaginary part (X_i) of Z_i must be canceled using

either the capacitor (C_i) or inductor (L_i). Depending on the sign of the imaginary part of Z_i , the values of C_i and L_i can be determined as follows:

$$C_i = \frac{1}{\omega_0 |X_i|} \quad \text{for } +X_i \quad (1)$$

$$L_i = \frac{|X_i|}{\omega_0} \quad \text{for } -X_i \quad (2)$$

where ω_0 is the angular center frequency. Once X_i is cancelled, pure real impedance R_i remains at each node, which will transform from one node impedance to another node impedance by the proposed 3 dB branch-line hybrid. Since the 3 dB branch-line hybrid consists of quarter-wavelength ($\lambda/4$) transmission lines (TLs), the characteristic impedances can be found by using equivalent structure with isolation characteristics shown in Figure 3b,c.

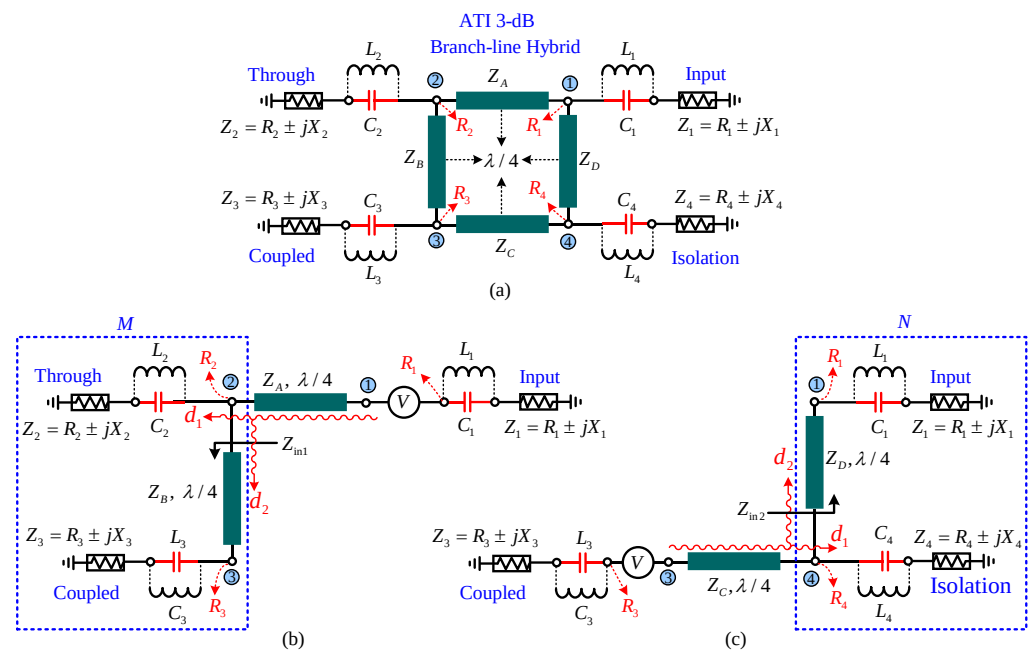


Figure 3. Arbitrary termination impedance 3 dB branch-line coupler: (a) full structure, (b) simplified structure under the assumption of $|S_{41}| = 0$, and (c) reciprocal structure of (b) (ATI: arbitrary termination impedance).

If the isolation port is perfectly isolated (i.e., $|S_{41}| = 0$), there is no signal coupled to node 4 when the signal is excited at node 1. Under this condition, the input power is equally divided between nodes 2 and 3, denoted as d_1 and d_2 , respectively, as shown in Figure 3b. The resulting power ratio between d_1 and d_2 is 1:1. So, the voltage across R_2 at node 2 is equal to that across TL Z_B terminated with R_3 at node 3. Therefore, the impedance (Z_{in1}) looking into the TL Z_B terminated with R_3 must be R_2 [24]. Denoting $M = R_2/2$ as the equivalent impedance of parallel Z_{in1} with R_2 at node 2, the characteristic impedances Z_A and Z_B of TL can be defined as follows:

$$Z_A = \sqrt{\frac{R_1 R_2}{2}} \quad (3)$$

$$Z_B = \sqrt{R_2 R_3} \quad (4)$$

Similarly, due to reciprocal characteristics, the circuit in Figure 3c is the simplified equivalent circuit in condition input signal excited at node 3. The voltage across R_4 at node 4 is equal to that across TL Z_D terminated with R_1 at node 1. Thus, the impedance (Z_{in2})

of TL Z_D terminated with R_1 must be R_4 . Assuming $N = R_4/2$ as equivalent impedances of parallel Z_{in2} with R_4 at node 4, the characteristic impedances Z_C and Z_D of TL can be expressed as follows:

$$Z_C = \sqrt{\frac{R_3 R_4}{2}} \quad (5)$$

$$Z_D = \sqrt{R_1 R_4}. \quad (6)$$

For equation validation, the calculated parameters for arbitrary terminated 3 dB branch-line hybrids with different termination impedances are shown in Table 1. The simulation was performed using advanced design system (ADS) software. Based on (1)–(6), the values of C_m , L_m , Z_A , Z_B , Z_C , and Z_D are calculated at f_0 of 2.45 GHz.

Table 1. Calculated values of C_m , L_m , Z_A , Z_B , Z_C , and Z_D according to different Z_m at $f_0 = 2.45$ GHz.

	Z_1 [Ω]	Z_2 [Ω]	Z_3 [Ω]	Z_4 [Ω]	C_i [pF] or L_i [nH] ($i = 1\sim4$)	Z_A [Ω]	Z_B [Ω]	Z_C [Ω]	Z_D [Ω]
Case 1	20	50	30	80	0/0/0/0	22.36	38.73	34.64	40
Case 2	25	$50 - j60$	$40 + j20$	120	0/3.9 nH/3.25 pF/0	25	44.72	48.99	54.73
Case 3	$35 - j40$	$30 + j125$	$75 - j60$	$45 + j10$	2.6 nH/0.52 pF/3.9 nH/6.5 pF	22.91	47.43	41.08	39.69

The simulated electrical performances of the designed 3 dB branch-line hybrids are shown in Figure 4. Although termination impedances are different from 50Ω , the optimum RL characteristic can be achieved at f_0 , as can be seen in Figure 4a,b. Moreover, the proposed arbitrary termination impedance 3 dB branch-line hybrid provides high isolation between the through and coupled ports, as shown in Figure 4c. Figure 4d depicts the phase difference between the output ports of the arbitrary termination impedance 3 dB branch-line hybrid. A phase difference of 90° is achieved at the center frequency f_0 .

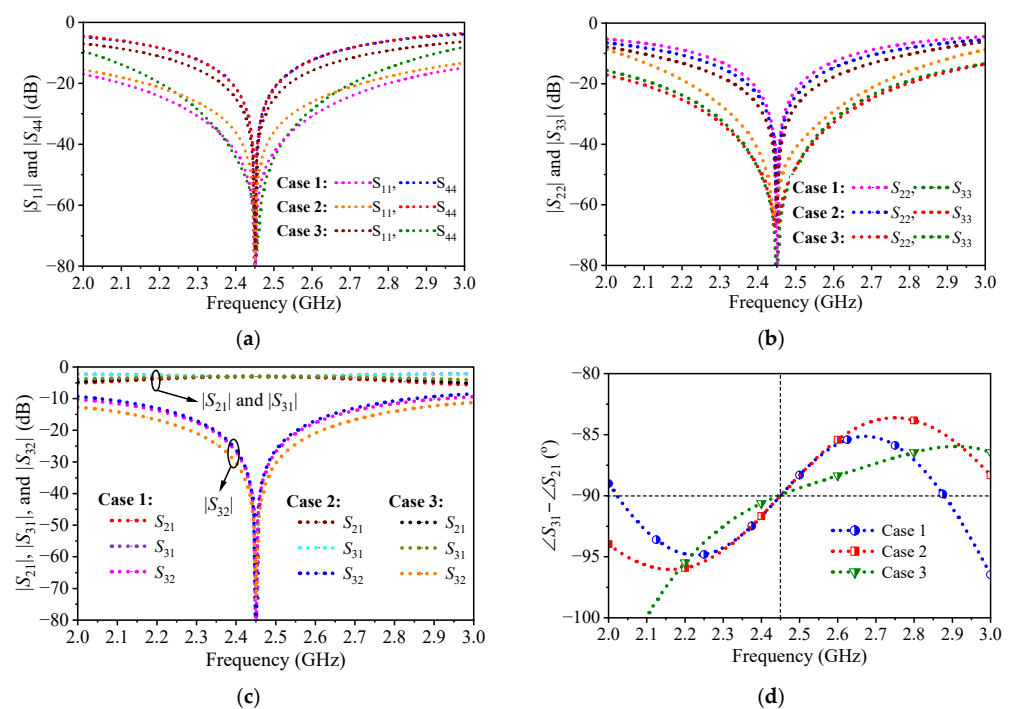


Figure 4. Simulated electrical performances of the arbitrary termination impedance 3 dB branch-line hybrid: (a) $|S_{11}|$ and $|S_{44}|$, (b) $|S_{22}|$ and $|S_{33}|$, (c) $|S_{21}|$, $|S_{31}|$, and $|S_{32}|$, and (d) $\angle S_{31} - \angle S_{21}$.

2.2. Arbitrary Termination Impedance Reflection-Type Phase Shifter

Figure 5a shows the proposed structure of the arbitrary termination impedance reflection-type phase shifter that consists of the arbitrary termination impedance 3 dB hybrid, where the through and coupled ports are terminated with two identical reflection loads. In this proposed reflection-type phase shifter, the input, through, and coupled ports are terminated with Z_0 ($Z_1 = Z_2 = Z_3 = Z_0$), while the output (or isolation) port is terminated with arbitrary termination impedance $Z_4 = R_4 \pm jX_4$. Figure 5b depicts the structure of reflection load, which is composed of a short-circuited TL with characteristic impedance Z_r and electrical length θ_r , in a series with a varactor diode with capacitance C_v .

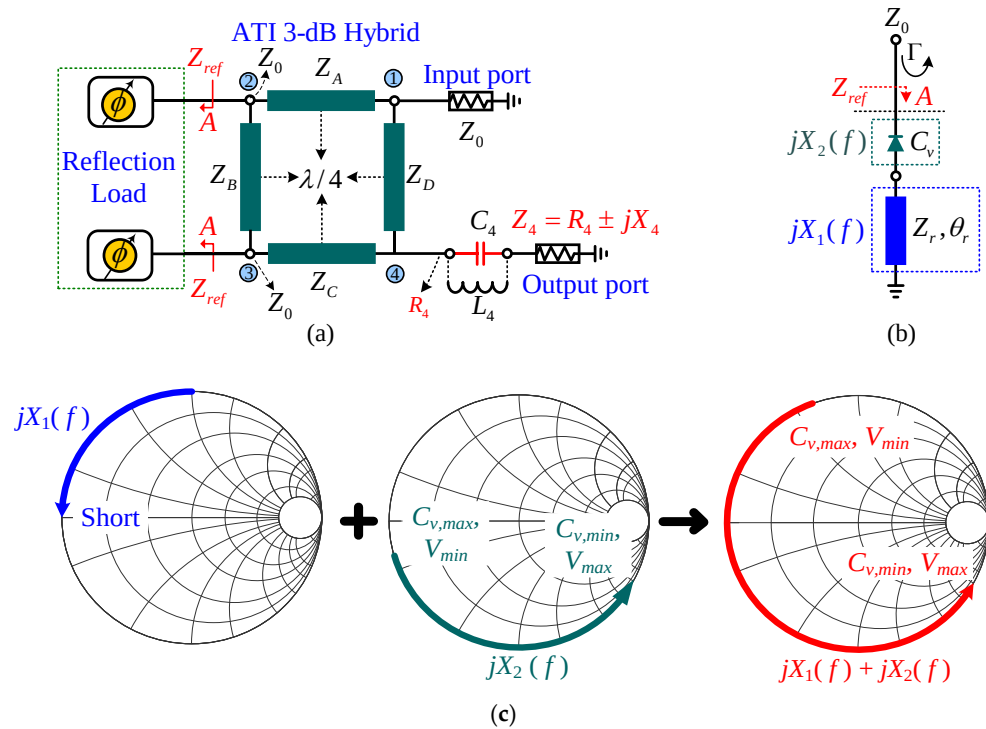


Figure 5. Proposed arbitrary termination impedance reflection-type phase shifter: (a) overall structure, (b) equivalent circuit of reflection load, and (c) phase shift range concept (ATI: arbitrary termination impedance).

By considering ideal TLs (i.e., no loss and no dispersion) and an ideal varactor diode with no junction resistance (i.e., $R_j = 0 \Omega$), the input impedance of the reflection load Z_{ref} can be defined as follows:

$$Z_{ref} = j[X_1(f) + X_2(f)] \quad (7)$$

where

$$X_1(f) = Z_r \tan \theta_r \quad (8)$$

$$X_2(f) = -\frac{1}{\omega C_v} \quad (9)$$

Assuming ideal perfect characteristics of the 3 dB arbitrary termination impedance hybrid, the transmission phase of the reflection-type phase shifter is equal to the phase of the reflection coefficient (Γ) of the load. Therefore, the Γ of the reflection load can be expressed as follows:

$$\Gamma = \frac{j[X_1(f) + X_2(f)] - Z_0}{j[X_1(f) + X_2(f)] + Z_0} \quad (10)$$

As shown in Figure 5b, the reflection load consists of two reactive elements, which are $jX_1(f)$ and $jX_2(f)$. By properly selecting θ_r of the short-circuited TL between 0° to 90° ,

$jX_1(f)$ can cover part of the upper side of the Smith chart, shown as $jX_1(f)$ in Figure 5c. Simultaneously, $jX_2(f)$ consists of a varactor diode that can cover the lower part of Smith chart, transitioning from $C_{v, max}$ to $C_{v, min}$ as the bias voltage varies from V_{min} to V_{max} , as shown in $jX_2(f)$ in Figure 5c. Once $jX_1(f)$ is series connected with $jX_2(f)$, their combined effect enables a wider reactance range, thereby broadening the achievable phase shift range, as demonstrated by $jX_1(f) + jX_2(f)$ in Figure 5c. Therefore, the transmission phase (ϕ) of the reflection-type phase shifter can be calculated as (11) using the reflection coefficient of the load:

$$\phi = \angle \Gamma = -2 \tan^{-1} \left[\frac{X_1(f) + X_2(f)}{Z_0} \right] \quad (11)$$

As a result, the phase shift range ($\Delta\phi$) of the proposed arbitrary termination impedance reflection-type phase shifter can be expressed in the following equation, where V varies from V_{min} to V_{max} [9]:

$$\Delta\phi|_V = \begin{cases} \phi|_V - \phi|_{V_{min}} & \text{if } \phi|_{V_{min}} < \phi|_{V_{max}} \\ \phi|_V - \phi|_{V_{max}} & \text{if } \phi|_{V_{max}} < \phi|_{V_{min}} \end{cases} \quad (12)$$

Lastly, the in-band phase derivation error (ϕ_{err}) within the operating bandwidth (BW) can be defined as the difference between the maximum and minimum values of $\Delta\phi|_V$:

$$\phi_{err} = \pm \frac{\max(\Delta\phi)|_V - \min(\Delta\phi)|_V}{2} \quad (13)$$

To achieve high phase shift range with low phase deviation error within the operating BW, values of Z_r , θ_r , and C_v should be chosen carefully. In this work, the SMV1231 varactor diode from Skyworks Inc. is used as variable capacitor C_v , which provides a tuning capacitance range of 0.33 pF to 4.78 pF, as shown in Figure 5a.

To select the optimum circuit parameters of the proposed arbitrary termination impedance reflection-type phase shifter, Figure 6b,c illustrate the calculated phase shift range and phase deviation error according to the variations of Z_r and θ_r . In these figures, the color bars represent phase shift range and phase deviation error values. As can be observed in Figure 6b, higher phase shift ranges can be achieved by increasing Z_r while keeping θ_r fixed, or vice versa. However, the maximum phase shift range of 245° can be achieved, and further increasing Z_r and/or θ_r results in a decreasing phase shift range. Meanwhile, once the phase shift range reaches its maximum value, further increases in Z_r and/or θ_r result in a continued increase in phase deviation error up to a certain point before it starts to decrease, as demonstrated in Figure 6c. Therefore, appropriate Z_r and θ_r can be selected through a trade-off between the desired values of phase shift range and phase deviation error.

In this paper, the arbitrary termination impedance reflection-type phase shifter was designed to achieve a phase shift range of over 180° while maintaining a phase deviation error of less than $\pm 8^\circ$ across the operational BW of 200 MHz at the center frequency (f_0) of 2.45 GHz. For the validation of the proposed arbitrary termination impedance reflection-type phase shifter, Figure 7 shows the simulated S -parameters of the three proposed reflection-type phase shifters where the input port is terminated with 50Ω and the isolation port is terminated with three different Z_4 , as shown in Table 2. To see the phase shift range and phase deviation error tendencies, each proposed reflection-type phase shifter is terminated with three different reflection loads, specifically three different pairs of Z_r and θ_r . As can be seen from Table 2, a high-phase shift range ($>200^\circ$) and low-phase deviation error ($<\pm 7.4^\circ$) within the desired BW can be achieved by selecting $Z_r = 50 \Omega$ and $\theta_r = 45^\circ$. This result reveals that by carefully selecting Z_r and θ_r , the proposed arbitrary termination

impedance reflection-type phase shifter can simultaneously provide a high-phase shift range, a low-phase deviation error, and a good RL.

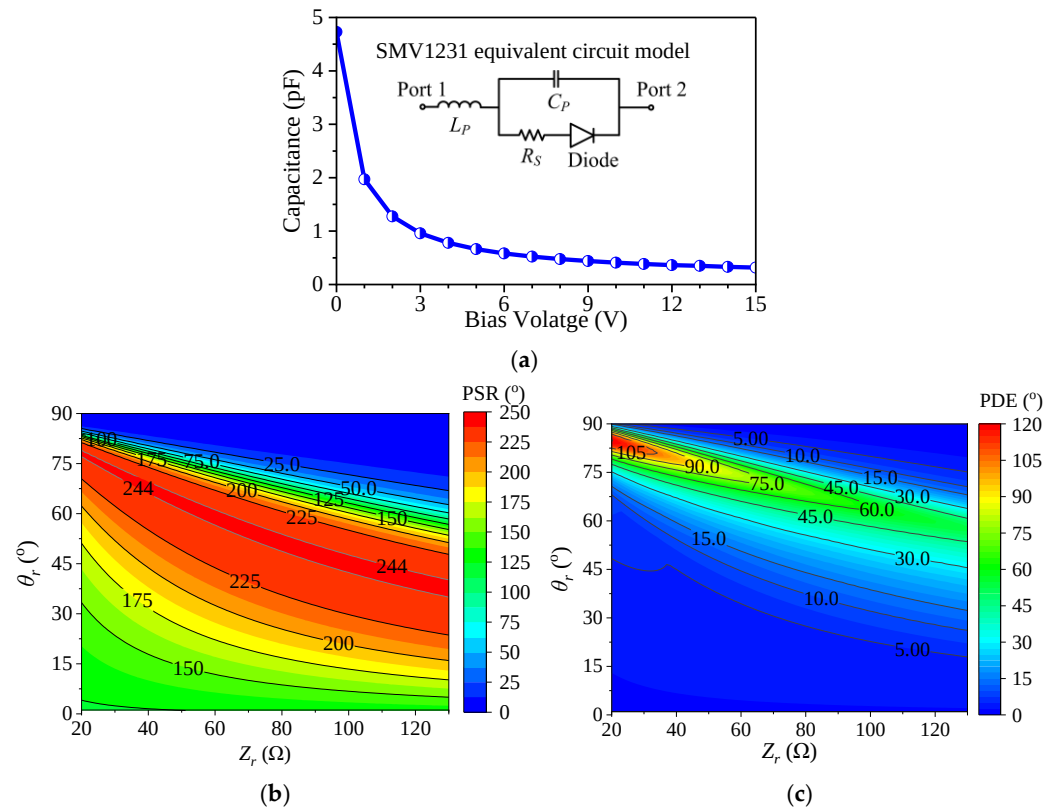


Figure 6. Parametric study of reflection load: (a) capacitance variation according to DC-bias voltage, (b) phase shift range according to Z_r and θ_r at $f_0 = 2.45$ GHz, and (c) phase deviation error according to Z_r and θ_r within BW of 200 MHz (PSR: phase shift range; PDE: phase deviation error).

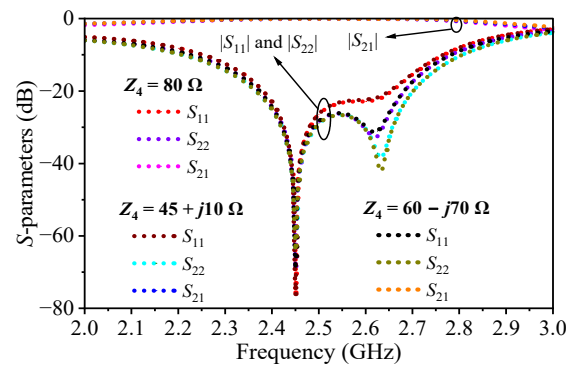


Figure 7. Simulated S-parameters of ATI reflection-type phase shifter.

Table 2. Simulated values of phase shift range = $\Delta\phi_{\max}$ and phase deviation error = ϕ_{err} within BW of 200 MHz at $f_0 = 2.45$ GHz for different Z_4 , Z_r , and θ_r . Z are units in [Ω] (PSR: phase shift range; PDE: phase deviation error).

Z_1, Z_2, Z_3	Z_4	C_4/L_4	Z_A	Z_B	Z_C	Z_D	Z_r	θ_r (°)	PSR (°)	PDE (°)
50	80	0	35.36	50	44.72	63.25	30/50/80	30/45/60	157.8/214.5/233.2	3.8/7.4/39.8
50	$45 + j10$	$C_4 = 6.5$ pF	35.36	50	33.54	47.43	30/50/80	30/45/60	157.8/214.5/233.2	3.8/7.4/39.8
50	$60 - j70$	$L_4 = 4.55$ nH	35.36	50	38.73	54.77	30/50/80	30/45/60	157.8/214.5/233.2	3.8/7.4/39.8

3. Experimental Results

For performance comparison, all microwave amplifiers in this work were designed at $f_0 = 2.45$ GHz with a target operational BW of 200 MHz. The type of microwave transistor used in this work is GaAs MESFET, specifically AH102A-G from Qorvo, which operates at frequencies from 350 MHz to 3 GHz. The optimum input (Z_{IN}) and output (Z_{OUT}) impedances of microwave transistors can be obtained by performing source and load impedance analyses based on the provided S2P file using ADS software. The S-parameters of the transistor are chosen under the bias conditions of $V_{DS} = 9$ V, $I_{DS} = 200$ mA.

Figure 8a shows a small signal equivalent model of GaAs MESFET [25]. Additionally, Figure 8b illustrates gain circles plotted at the three frequencies (i.e., $f_L = 2.35$ GHz; $f_0 = 2.45$ GHz; and $f_H = 2.55$ GHz). The flat gain performance can be achieved by selecting Z_{OUT} at the point where these three gain circles are closest to each other. Therefore, the impedances $Z_{IN} = 18 - j7 \Omega$ and $Z_{OUT} = 47 + j18 \Omega$ are selected. The source impedance (Z_S) and the load impedance (Z_L) are complex conjugates of Z_{IN} and Z_{OUT} , respectively.

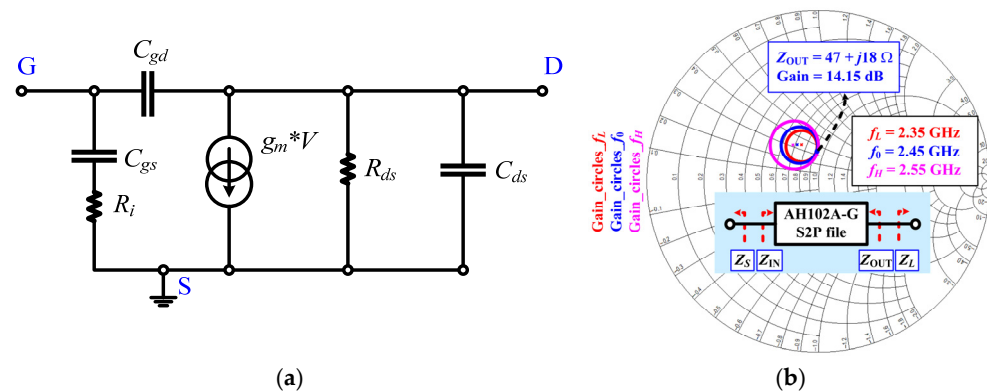


Figure 8. AH102A-G transistor (a) small signal equivalent model and (b) gain circles and impedances selection. Parameters: $C_{gs} = 2.09$ pF; $C_{gd} = 0.116$ pF; $C_{ds} = 0.3$ pF; $g_m = 0.393$ S; $R_i = 36 \Omega$; $R_{ds} = 55 \Omega$; $V = 9$ V.

Figure 9a shows the structure of the conventional microwave amplifier, where the conventional input matching network and conventional output matching network are realized using T-type TL topology. The conventional input matching network is implemented using an open-circuited stub (TL_1) in series with a transmission line (TL_2) to match the optimum source impedance Z_S . Similarly, the conventional output matching network is realized using a short-circuited stub (TL_3) in series with TL_4 to match the optimum load impedance Z_L . The TL_3 of conventional output matching network is also used as a DC feed line by shorting TL_3 with the bypass capacitor (C_{bp}). All circuits presented in this paper are implemented using Taconic-TLY PCB with $\epsilon_r = 2.2$ and $h = 0.787$ mm. For soldering and realization purposes, three TLs with $Z_{add} = 50 \Omega$ and $\theta_{add} = 10^\circ$ are added to the network. The circuit layout with the physical dimensions of the conventional microwave amplifier is presented in Figure 9b. Figure 9c demonstrates the comparison of S-parameters obtained from EM simulation by using high-frequency structure simulator (HFSS) software and measurements. The simulation and measurement results are agreed well, and a minimum gain of 14.13 dB and a RL of 20.1 dB are achieved within the desired BW.

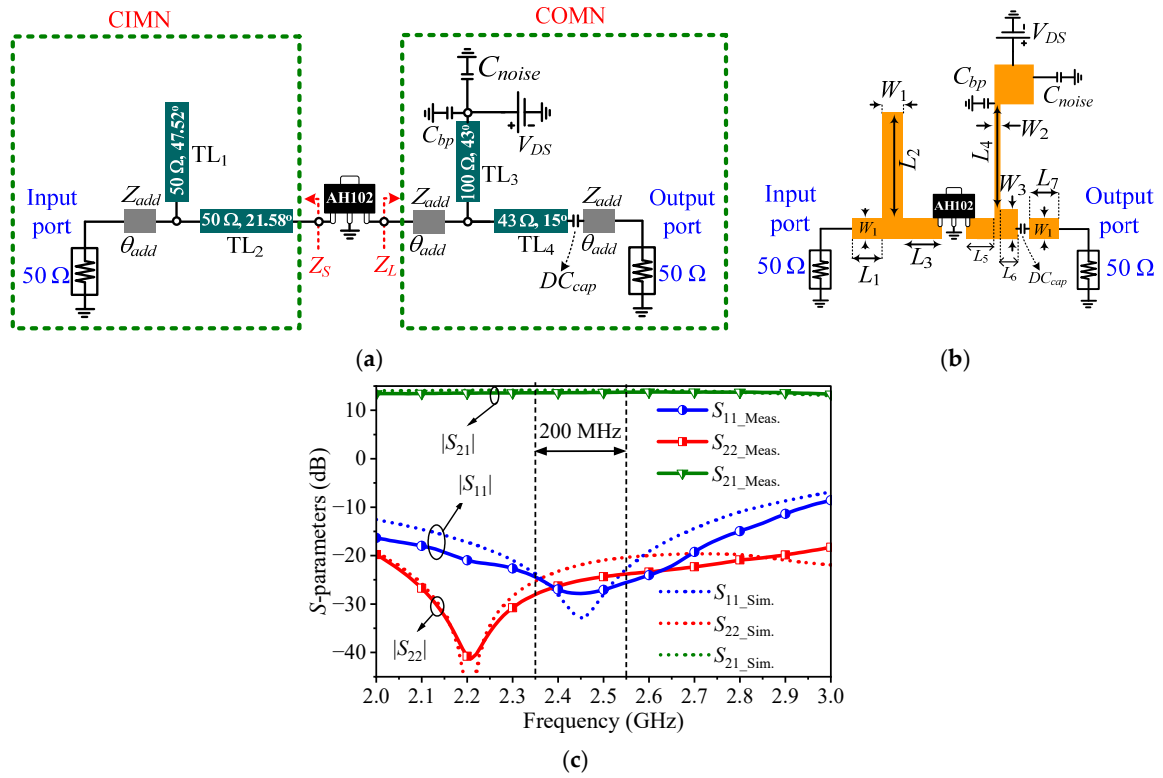


Figure 9. Conventional microwave amplifier design: (a) structure, (b) PCB layout with physical dimensions, and (c) S-parameters. Circuit dimensions: $W_1 = 2.38$; $W_2 = 0.7$; $W_3 = 3.4$; $L_1 = 10$; $L_2 = 11.82$; $L_3 = 4.45$; $L_4 = 11.9$; $L_5 = 2$; $L_6 = 4.7$; $L_7 = 12$ [unit: millimeter] (CIMN: conventional input matching network; COMN: conventional output matching).

3.1. Conventional Cascaded Design of the Microwave Amplifier–Phase Shifter

To design the conventional cascaded MA-PS, the microwave amplifier and the reflection-type phase shifter are designed individually by assuming port impedances of 50-to-50 Ω. Figure 10a depicts the PCB layout with the physical dimensions of the conventional reflection-type phase shifter by using the 3 dB branch–line hybrid [26]. The design goals of the conventional reflection-type phase shifter are to achieve a phase shift range of $>180^\circ$ and a phase deviation error of $<\pm 8^\circ$ within a bandwidth of 200 MHz at $f_0 = 2.45$ GHz. The circuit parameters of the conventional reflection-type phase shifter are given as $Z_A = Z_C = 35.36$ Ω; $Z_B = Z_D = 50$ Ω; $Z_r = 50$ Ω; $\theta_r = 45^\circ$; and C_v of varactor diode SMV1231, which provide capacitance ranging from 0.33 pF to 4.87 pF by varying the voltage from 0 V to 15 V. The values of Z_r and θ_r are selected using the method described in Section 2.2. For practical realization, additional TLs with parameters of (W_1'/L_1') and (W_1'/L_3') are used as soldering pads. Additionally, to vary the capacitance of the varactor diode, a single V_{DC} feeding line (W_3'/L_2') with $Z_{DC} = 100$ Ω and $\theta = 90^\circ$ is connected to the middle of the conventional reflection-type phase shifter vertical branch. These additional TLs affect the phase shift range and phase deviation error of conventional reflection-type phase shifter. Figure 10b illustrates the comparison of the phase shift range between the EM simulation and measurement results. The measured phase shift range is 188.97° with a maximum phase deviation error of $\pm 6^\circ$ within the desired BW. Furthermore, a comparison of S-parameters between EM simulation and measurement results is shown in Figure 10c, where the maximum IL of 0.52 dB and minimum RL of 22 dB are obtained within an operation band ranging from 2.35 GHz to 2.55 GHz.

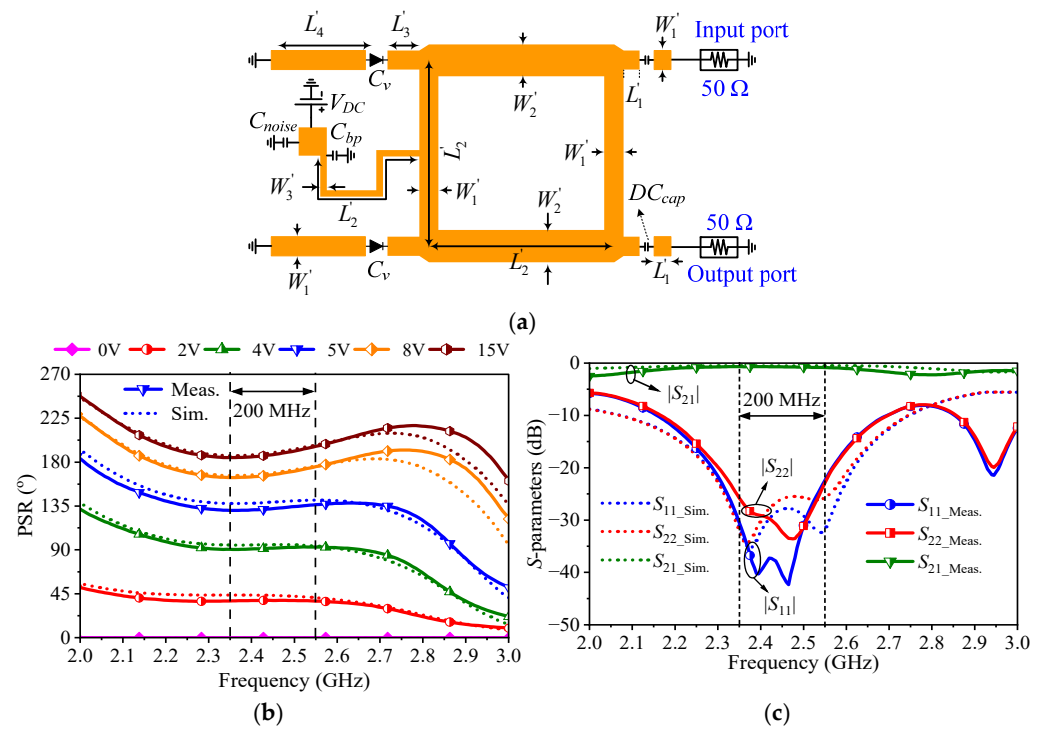


Figure 10. Conventional reflection-type phase shifter: (a) PCB layout with physical dimension; (b) phase shift range; and (c) S-parameters. Circuit dimensions: $W_1' = 2.38$; $W_2' = 3.95$; $W_3' = 0.8$; $L_1' = 2$; $L_2' = 23.2$; $L_3' = 4$; $L_4' = 11.87$ [unit: millimeter] (PSR: phase shift range).

Figure 11a depicts the circuit layout of the cascaded microwave amplifier–phase shifter, where the output port of the conventional reflection-type phase shifter is connected to the input port of the conventional microwave amplifier. As the space between the input port of conventional reflection-type phase shifter and conventional input matching network of conventional microwave amplifier is too small, the additional TL with physical parameters of (W_1''/L_1'') is used and bent by 90° for the input port connection. Figure 11b,c show the phase shift range and S-parameters comparison between the simulation and measurement results of the conventional reflection-type phase shifter cascaded with a conventional microwave amplifier at f_0 of 2.45 GHz, respectively. Since the conventional reflection-type phase shifter contains IL, the cascading conventional reflection-type phase shifter with conventional microwave amplifier can degrade electrical performances such as gain and output power. In addition, if each of these circuits is not perfectly matched to 50 Ω , additional losses will occur due to an impedance mismatch. As a result, the measured maximum phase shift range is 195.35° at f_0 and the maximum phase deviation error over BW of 200 MHz is $\pm 6.14^\circ$, which is comparable to the simulation. Moreover, minimum return losses (S_{11} and S_{22}) of 16.2 dB and a minimum gain (S_{21}) of 13.28 dB are achieved.

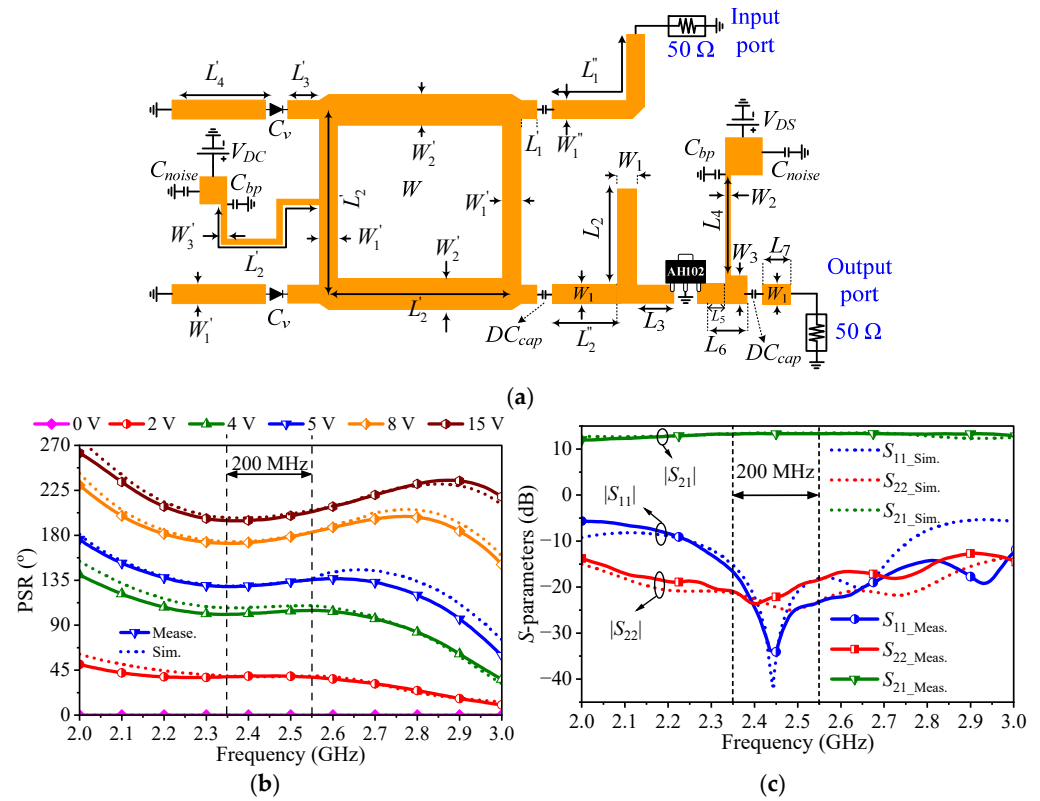


Figure 11. Conventional reflection-type phase shifter cascaded with conventional microwave amplifier: (a) PCB layout with physical dimension, (b) phase shift range, and (c) S-parameters. Circuit dimensions: $W_1'' = 2.38$; $L_1'' = 15.5$; $L_2'' = 14$ [unit: millimeter] (PSR: phase shift range).

3.2. Co-Design of the Microwave Amplifier–Phase Shifter

Figure 12a shows the physical layout of the proposed co-design microwave amplifier–phase shifter. In this proposed microwave amplifier–phase shifter, the input matching network is designed using an arbitrary termination impedance reflection-type phase shifter, whereas the output matching network remained the same as the conventional output matching network. Since it is difficult to connect the capacitor C_4 to the gate pin of the microwave transistor, a TL with (W_1'/L_3^*) parameters is used. According to this additional TL, $Z_S = 18 + j7 \Omega$ is transformed to $Z_4 = 19.05 + j13.33 \Omega$. The impedance Z_4 will be used in the calculation for the design parameters of the proposed reflection-type phase shifter input matching network. By using (1), the high-Q capacitor $C_4 = 4.87$ pF is used to cancel the imaginary part of $+j13.33 \Omega$. It is noteworthy that this C_4 also functions as a DC-block for the supply voltage fed from the reflection-type phase shifter. The remaining resistance of 19.05Ω is considered terminated impedance R_4 at node 4. Therefore, $Z_A = 35.36 \Omega$, $Z_B = 50 \Omega$, $Z_C = 21.82 \Omega$, and $Z_D = 30.86 \Omega$ are calculated using (3)–(6). Meanwhile, the reflection load is identical to that used in Section 3.1, specifically $Z_r = 50 \Omega$, $\theta_r = 45^\circ$, and C_v , ranging from 0.33 pF to 4.87 pF.

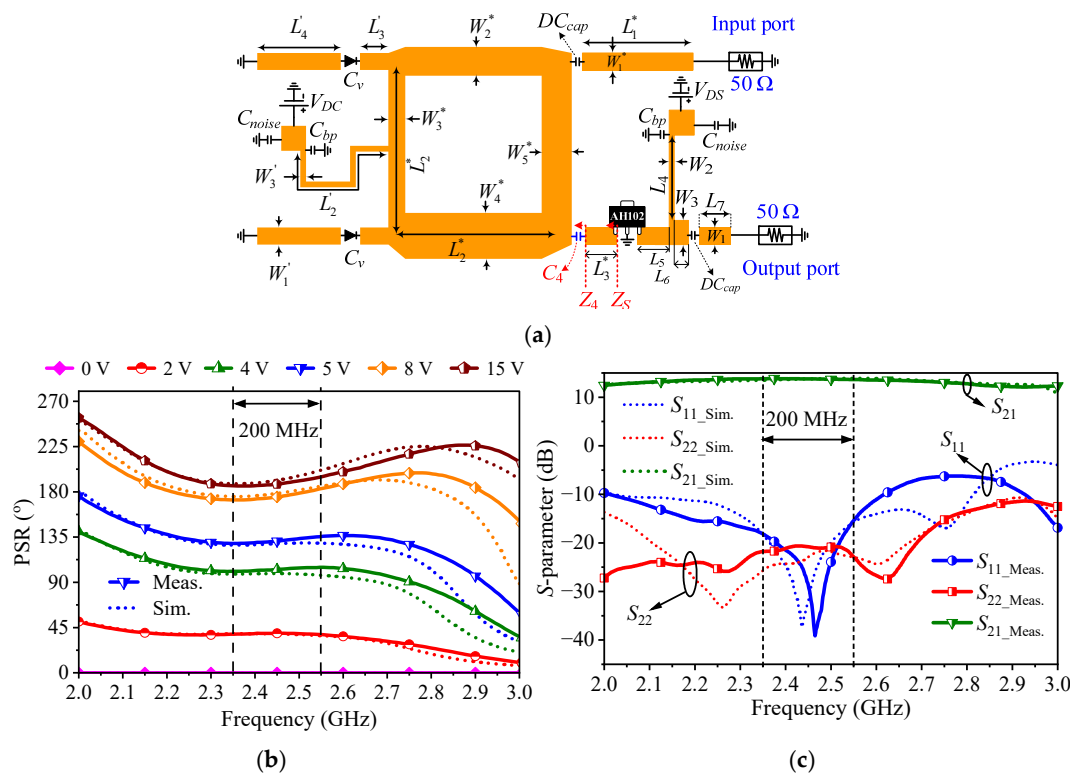


Figure 12. Proposed co-design microwave amplifier with arbitrary termination impedance reflection-type phase shifter: (a) realized structure with dimensions, (b) phase shift range, and (c) S-parameters. Circuit dimensions: $W_1^* = 2.38$; $W_2^* = 3.95$; $W_3^* = 2.38$; $W_4^* = 6.4$; $W_5^* = 4.25$; $L_1^* = 5$; $L_2^* = 23.2$; $L_3^* = 2$ [unit: millimeter] (PSR: phase shift range).

Figure 12b,c present a detailed comparison of phase shift range and S-parameters between the simulated and measured results of the proposed co-design microwave amplifier–phase shifter circuit at f_0 of 2.45 GHz. Figure 12b shows the phase shift as a function of frequency for various DC bias voltages from 0 V to 15 V, which demonstrate that the measured phase increases with the increasing of bias voltage and reaches a maximum of 185.62° at 15 V. Notably, the phase shift range has been observed to exhibit consistent behavior across the BW of 200 MHz with a maximum in-band phase deviation error of $\pm 4.3^\circ$. The results reveal that our proposed structure exhibits stability and tunability. The corresponding S-parameters across the frequency range from 2 GHz to 3 GHz are displayed in Figure 12c. In 200 MHz, the maximum return losses (S_{11} and S_{22}) of 16 dB and the minimum gain (S_{21}) of 13.74 dB are achieved.

3.3. Results Comparison

Figure 13 shows photographs of prototypes of a standalone microwave amplifier, a proposed co-design microwave amplifier with a reflection-type phase shifter input matching network, and a conventional reflection-type phase shifter cascaded with a conventional microwave amplifier. In this work, capacitors from Tekelec are used and are listed in Table 3. Meanwhile, the measured S-parameters of all fabricated prototypes are shown in Figure 14a, where the gains of the standalone conventional microwave amplifier, the conventional reflection-type phase shifter cascaded with conventional microwave amplifier, and the proposed co-design microwave amplifier–phase shifter are 14.13 dB, 13.28 dB, and 13.74 dB, respectively. As can be seen from these results, the proposed co-design microwave amplifier–phase shifter exhibits a higher gain across the entire frequency range than the conventional reflection-type phase shifter cascaded with a conventional microwave amplifier, highlighting its effectiveness in enhancing gain performance. For output power

measurements, a continuous-wave signal is used. Figure 14b shows the measured output power and gain of all fabricated microwave amplifiers according to input power (P_{in}). By increasing P_{in} from -15 dBm to 15 dBm, the output power reached its 1 dB compression point (P_{1dB}). At f_0 , the measured output power at P_{1dB} of the conventional microwave amplifier is 25.72 dBm. Similarly, the P_{1dB} of conventional reflection-type phase shifter cascaded with the conventional microwave amplifier and the proposed co-design microwave amplifier–phase shifter are 24.77 dBm and 25.26 dBm, respectively. At P_{1dB} , the gains of the standalone conventional microwave amplifier, the conventional reflection-type phase shifter cascaded with the conventional microwave amplifier, and the proposed co-design microwave amplifier–phase shifter are 13.11 dB, 12.25 dB, and 12.73 dB, respectively. Even though the conventional microwave amplifier exhibits the highest gain, this circuit does not consider phase shift functionality. The phase shift range and phase deviation error performance comparison of the conventional reflection-type phase shifter cascaded with conventional microwave amplifier and the proposed co-design microwave amplifier phase shifter can be found in Table 4.

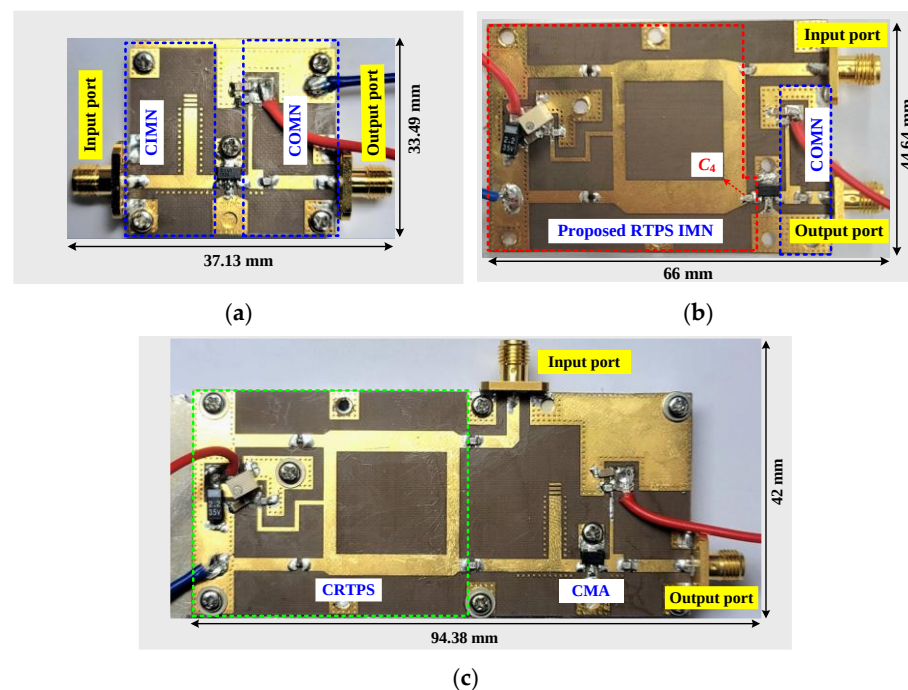


Figure 13. Photograph of fabricated microwave amplifiers: (a) conventional microwave amplifier, (b) proposed co-design microwave amplifier with reflection-type phase shifter input matching network, and (c) conventional reflection-type phase shifter cascaded with conventional microwave amplifier (IMN: input matching network; CIMN: conventional IMN; COMN: conventional output matching network; RTPS: reflection-type phase shifter; CRTPS: conventional RTPS; CMA: conventional microwave amplifier).

Table 3. Parameters of capacitors used in Figures 10a, 11a and 12a.

Target	Part Number of Used Capacitors	Measured Capacitances at 2.45 GHz
$C_{noise} = 0.1$ pF	VJ0603D0R1VXPAJ	$C_{noise_meas.} = 0.13$ pF
$C_{bp} = 33$ pF	0603N8R2AW251	$C_{bp_meas.} = 32.92$ pF
$DC_{cap} = 20$ pF	0603N5R6AW251	$DC_{cap_meas.} = 20.1$ pF
$C_4 = 4.87$ pF	0603N3R3AW251	$C_{4_meas.} = 4.9$ pF

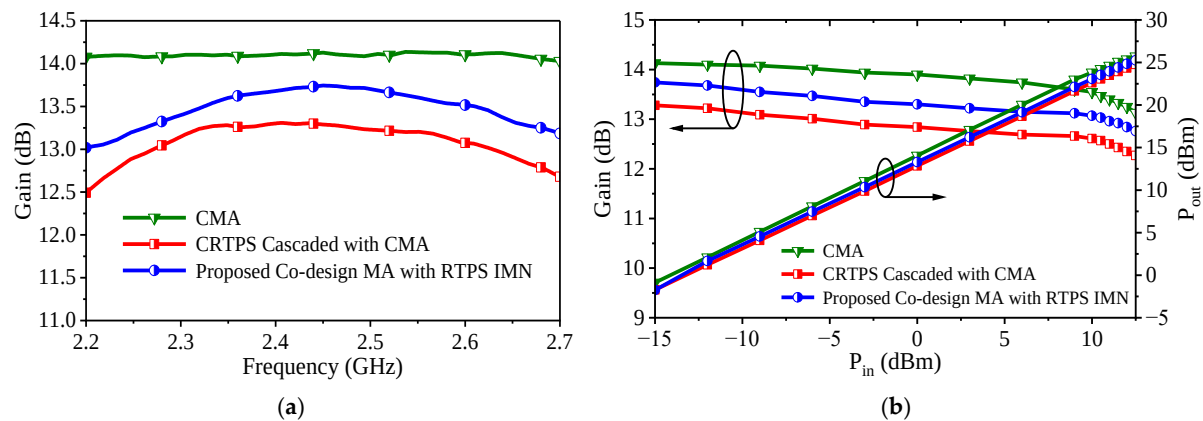


Figure 14. Comparison results of three fabricated microwave amplifiers: (a) gain and (b) P_{out} and gain according to P_{in} (CMA: conventional microwave amplifier; RTPS: reflection-type phase shifter; CRTPS: conventional RTPS; MA: microwave amplifier; CMA: conventional MA; IMN: input matching network).

Table 4. Electrical performances summary of three fabricated microwave amplifiers above a 200 MHz bandwidth (MA: microwave amplifier; PSR: phase shift range; PDE: phase deviation error; CRTPS: conventional reflection-type phase shifter; CMA: conventional microwave amplifier; IMN: input matching network; MA-PS: microwave amplifier–phase shifter).

MAs	Gain (dB)	P_{1dB} (dBm)	PSR ($^{\circ}$)	PDE ($^{\circ}$)	Size
CMA	14.13	25.72	–	–	$0.42 \lambda_0 \times 0.37 \lambda_0$
CRTPS cascaded with CMA	13.28	24.77	195.35	± 6.14	$1.06 \lambda_0 \times 0.47 \lambda_0$
Proposed co-design MA-PS	13.74	25.26	185.62	± 4.3	$0.74 \lambda_0 \times 0.50 \lambda_0$
			201.17	± 6.14	

Table 4 summarizes the key performance metrics of the three fabricated prototypes. The proposed co-design circuit demonstrates superior gain and power compared to the conventional cascaded design. Within 200 MHz BW, the proposed co-design microwave amplifier–phase shifter exhibits a maximum phase shift range of 185.62° with a phase deviation error of $\pm 4.3^{\circ}$ while the conventional reflection-type phase shifter cascaded with a conventional microwave amplifier provides a maximum phase shift range of 195.35° with a phase deviation error of $\pm 6.14^{\circ}$. This result indicates that the proposed co-design microwave amplifier–phase shifter provides a lower phase deviation error than a conventional reflection-type phase shifter cascaded with a conventional microwave amplifier. If a phase deviation error of $\pm 6.14^{\circ}$ within the operating BW is used, the proposed co-design microwave amplifier–phase shifter provides a higher phase shift range and a wider BW than the conventional reflection-type phase shifter cascaded with a conventional microwave amplifier. Besides this, the proposed co-design microwave amplifier with the reflection-type phase shifter input matching network also achieves an approximate size reduction of 38.5%. These improved performances and compact sizes demonstrate the advantages of the co-design method, rather than cascading individual circuits.

4. Conclusions

This work introduces a co-design methodology for a microwave amplifier–phase shifter. By integrating the reflection-type phase shifter as the input matching network of the microwave transistor, the optimum source impedance of the microwave transistor matches 50Ω directly. For experimental validation, a standalone conventional microwave amplifier, a conventional reflection-type phase shifter cascaded with a conventional microwave amplifier, and a proposed co-design microwave amplifier–phase shifter were designed,

fabricated, and their results were compared at f_0 of 2.45 GHz. The co-designed structure exhibits a measured gain of 13.74 dB and P1dB of 25.26 dBm. At 200 MHz, the designed phase shifter achieves a phase shift range of 185.62° with a phase deviation error of $\pm 4.3^\circ$. The proposed co-design microwave amplifier–phase shifter not only offers higher gains and better output power performances but it also has a 38.5% smaller size and less design complexity than the typical conventional reflection-type phase shifter cascaded with a conventional microwave amplifier. The proposed topology is effectively suited for the compact and high electrical performances of RF front-end transmitter systems.

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Abbreviations

The following abbreviations are used in this manuscript:

ADS	Advanced Design System
BW	Bandwidth
HFSS	High-Frequency Structure Simulator
IL	Insertion Loss
MIMO	Multiple Input and Multiple Output
RF	Radio Frequency
RL	Return Loss
TL	Transmission Line

References

1. Pech, P.; Kim, P.; Jeong, Y. Co-design of a low noise amplifier with compact substrate integrated waveguide bandpass filtering matching network. *Int. J. RF Microw. Comput. Aided Eng.* **2022**, *32*, e23498. [\[CrossRef\]](#)
2. Su, Z.; Yu, C.; Tang, B.; Liu, Y. Bandpass filtering power amplifier with extended band and high efficiency. *IEEE Microw. Wirel. Compon. Lett.* **2020**, *30*, 181–184. [\[CrossRef\]](#)
3. Chang, C.-H.; Chen, J.-Y.; Shen, C.-T.; Tsai, M.-J.; Tai, T.-S. Reflection-type phase shifter integrated with tunable power attenuation mechanism for sub-6 GHz wireless applications. *IEEE Access.* **2022**, *10*, 115532–115540. [\[CrossRef\]](#)
4. Zheng, Z.Z.; Zhang, Y.-M.; Zheng, S.Y. A continuously tunable wideband differential phase shifter with filtering function and passband alignment. *IEEE Trans. Microw. Theory Techno.* **2025**, *73*, 620–630. [\[CrossRef\]](#)
5. Zhang, Z.; Psychogiou, D. Multifunctional bandpass filter with codesigned tunable attenuator and reflectionless phase shifter functionalities. *IEEE Microw. Wirel. Technol. Lett.* **2024**, *34*, 737–740. [\[CrossRef\]](#)
6. Lyu, Y.-P.; Zhu, L.; Cheng, C.-H. Proposal and synthesis design of differential phase shifters with filtering function. *IEEE Trans. Microw. Theory Techn.* **2017**, *65*, 2906–2917. [\[CrossRef\]](#)

7. Moon, S.-M.; Yun, S.; Yom, I.-B.; Lee, H.L. Phase array shaped-beam satellite antenna with boosted-beam control. *IEEE Trans. Antennas Propag.* **2019**, *67*, 7633–7636. [\[CrossRef\]](#)
8. Lee, W.-S.; Song, H.-K. Efficient channel feedback scheme for multi-user MIMO hybrid beamforming systems. *Sensors* **2021**, *21*, 5298. [\[CrossRef\]](#)
9. Ayed, A.B.; Jaffri, I.; Darwish, A.M.; Mitran, P.; Boumaiza, S. Proof-of-concept of millimeter-wave RF beamforming transmitter architecture employing frequency-multiplier-based up-converters. *IEEE Microw. Wirel. Compon. Lett.* **2022**, *32*, 776–779. [\[CrossRef\]](#)
10. Kim, S.; Park, H.; Min, B.-W. A center-fed beam-steerable series antenna array with a wide matching bandwidth. *IEEE Open J. Antennas Propag.* **2024**, *5*, 879–887. [\[CrossRef\]](#)
11. Zheng, S.Y.; Chan, W.S.; Man, F.K. Broadband phase shifter using loaded transmission line. *IEEE Microw. Wirel. Compon. Lett.* **2010**, *20*, 498–500. [\[CrossRef\]](#)
12. Gaun, C.-E.; Kanaya, H. 360° Phase shifter design using dual-branch switching network. *IEEE Microw. Wirel. Compon. Lett.* **2018**, *28*, 675–677. [\[CrossRef\]](#)
13. Burdin, F.; Iskandar, Z.; Podevin, F.; Ferrari, P. Design of compact reflection-type phase shifters with high figure-of-merit. *IEEE Trans. Microw. Theory Techn.* **2015**, *63*, 1883–1893. [\[CrossRef\]](#)
14. Yang, G.; Lee, D.; Min, B.-W. Low delay and loss variation reflection type phase shifter with sequentially controlled varactor diodes. In Proceedings of the 2023 IEEE/MTT-S International Microwave Symposium-IMS 2023, San Diego, CA, USA, 11–16 June 2023; pp. 1050–1053.
15. Saron, S.; Chaudhary, G.; Jeong, Y. Concurrent dual-band shifter with independent tunable phase shift control. *IEEE Microw. Wirel. Technol. Lett.* **2024**, *34*, 163–166. [\[CrossRef\]](#)
16. Liu, W.J.; Zheng, S.Y.; Pan, Y.M.; Li, Y.X.; Long, Y.L. A wideband tunable reflection-type phase shifter with wide relative phase shift. *IEEE Trans. Circuits Syst. II Exp. Briefs.* **2017**, *64*, 1442–1446. [\[CrossRef\]](#)
17. An, B.; Chaudhary, G.; Jeong, Y. Wideband tunable phase shifter with low in-band phase deviation using coupled line. *IEEE Microw. Wirel. Compon. Lett.* **2018**, *28*, 678–680. [\[CrossRef\]](#)
18. Chaudhary, G.; An, B.; Jeong, Y. In-band phase deviation minimization method for wideband tunable phase shifter. *Microw. Opt. Technol. Lett.* **2018**, *61*, 537–541. [\[CrossRef\]](#)
19. Chaudhary, G.; Jeong, Y. Wideband tunable differential phase shifter with minimized in-band phase deviation error. *IEEE Microw. Wirel. Compon. Lett.* **2019**, *29*, 468–470. [\[CrossRef\]](#)
20. Amin, F.; Liu, L.; Liu, Y.; Zhao, Y. Compact single-unit two-bit reflection-type phase shifters with large phase shift range. *Electronics* **2023**, *12*, 2412. [\[CrossRef\]](#)
21. Singh, A.; Mandal, M.K. Electronically tunable reflection type phase shifters. *IEEE Trans. Circuits Syst. II Exp. Briefs.* **2020**, *67*, 425–429. [\[CrossRef\]](#)
22. Zhu, H.; Abbosh, A.M. Wideband tunable reflection-type phase shifter using high-directivity directional coupler. In Proceedings of the 2018 Australian Microwave Symposium (AMS), Brisbane, Australia, 6–7 February 2018; pp. 49–50.
23. Lyu, Y.-P.; Zhu, L.; Wang, S.-T.; Cheng, C.-H. A new synthesis method to design broadband continuously tunable phase shifters based on coupled line with low phase error. *IEEE Trans. Microw. Theory Techn.* **2025**, *73*, 2869–2878. [\[CrossRef\]](#)
24. Ahn, H.-R. *Asymmetric Branch-Line Hybrids in Asymmetric Passive Component in Microwave Integrated Circuits*; John Wiley & Sons, Inc.: Hoboken, NJ, USA, 2006; pp. 141–144.
25. Kameche, M.; Feham, M.; Meliani, M.; Benahmed, N.; Dali, S. *Intrinsic Small-Signal Equivalent Circuit of GaAs MESFET's*; National Centre of Space Techniques, AlgeriaTelecom Laboratory, University of Tlemcen: Tlemcen, Algeria, 2006.
26. Pozar, D.M. Power dividers and directional coupler. In *Microwave Engineering*, 4th ed.; John Wiley & Sons, Inc.: Hoboken, NJ, USA, 2012; pp. 343–346.

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