



Ultra-wideband miniaturized Wilkinson power divider based on rectangular multi-section resonators with neural network-based enhancement

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ABSTRACT

In this study, a novel compact Wilkinson power divider (PD) with high fractional bandwidth (FBW) and excellent isolation is proposed, designed, and experimentally validated. The structure employs rectangular multi-section resonators, replacing conventional transmission lines, achieving a significant size reduction of over 76% compared to traditional designs. Additionally, the multilayer perceptron (MLP) neural network algorithm is used to optimize performance, resulting in a very small mean relative error (MRE) and mean absolute error (MAE), indicating the design's optimality. The fabricated PD eliminates 16 unwanted harmonics with suppression levels below -20 dB across a wide frequency range of 1.15–2.5 GHz. Fabricated on an RT Duroid 5880 substrate, the measured results show excellent agreement with simulations, validating the design's reliability. Compared to existing Wilkinson power dividers, the proposed solution offers several key advantages: a 76% reduction in size, a FBW of 96%, and the suppression of harmonics up to the 16th order, which significantly enhances signal integrity. Additionally, the design's simplicity and efficiency ensure lower manufacturing costs, making it suitable for high-performance, cost-effective wireless communication systems.

1. Introduction

In recent decades, the development of high-frequency telecommunication circuits has played an important role in modern society by revolutionizing communication, information exchange, and technological progress. These circuits are at the heart of various devices and systems that enable fast and efficient data transmission over long distances [1].

Power dividers and couplers are now an integral part of high-frequency systems and circuits, and the applications of power dividers, include wireless and Bluetooth communication systems, mobiles, phase shifters, biomedical instruments, antenna arrays, etc. This is because one of the main tasks of these passive circuits is to divide or combine the power of signals at radio frequencies [1–3]. One of the common PDs is the Wilkinson PD, which has been highly regarded by designers and engineers due to its small dimensions and high isolation capability. Thus, several methods have been used to improve the PD's performance. These include, but are not limited to, the following: creating wideband filters and PDs utilizing a defective ground structure [4]; using spiral-

coupled-lines [5]; using parallel resonant shunt stubs [6]; using cascading sections [7]; using tapered transmission lines (TLs) [8]; etc.

As shown in [9], a two-stage wideband Wilkinson PD utilizing quarter-wave TLs and short-circuit stubs is achieved. Furthermore, the given structure was analyzed using the even- and odd-mode analysis approach, which produced an appropriate FBW; however, its dimensions are extremely large, measuring more than 66 cm^2 . To generate transmission zeros and increase the stopband bandwidth of the PD, circular patch-shaped resonators are employed in [10]. This arrangement eliminates up to six deleterious harmonics at a power division frequency of 2.55 GHz.

In [11], the quarter-wavelength lines of the conventional structure are replaced by a stub-loaded TL to reduce the size of the circuit. Also, the defected ground structure technique is used to eliminate higher-order harmonics. This circuit suffers from inadequate insertion losses. The utilization of two-part resonators connected to the input and output ports and stepped-impedance TLs allows for the suppression of harmonics up to the ninth order in [12]. Although the structure's insertion losses are low, the isolation between its output ports is poor.

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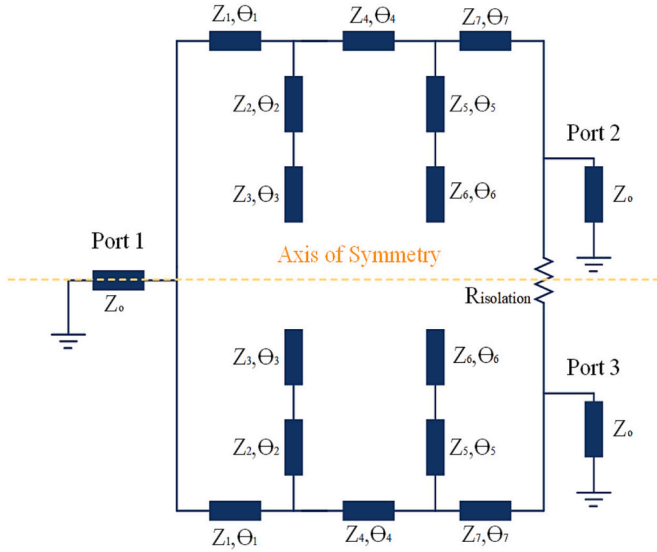


Fig. 1. The final structure of the designed Wilkinson PD.

The PD described in [13] is based on an asymmetric metastructure and has a basic and symmetrical structure, which makes it simple to build and operate. It only eliminates four unwanted harmonics. For accurate impedance transformation and more design flexibility in power dividers, coupled microstrip lines with capacitor-loaded and two short-circuited ports are shown in [14]. In [15], a comprehensive study of odd and even modes was utilized to compute the characteristic impedances and electrical lengths of TLs. Unfortunately, the bandwidth of the structure is limited, and no harmonics were eliminated.

Some researchers have focused on the mathematical analysis of PD circuits. Ref. [16] uses the particle swarm optimization algorithm; Refs. [17,18] use neural network algorithms; and Ref. [19] uses genetic algorithms.

In [20], using an optimization technique and filter synthesis theory, the suggested topology's scattering parameters are fine-tuned to provide an equal-ripple response. Its equal-ripple level and bandwidth allow for controllable outcomes. Particle swarm optimization was used to find the best possible values for the electrical lengths of TLs in [21]. In addition, to construct a structure with compact dimensions, the typical admittances have been computed. Significant input return loss and poor isolation are some of the circuit's drawbacks.

Creating a PD with sufficient insertion loss, a small size, a high FBW, and a simple topology has been a major challenge for designers.

The design and realization of a unique structure of ultra-small Wilkinson PD based on an MLP neural network employing rectangular multi-section resonators is presented in this study. Our PD has a compact size, a wide stopband, high-order harmonic suppression, and high isolation. This PD is designed based on even- and odd-mode analysis for TLs. To check the accuracy of the design, the dimensions (length and width) of the resonator, its number, and the total length of the TL have been applied as inputs to the MLP network, resulting in very small MRE and MAE have been obtained for the outputs of insertion loss, return loss, and isolation. This power divider operates at a frequency of 1.4 GHz, and its dimensions are only 9.04 mm × 11.4 mm. It exhibits favorable isolation and wideband performance, reflecting its suitability for modern microwave applications.

As a result, the novelty of the proposed PD lies in its ability to achieve a highly compact (76% size reduction) using newly introduced trapezoidal-shaped microstrip resonators capable of suppressing 16 harmonics below -20 dB. In contrast to other approaches, which used defected ground structures (DGS) [4], stub-loaded TLs [21], or multi-stage quarter-wave sections [9], the proposed structure achieves simultaneous miniaturization, FBW of 96%, high isolation, and deep

stopband attenuation over a wide range. Furthermore, an MLP neural network verification framework is used to accurately predict S-parameters from physical design variables, ensuring design reliability, as confirmed by measurement results. So, the presented design provides a compact and efficient solution for power splitting and combining in RF systems.

2. Design of Wilkinson PD

The structure of the TL of the newly designed PD is presented in Fig. 1. This circuit consists of seven TLs in each branch, and according to the original Wilkinson structure, an isolation resistor is placed between the two output ports. It should be noted that in the design of this circuit, impedance matching of 50 Ohms (Z_0) has been used for the output and input ports. Furthermore, each TL consists of a characteristic impedance and an electrical length. This structure is designed symmetrically, and the TLs of opposite branches are equal.

The symmetry line divides the structure into two mirrored halves, each containing identical components and configurations. The symmetry ensures balanced operation in the Wilkinson power divider, where equal power is delivered to the output ports (Port 2 and Port 3) with proper isolation between them.

The characteristic impedances Z_1 to Z_7 are fundamental parameters in the design of the PD, directly influencing its performance in terms of power division, harmonic suppression, and isolation. These values are determined by employing standard transmission line theory, which relates the physical dimensions of the transmission lines and resonators to their impedance. The characteristic impedance Z of a microstrip line, which forms the backbone of the Wilkinson PD, depends critically on the line width (w) and substrate height (h), as well as the dielectric properties of the substrate. These parameters are crucial for ensuring optimal impedance matching, high isolation between output ports, and suppression of unwanted harmonics.

The effective relative permittivity ϵ_{re} of the microstrip line, which is required to calculate the impedance, is highly dependent on the ratio of the line width to the substrate height (w/h). This ratio determines whether the trace is narrow or wide relative to the substrate, influencing both the effective permittivity and the characteristic impedance. For narrow traces ($w/h \leq 1$), the effective permittivity ϵ_{re} is given by [1,2]:

$$\epsilon_{re} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \times \left\{ \frac{1}{\sqrt{1 + 12 \frac{h}{w}}} + 0.04 \left[1 - \frac{w}{h} \right]^2 \right\} \quad (1)$$

where ϵ_r is the relative permittivity of the substrate, and h and w are the height and width of the microstrip line, respectively. The characteristic impedance in this case is calculated using [1,2]:

$$Z = \frac{\eta}{2\pi\sqrt{\epsilon_{re}}} \times \ln \left[8 \frac{h}{w} + 0.25 \frac{w}{h} \right] \quad (2)$$

For wide traces ($w/h \geq 1$), a different approach is used to calculate ϵ_{re} and Z , reflecting the changing behavior of the line [1,2]:

$$\epsilon_{re} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left\{ \left[1 + 12 \frac{h}{w} \right]^{-0.5} \right\} \quad (3)$$

$$Z = \frac{\eta}{\sqrt{\epsilon_{re}}} \times \left\{ \frac{h}{w} + 1.393 + 0.677 \times \ln \left[\frac{w}{h} + 1.444 \right] \right\}^{-1} \quad (4)$$

These equations provide an accurate estimation of the microstrip line's impedance, enabling the design of efficient transmission lines for the Wilkinson PD.

Also, even- and odd-mode analysis is used to analyze the structure. In this way, the input signal (V_{in}) is considered first. This signal is divided into two branches and enters each node. By solving the electromagnetic

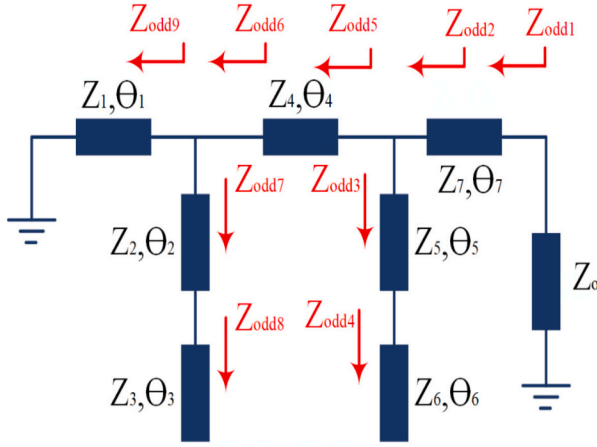


Fig. 2. The odd-mode equivalent circuit.

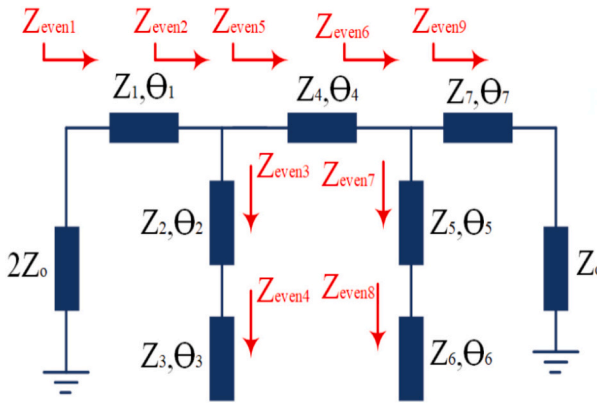


Fig. 3. The even-mode equivalent circuit.

field equations for each node, the current and voltage in each node are obtained. These values are then used to calculate the reflection and transmission coefficients.

2.1. Odd mode

When the odd mode is excited in our Wilkinson PD, the voltage at the center of the circuit becomes zero. This phenomenon is based on the symmetry of the circuit under odd-mode excitation, where the signal is equally distributed across both branches but in opposite phases. As a result, the central node can be grounded without affecting the circuit's performance, allowing the entire circuit to be split symmetrically into two halves for analysis.

The odd-mode equivalent circuit is shown in Fig. 2. This equivalent circuit facilitates a systematic analysis of the TL sections by representing their electrical behavior in terms of impedances and phase angles. Each TL section is characterized by its characteristic impedance (Z_i) and electrical length (θ_i). The step-by-step calculation of the equivalent impedance under odd-mode excitation proceeds as follows:

$$Z_{\text{odd}9} = jZ_1 \tan \theta_1 \quad (5)$$

$$Z_{\text{odd}8} = -jZ_3 \cot \theta_3 \quad (6)$$

$$Z_{\text{odd}7} = Z_2 \left(\frac{Z_{\text{odd}8} + jZ_2 \tan \theta_2}{Z_2 + jZ_{\text{odd}8} \tan \theta_2} \right) \quad (7)$$

The impedance at this stage ($Z_{\text{odd}6}$) is obtained by combining $Z_{\text{odd}7}$ and $Z_{\text{odd}9}$ in parallel:

$$Z_{\text{odd}6} = \frac{Z_{\text{odd}7} \times Z_{\text{odd}9}}{Z_{\text{odd}7} + Z_{\text{odd}9}} \quad (8)$$

The impedance calculations proceed through successive sections of the circuit:

$$Z_{\text{odd}5} = Z_4 \left(\frac{Z_{\text{odd}6} + jZ_4 \tan \theta_4}{Z_4 + jZ_{\text{odd}6} \tan \theta_4} \right) \quad (9)$$

$$Z_{\text{odd}4} = -jZ_6 \cot \theta_6 \quad (10)$$

$$Z_{\text{odd}3} = Z_5 \left(\frac{Z_{\text{odd}4} + jZ_5 \tan \theta_5}{Z_5 + jZ_{\text{odd}4} \tan \theta_5} \right) \quad (11)$$

$$Z_{\text{odd}2} = \frac{Z_{\text{odd}3} \times Z_{\text{odd}5}}{Z_{\text{odd}3} + Z_{\text{odd}5}} \quad (12)$$

Finally, the impedance at the input port ($Z_{\text{odd}1}$) is calculated as:

$$Z_{\text{odd}1} = Z_7 \left(\frac{Z_{\text{odd}2} + jZ_7 \tan \theta_7}{Z_7 + jZ_{\text{odd}2} \tan \theta_7} \right) \quad (13)$$

The reflection coefficient at the output ports under odd-mode excitation ($\Gamma_{\text{out}}^{\text{odd}}$) is determined by comparing the calculated input impedance ($Z_{\text{odd}1}$) with the reference impedance (Z_0):

$$\Gamma_{\text{out}}^{\text{odd}} = \frac{Z_{\text{odd}1} - Z_0}{Z_{\text{odd}1} + Z_0} \quad (14)$$

Each formula (Eqs. 1–9) is derived from fundamental transmission line theory, using the relationships for impedance transformation and reflection coefficients. These calculations ensure proper matching at the output ports and provide the basis for deriving the odd-mode reflection coefficient ($\Gamma_{\text{out}}^{\text{odd}}$) in Eq. (10). The odd-mode analysis guarantees isolation between the output ports while maintaining consistent signal splitting.

The odd-mode analysis not only ensures proper impedance matching at the output ports but also provides insight into the isolation characteristics of the proposed PD. The calculated odd-mode impedance values guide the design and optimization of the circuit, ensuring high isolation and minimal reflections in odd-mode operation.

2.2. Even mode

When the even mode is excited in the proposed Wilkinson power divider, the circuit behaves symmetrically. This symmetry allows the analysis to be simplified by effectively dividing the circuit along its horizontal axis. Under even-mode excitation, the isolation resistor does not conduct current because the voltage difference across it is zero. Consequently, the resistor does not affect the even-mode behavior of the circuit. In this analysis, each branch of the transmission line network can be modeled independently. The input signal is equally split between the two branches, and the voltages at symmetrical points in the circuit are identical. The equivalent circuit for the even mode is shown in Fig. 3, where the transmission line segments are represented by their characteristic impedances (Z_i) and electrical lengths (θ_i).

The following is the calculation that is used to obtain the equivalent impedance of the output in even mode:

$$Z_{\text{even}9} = Z_7 \left(\frac{Z_0 + jZ_7 \tan \theta_7}{Z_7 + jZ_0 \tan \theta_7} \right) \quad (15)$$

$$Z_{\text{even}8} = -jZ_6 \cot \theta_6 \quad (16)$$

$$Z_{\text{even}7} = Z_5 \left(\frac{Z_{\text{even}8} + jZ_5 \tan \theta_5}{Z_5 + jZ_{\text{even}8} \tan \theta_5} \right) \quad (17)$$

$$Z_{\text{even}6} = \frac{Z_{\text{even}7} \times Z_{\text{even}9}}{Z_{\text{even}7} + Z_{\text{even}9}} \quad (18)$$

Table 1
Impedance values of transmission lines in the proposed Wilkinson power divider.

Impedance Parameter	Value (Ω)
Z_1	75.2
Z_2	120.8
Z_3	90.4
Z_4	135.5
Z_5	110.4
Z_6	163.3
Z_7	80.8

$$Z_{\text{even5}} = Z_4 \left(\frac{Z_{\text{even6}} + jZ_4 \tan \theta_4}{Z_4 + jZ_{\text{even6}} \tan \theta_4} \right) \quad (19)$$

$$Z_{\text{even4}} = -jZ_3 \cot \theta_3 \quad (20)$$

$$Z_{\text{even3}} = Z_2 \left(\frac{Z_{\text{even4}} + jZ_2 \tan \theta_2}{Z_2 + jZ_{\text{even4}} \tan \theta_2} \right) \quad (21)$$

$$Z_{\text{even2}} = \frac{Z_{\text{even3}} \times Z_{\text{even5}}}{Z_{\text{even3}} + Z_{\text{even5}}} \quad (22)$$

$$Z_{\text{even1}} = Z_1 \left(\frac{Z_{\text{even2}} + jZ_1 \tan \theta_1}{Z_1 + jZ_{\text{even2}} \tan \theta_1} \right) \quad (23)$$

According to Eq. (19), the reflection coefficient obtained at the input port or parameter S_{11} can be obtained as follows:

$$S_{11} = \Gamma_{\text{in}}^{\text{even}} = \frac{Z_{\text{even1}} - 2Z_0}{Z_{\text{even1}} + 2Z_0} \quad (24)$$

It is also possible to derive the parameter S_{23} of the proposed PD by using the formula [22]:

$$S_{23} = \frac{\Gamma_{\text{out}}^{\text{even}} - \Gamma_{\text{out}}^{\text{odd}}}{2} \quad (25)$$

Furthermore, the return loss (RL) and isolation (I) can be written as:

$$\text{RL(dB)} = -20 \log |S_{11}| \quad (26)$$

$$I(\text{dB}) = -20 \log |S_{23}| \quad (27)$$

In the context of the proposed Wilkinson PD, the values of Z_1 to Z_7 are optimized to meet the design requirements for equal power division, high isolation, and harmonic suppression. Specifically, Z_1 , Z_2 , and Z_3 represent the impedances of the main transmission lines that split the power between the input and the output ports. These impedances are carefully calculated to achieve a balanced power division at the operating frequency, ensuring that the power is split evenly between the output ports. Furthermore, Z_4 , Z_5 , and Z_6 correspond to the resonators integrated into the design to suppress higher-order harmonics. These resonators are designed to have specific impedance values that enhance the stopband performance, effectively reducing spurious harmonics and improving the overall frequency response of the power divider. Lastly, Z_7 , which is the impedance of the isolation resistor, is optimized to provide adequate isolation between the output ports while minimizing insertion losses.

The impedance values Z_1 to Z_7 are computed using the transmission line formulas mentioned above, along with the physical dimensions of the resonators and transmission lines. These values are crucial for ensuring that the Wilkinson PD meets the desired performance criteria, including high fractional bandwidth, minimal return loss, and

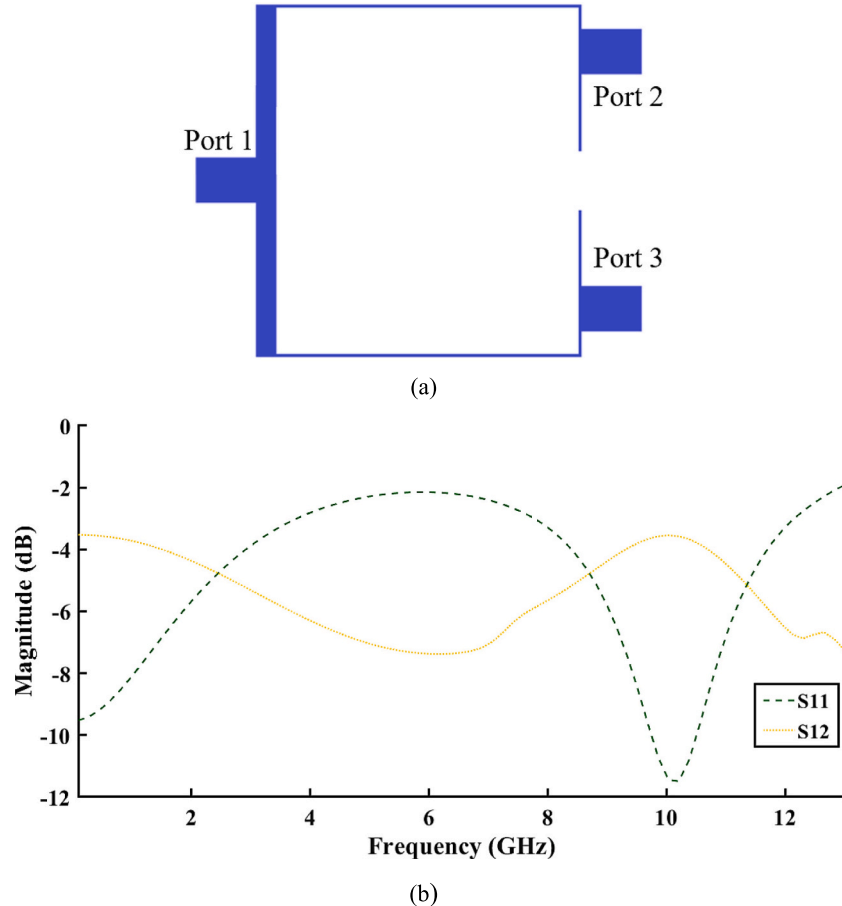


Fig. 4. (a) The basic layout of Wilkinson's PD, (b) Simulation result of S_{12} and S_{11} parameters.

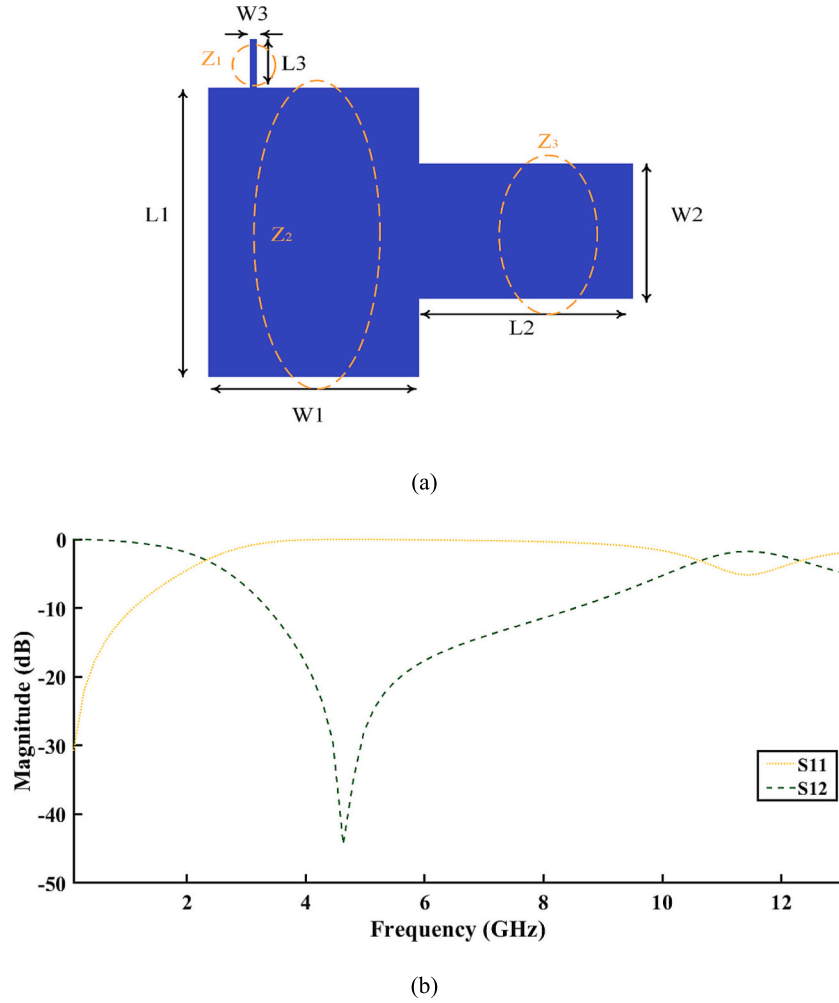


Fig. 5. (a) The layout of first rectangular multi-section resonator, (b) Simulation result of S_{12} and S_{11} parameters.

Table 2

Dimensions of the deigned rectangular multi-section resonators (dimensions in mm).

L1	L2	L3	L4	L5	L6	W1
4.81	3.15	0.8	6.43	3.24	0.25	3.1
W2	W3	W4	W5	W6	W7	
2.24	0.1	4.82	2.63	1.07	0.1	

substantial harmonic suppression.

The final optimized values for Z_1 to Z_7 , obtained through iterative design and optimization, are presented in Table 1. These values correspond to the optimal impedances that balance performance, size, and harmonic suppression in the proposed compact Wilkinson PD.

By adjusting the magnitude of the input reflection coefficient $|S_{11}|$, a wide impedance-matched passband is achieved, with the return loss (RL) better than 21 dB. This adjustment is critical for minimizing signal reflections and maximizing the transmission efficiency. Moreover, when $S_{23} = 0$ at the operating frequency, a transmission zero is created between output ports 2 and 3. This transmission zero allows the isolation frequency band to be extended, further enhancing the separation between the output ports and improving the overall isolation.

Through this optimization process, the impedance values Z_1 to Z_7 are determined to achieve the desired performance of the Wilkinson PD. The resulting values, as shown in Table 1, demonstrate the effectiveness of the design in achieving a compact, high-performance PD with excellent harmonic suppression and isolation.

The formulas and relationships presented in this study for even- and odd-mode analysis of the proposed Wilkinson PD are based on well-established principles of microwave circuit design and transmission line theory. These relationships have been thoroughly derived and verified using fundamental equations documented in Refs. [23–26].

The even-mode analysis assumes that the isolation resistor does not conduct current under even-mode excitation, allowing the circuit to be symmetrically divided and analyzed. The odd-mode analysis considers the circuit's central voltage being zero, allowing equivalent impedances to be derived.

Moreover, the derived expressions, including Z_{even} and Z_{odd} , and their subsequent reflection coefficients, match the outcomes predicted by simulation and experimental validation. The methodologies employed are consistent with previous works and widely accepted practices in the field of PD design.

To ensure the accuracy and reliability of the derived impedance values for our Wilkinson PD, Mathematica software is employed for complex arithmetic and iterative calculations. Mathematica's precision capabilities make it an ideal tool for analyzing complex impedance transformations and verifying results derived from transmission line theory.

In the even-mode analysis, it is essential to note that this method relies on the symmetry of the Wilkinson power divider design. Specifically, the even-mode analysis assumes that the divider is symmetrically configured, ensuring that power is evenly distributed between the output ports. This symmetry allows for simplification in the calculation of the impedance and reflection coefficients. Thus, the even-mode

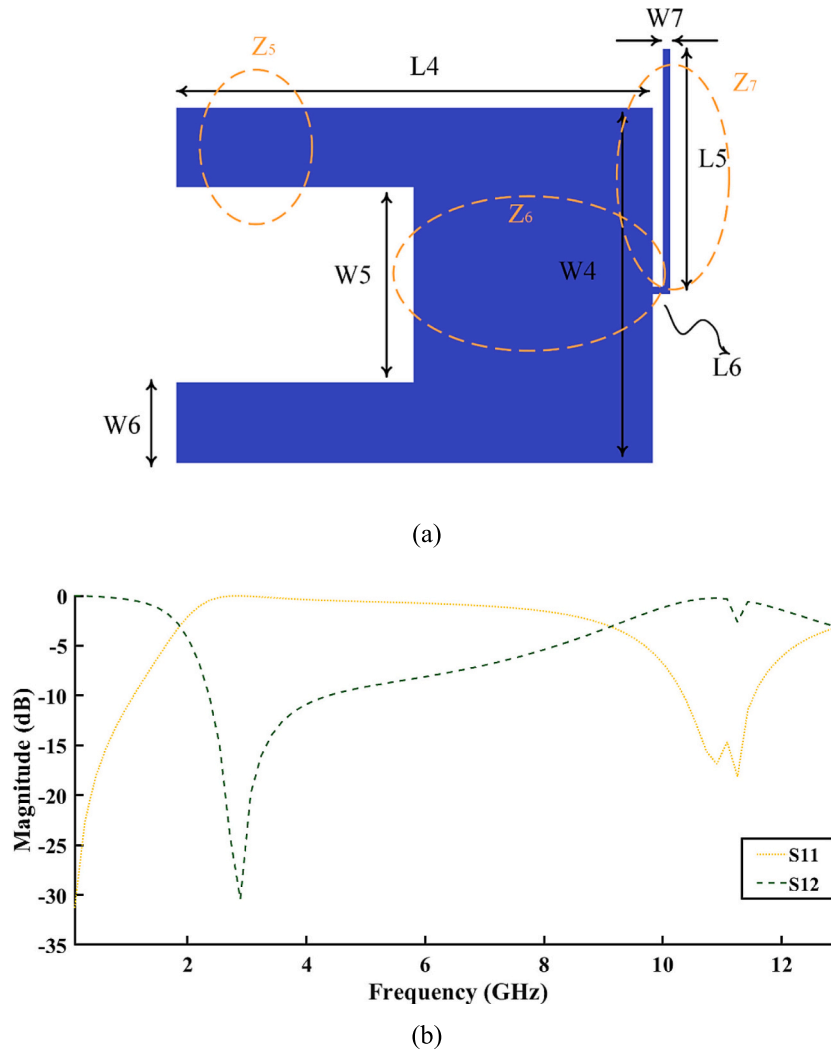


Fig. 6. (a) The layout of second rectangular multi-section resonator, (b) Simulation result of S_{12} and S_{11} parameters.

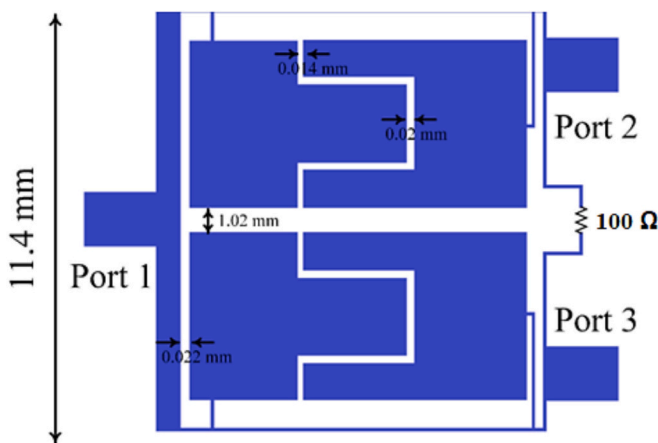


Fig. 7. The final structure of the PD.

analysis is strictly applicable only to power dividers designed for equal power splitting, where both output branches have identical impedances and lengths [27,28].

Similarly, the odd-mode analysis also relies on symmetry in its calculation of impedance values, but it focuses on the behavior of the

circuit when the input signal is applied in opposite phases to the two branches. Therefore, both the odd- and even-mode analysis methods are designed for balanced, symmetric Wilkinson power dividers. For asymmetric power divider configurations, different analysis techniques must be employed to account for the unequal distribution of power and the impact on impedance matching and isolation.

3. The basic layout of the Wilkinson power divider

Fig. 4(a) shows a basic and simple architecture that is utilized to create a Wilkinson PD with a high FBW, a broad cut-off band, and sufficient isolation between the output ports. The RT/Duroid 5880 substrate, with a dielectric constant of 2.2, a loss tangent of 0.0009, and a substrate thickness of 0.508 mm, is used to simulate this basic circuit in ADS software. No resonator or stub is utilized. Fig. 4(b) displays the generated scattering parameter values (S_{12} and S_{11}) from the simulation.

As is clear in Fig. 4(b), this basic structure does not have improved passband and cutoff and does not create any proper power distribution between the output ports. To improve the performance of the proposed PD shown in Fig. 1, resonators corresponding to TLs with impedances Z_1 , Z_2 , Z_3 , Z_5 , Z_6 , and Z_7 are added to this basic layout.

3.1. Design of the first rectangular multi-section resonator

Using microstrip technology and rectangular geometry, the layout of

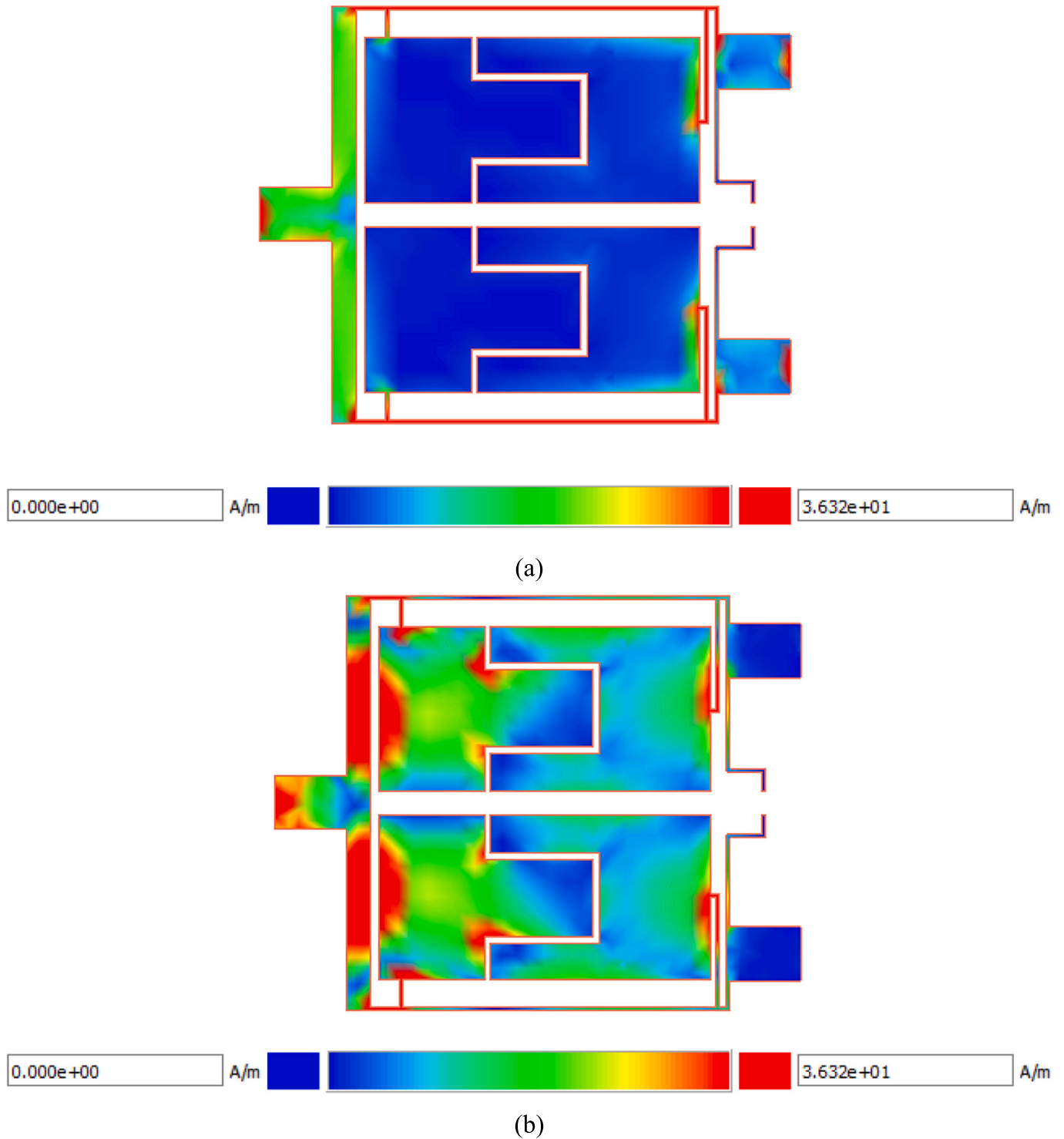


Fig. 8. Current density distribution at (a) 1.4 and (b) 10 GHz frequencies.

the first rectangular multi-section resonator is designed as shown in Fig. 5(a). This resonator corresponds to the impedance values Z_1 , Z_2 , and Z_3 in the Wilkinson PD. The layout of the resonator plays a crucial role in achieving the desired performance by ensuring the proper harmonic suppression and impedance matching.

Fig. 5(a) presents the design of this resonator, which directly affects the impedance characteristics of the PD. The arrangement of the resonator sections and their respective dimensions are optimized to maximize the suppression of unwanted harmonics, particularly at higher frequencies, while ensuring that the power division remains accurate at

the operating frequency.

Additionally, Z_4 represents the characteristic impedance of the transmission line connecting the first and second rectangular multi-section resonators. This transmission line is critical for maintaining proper impedance matching and ensuring efficient signal transfer between the resonators. The value of Z_4 is carefully selected to optimize the overall performance of the Wilkinson PD, contributing to improved isolation and minimal insertion loss.

In Fig. 5(b), the simulation results for the S_{12} and S_{11} parameters are presented, showcasing the power division and return loss characteristics

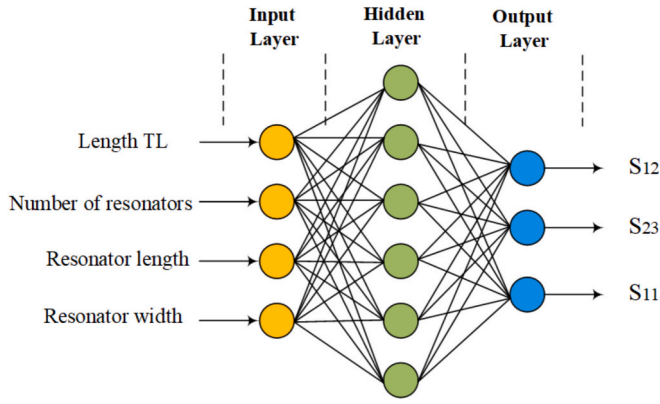


Fig. 9. MLP neural network architecture for the proposed PD.

Table 3

Characteristics of the implemented neural network.

Type of artificial intelligence algorithm	MLP model
The number of neurons in the input layer	4
The number of neurons in the hidden layer	6
The number of neurons in the output layer	3
The number of epochs	2500
Activation function	tansig

Table 4

The obtained errors of the Wilkinson PD.

Output	MRE test error	MRE train error	MAE test error	MAE train error
S_{12}	3.23	3.51	2.67	2.94
S_{23}	1.21	1.85	1.05	1.33
S_{11}	1.24	1.07	2.14	1.41

of the resonator. The S_{12} parameter indicates the transmission efficiency between the input and output ports, while the S_{11} parameter shows the reflection from the input port, which is indicative of the return loss. The simulation results demonstrate that the resonator design achieves a wide operating bandwidth with high isolation and low reflection loss, confirming the effectiveness of the design in meeting the required performance specifications.

In Table 2, the dimensions of the rectangular multi-section resonators are essential in determining the characteristic impedance and overall performance of the Wilkinson PD. The characteristic impedance of each section is derived from the standard transmission line equations, specifically Eqs. (1–4), which relate the physical dimensions of the resonators to the impedance of the microstrip lines.

These equations involve the width of each resonator $W_1, W_2, \dots, W_7$, listed in Table 2, along with the substrate height h and the effective relative permittivity ϵ_{re} of the substrate material. By applying these equations, the characteristic impedance for each transmission line section is calculated, ensuring that the design achieves the desired power division and harmonic suppression.

The result of the simulation of the S parameters of the designed resonator is shown in Fig. 5(b). As expected, this resonator has a transmission zero around 4.85 GHz and a relatively wide cutoff bandwidth. As a result of the fact that its operation is comparable to that of a low-pass filter, the value of the return loss in the pass band is less than -15 dB. This indicates that there is no return loss, which is highly desirable. But, as shown in Fig. 5(b), the use of a resonator does not create a high cut-off bandwidth. Therefore, another resonator is required to provide more bandwidth.

3.2. Design of the second rectangular multi-section resonator

Similar to the first resonator, using microstrip technology and rectangular geometry, the layout of the second rectangular multi-section resonator is designed according to Fig. 6(a). This structure in Fig. 6(a) is equivalent to impedances Z_5 to Z_7 .

The dimensions of the second rectangular multi-section resonator are also taken from Table 2. Fig. 6(b) displays the result of the resonator's S parameter simulation. A transmission zero at 2.92 GHz and a rather large cut-off bandwidth are predicted characteristics of this resonator.

3.3. The final structure of the PD

The equivalency between the rectangular multi-section resonators and the corresponding TL sections has been validated by optimizing the resonator dimensions (as presented in Table 1) and observing their impact on S -parameters. Figs. 5 and 6 clearly illustrate the performance of each resonator, while Fig. 7 demonstrates their integration into the final circuit structure. Simulation and experimental results (Figs. 13(a) and (b)) confirm the effectiveness of the rectangular multi-section resonators in achieving the design objectives.

The distances between the gaps were designed to optimize impedance matching and harmonic suppression. These gap values were determined using transmission line equations and the optimization process to ensure the Wilkinson PD operates efficiently across the desired frequency range, with good isolation and minimal insertion loss. The specific values of the gaps are indicated in Fig. 7, which shows the distances between the transmission line sections and resonators.

As seen in Fig. 7, the final structure of the presented PD consists of two new rectangular multi-section resonators, each of which creates two transmission zeros in the frequency response of the circuit. The overall dimensions of this circuit are $9.04 \text{ mm} \times 11.4 \text{ mm}$, which is very small compared to similar samples. This circuit has achieved a size reduction of more than 76% compared to the conventional PD.

The RF signal is applied to the input port and then transmitted to the bottom and top of the input TL. When the signal reaches the first resonator, it encounters a quarter-wavelength ($\lambda/4$) transmission line section. At this stage, the signal undergoes a 90-degree phase shift corresponding to a quarter wavelength. Next, the two output signals each experience an additional 90-degree phase shift when passing through the second resonator. As a result, the signals from the two output ports have a phase difference of 180 degrees, which means that they are in antiphase with each other. Also, according to the conventional Wilkinson structure, a 100 Ohm isolation resistor is placed between the two output ports.

The electric current distribution profile in the designed PD at two frequencies of 1.4 GHz and 10 GHz is shown in Figs. 8(a) and (b), respectively.

As expected, at the central frequency of 1.4 GHz, a high electric current distribution is observed at the output ports (darker regions). In contrast, at 10 GHz, the designed rectangular multi-section resonators absorb most of the current, effectively preventing it from passing through and reaching the output ports.

In the following, an artificial intelligence neural network is used to check the accuracy of the design.

4. The MLP neural network algorithm

The MLP is a type of artificial neural network commonly used for machine learning tasks such as classification, regression, and pattern recognition. It is a feedforward network where information flows in one direction from the input layer through one or more hidden layers to the output layer, without cycles or loops [29].

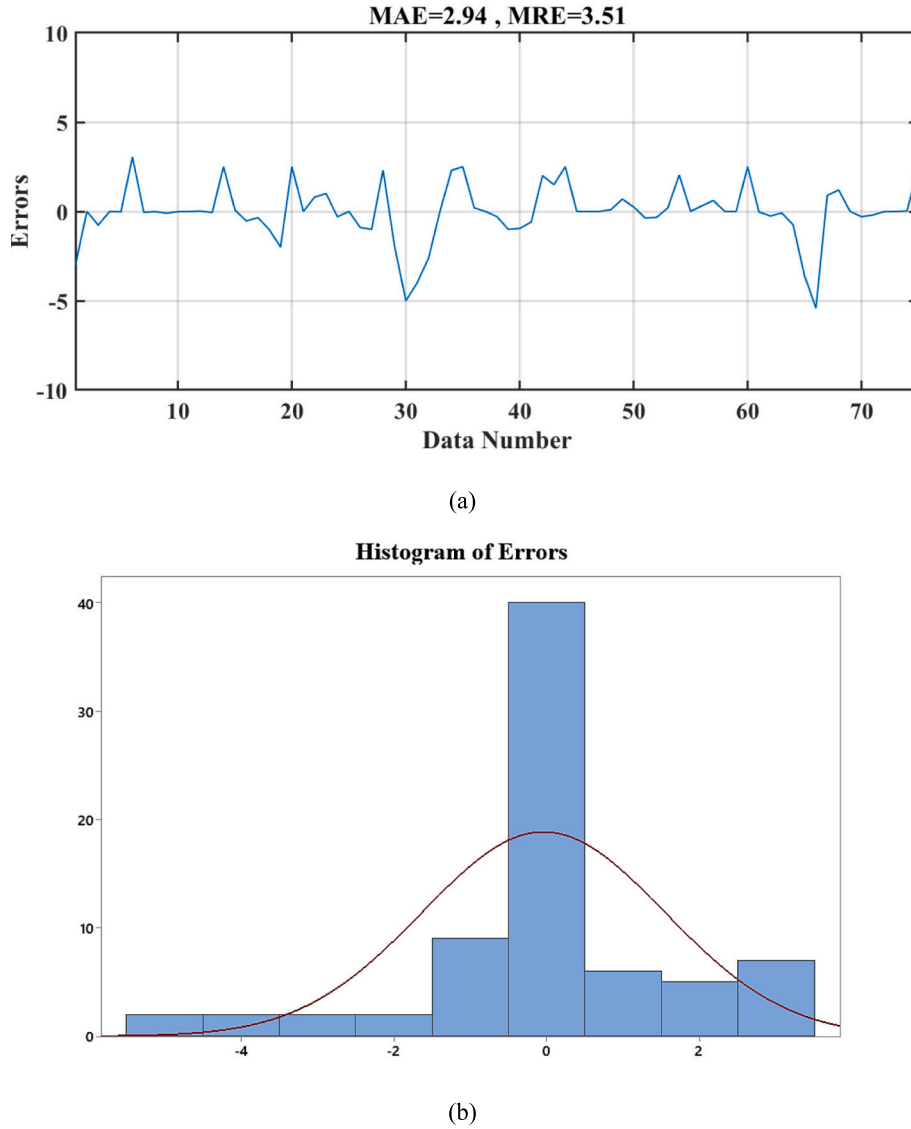


Fig. 10. Performance of neural network designed for training data (a) values, and (b) histogram chart of errors.

4.1. MLP implementation

The MLP neural network was trained using a dataset of verified Wilkinson PD designs. Here, MATLAB was used for implementing and simulating the network due to its computational capabilities and integration with advanced machine learning algorithms. The training process involved presenting known input parameters and their corresponding outputs to the network. Once trained, the network was tested to ensure it could accurately verify whether a given PD design met the established criteria [30].

The network was designed with input neurons representing parameters such as transmission line length and resonator dimensions, while the output layer corresponds to performance metrics like insertion loss, isolation, and return loss. The dataset was split into training and test sets to prevent overfitting and monitor performance [31].

During training, an optimization algorithm, such as Gradient Descent, adjusted the network's weights and biases to minimize the loss function, which measures the difference between predicted and actual values. Once trained, the network was used for optimization, identifying the best combination of parameters to meet the performance specifications for the Wilkinson PD [32,33].

4.2. The designed MLP network

To model the designed system, as shown in Fig. 9, the length of the TL, the width of the resonator, the length of the resonator, and the number of the resonators are chosen as the input parameters of the network, while S_{12} , S_{11} and S_{23} are chosen as the output parameters.

In this section, the MLP network is used to predict the performance and suitability of the parameters. It should be noted that the proper S_{12} in the pass band and the center frequency of the PD is -3 dB, S_{11} is below -15 dB, and the value of S_{23} should be approximately less than -15 dB.

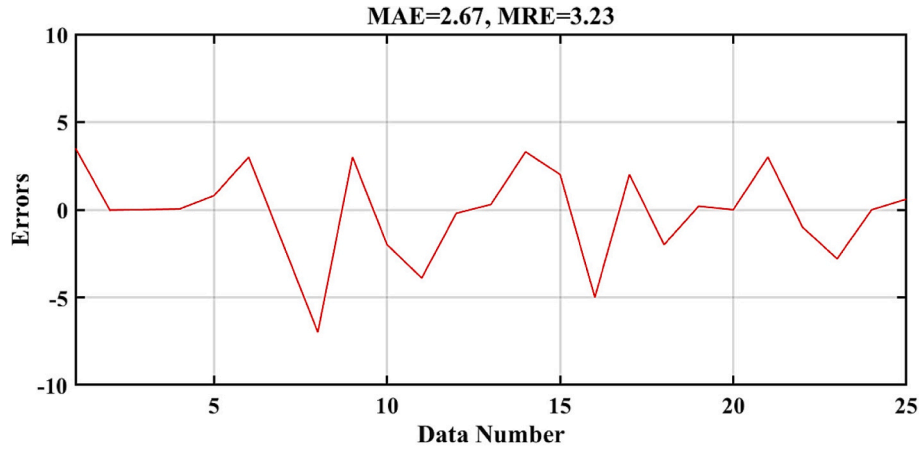
The parameters of the proposed MLP model are presented in Table 3.

The Levenberg-Marquardt (LM) learning algorithm has been used to train the proposed MLP network. This is because it provides correct and fast convergence and reduces complexity [34].

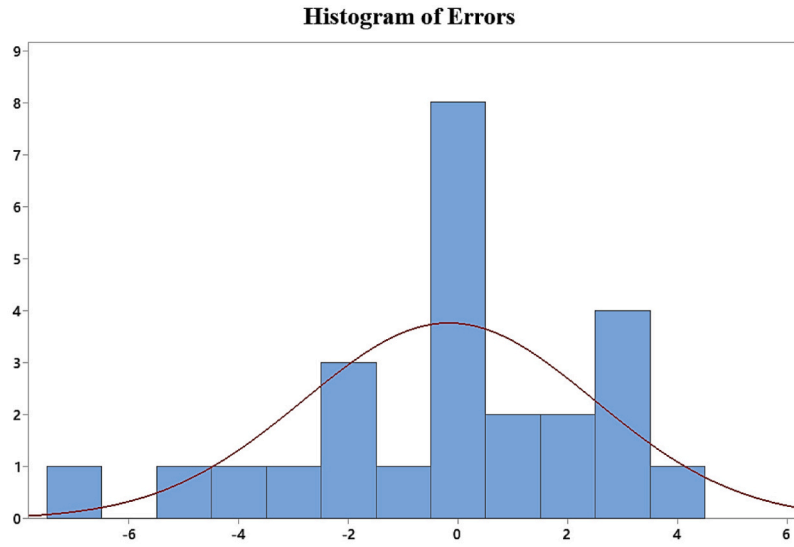
In the implemented model, 250 data are used where 75% of the data are considered for training the neural network and the remaining data are used for the test set. Also, MATLAB software has been used to implement and simulate the proposed neural network model.

In this neural network model, two errors are defined, which are calculated as follows [35,36].

$$MAE = \frac{1}{N} \sum_{i=1}^N |X_i(\text{Exp}) - X_i(\text{Pre})| \quad (28)$$



(a)



(b)

Fig. 11. Performance of neural network designed for test data (a) values, and (b) histogram chart of errors.

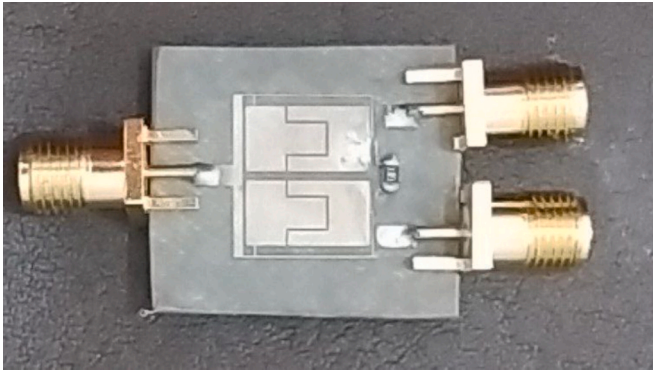


Fig. 12. The fabricated PD.

$$MRE = \frac{1}{N} \sum_{i=1}^N \left| \frac{X_i(\text{Exp}) - X_i(\text{Pre})}{X_i(\text{Exp})} \right| \quad (29)$$

In these equations, $X_i(\text{Exp})$ represents the expected measured output

value of the parameter, and $X_i(\text{Pre})$ denotes the predicted value obtained from the MLP neural network model. The results of the errors of the proposed MLP model are shown in Table 4.

As can be seen, the S_{12} and S_{23} obtained by the MLP model are almost identical to the actual values and have very small MAE and MRE errors for both test and training data. Therefore, the proposed model is valid for the dimensions and structure of the presented Wilkinson PD.

The MLP neural network is employed to optimize the design parameters and validate the theoretical models. The neural network architecture, detailed in Table 3 and Fig. 9, is specifically designed to map input parameters (e.g., transmission line length, resonator dimensions) to output performance metrics (S_{11} , S_{12} , S_{23}). By training the network on simulation data, we minimize prediction errors and ensure that the design meets performance specifications. The MLP uses the Levenberg-Marquardt learning algorithm to achieve fast convergence, and the resulting low MRE and MAE values confirm the accuracy of the theoretical and simulated models.

Figs. 10 and 11 provide a quantitative assessment of the predictive performance of the proposed MLP neural network for the S_{12} scattering parameter, using training and test datasets, respectively. In each figure, subfigure (a) presents the predicted S_{12} values versus the actual target

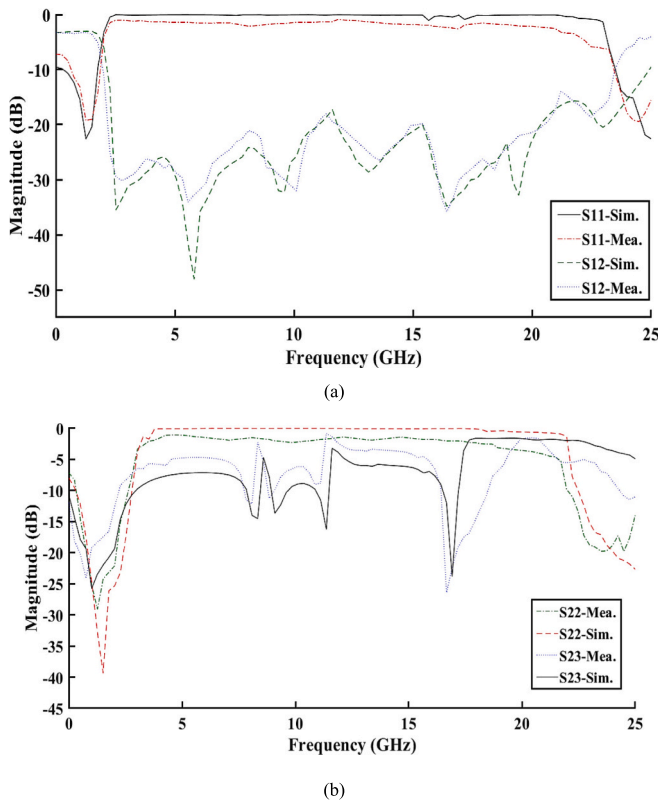


Fig. 13. The result of measurement and simulation of parameters (a) S_{11} , S_{12} , and (b) S_{22} , S_{23} .

Table 5

Comparison of the performance characteristics of this work with previous studies.

Ref.	Center frequency (GHz)	FBW (%)	S_{11} (dB)	S_{12} (dB)	Size (mm × mm)	Harmonic suppression
[11]	1.8	35	−21.68	−3.29	20.6 × 23.5	2nd—9th
[37]	2.45	19.6	−19.5	−3.59	16.4 × 26.1	2nd—3rd
[38]	2.85	52.9	−20	−3.62	13.5 × 17.8	2nd—4th
[39]	1.75	62.3	−20	−3.3	23.4 × 25.6	2nd—4th
[40]	2.2	18.2	−19.5	−3.95	11.35 × 14	2nd—3rd
[41]	1.5	41	−20.7	−3.22	15.3 × 11.6	2nd—8th
[42]	1	58	−25.7	−3.8	33.4 × 46.3	—
[43]	1.8	—	−33.2	−3.23	36 × 25	2nd—11th
This PD	1.4	96	−24	−3.12	9.04 × 11.4	2nd—16th

values, while subfigure (b) displays the histogram of absolute prediction errors. The complete dataset comprises 250 unique simulated instances of Wilkinson PD designs, each characterized by four input parameters: the transmission line length, resonator width, resonator length, and the total number of resonators. These input variables were systematically varied using parameter sweeps in ADS simulation software to span the design space. Subsequently, 75% of this dataset (188 samples) was allocated for supervised training of the network, while the remaining 25% (62 samples) formed the unseen test set used to evaluate generalization.

The error histograms in Figs. 10(b) and 11(b) offer a distributional perspective on model accuracy. For the training dataset, over 75 samples are shown due to denser sampling, while fewer are depicted in the test dataset (approximately 25) because of the smaller test set size. In both cases, the errors are tightly distributed around zero, confirming the absence of significant bias. The MAE and MRE, whose values are displayed at the top of each figure, are consistent with those listed in Table 4 and serve as global summary statistics. These metrics reflect the average prediction deviation in both absolute and relative terms and demonstrate that the MLP model has achieved high accuracy with low generalization error. This strong agreement between predicted and simulated values confirms that the network has successfully captured the underlying nonlinear relationships between design parameters and S_{12} behavior, enabling rapid and reliable performance prediction of future PD designs without requiring exhaustive simulation.

5. Fabrication and measurement

After ensuring the correct design and operation of the proposed PD, the simulated circuit is fabricated using microstrip technology on an RT/duroid Rogers 5880 substrate. The photograph of the fabricated circuit along with the SMA high-frequency ports connected to it is shown in Fig. 12.

An HP 8510 network analyzer has been used to measure the scattering parameters of the fabricated structure, and the results for parameters S_{11} and S_{12} are shown in Fig. 13(a), and the results of S_{22} and S_{23} are shown in Fig. 13(b).

The results obtained in Fig. 13(a) show that at the central frequency (f_0) of 1.4 GHz, the insertion loss rate, or S_{12} of the designed circuit is only −3.12 dB. At the same frequency, the value of S_{11} is approximately −24 dB, which means that the return loss is practically zero. This circuit has created a very wide pass band, while the return loss level is less than −15 dB in the frequency range from 1.15 GHz to 2.5 GHz, which is very ideal due to the very limited variation of the pass loss in this frequency range. The fractional bandwidth (FBW=BW/ f_0) is 96% given $f_0 = 1.5$ GHz.

The noteworthy point in the obtained results is the suppression of unwanted harmonics, i.e. the second to sixteenth harmonics are removed with a suppression level of −20 dB.

Also, from the results obtained in Fig. 13(b), it can be seen that in the pass band of the fabricated PD, the value of return loss from its output ports is less than −20 dB, and considering that the S_{23} parameter is also in the same range, it is zero (less than −15 dB), so the output ports are well isolated and separated from each other. The circuit is designed symmetrically, so only the return loss response from one output is simulated and measured, as the response of the other output will be exactly the same.

The presented PD introduces several key innovations that enhance its performance and practicality. First, the use of rectangular multi-section resonators—rather than conventional transmission lines—effectively suppresses unwanted harmonics, contributing to improved signal purity. This geometric approach is both novel and functionally advantageous.

Second, a MLP neural network is employed for design optimization and performance verification. The resulting low MRE and MAE confirm the accuracy and optimality of the proposed design.

Third, the structure achieves a size reduction of over 76% compared to traditional PDs, making it significantly more compact and suitable for space-constrained applications.

Fourth, the design suppresses 16 unwanted harmonics to levels below −20 dB across 1.15–2.5 GHz, while maintaining a power division frequency of 1.4 GHz, demonstrating excellent harmonic suppression. Finally, the PD is benchmarked against similar works in Table 5, showing superior performance in terms of size, suppression, and bandwidth.

6. Conclusion

In this paper, a high FBW Wilkinson PD based on rectangular multi-section resonators is presented. The design has been analyzed using even- and odd-mode analysis, with dimensions of $9.04 \text{ mm} \times 11.4 \text{ mm}$, fabricated using microstrip technology. The MLP neural network model was employed to optimize and verify the design. The proposed PD achieves a remarkable FBW of 96%, eliminating unwanted harmonics up to the 16th order, and an input return loss of -24 dB at the central frequency of 1.4 GHz . Compared to existing Wilkinson PDs, this solution offers several key advantages: it achieves over 76% size reduction, making it ideal for compact applications. The 96% FBW is notably higher than most traditional designs, providing superior broadband performance. Additionally, the PD suppresses up to the 16th harmonic, significantly reducing unwanted interference and improving signal purity. The simple structure ensures cost-effectiveness, which is an important consideration for mass production. These advantages position the proposed Wilkinson PD as a superior alternative to conventional designs, offering enhanced size, performance, harmonic suppression, and cost efficiency, making it highly suitable for modern wireless communication systems.

CRediT authorship contribution statement

Hesam Ghitasy Sarabi: Writing – original draft, Validation, Software, Methodology, Investigation, Formal analysis, Data curation. **Gholamreza Karimi:** Writing – review & editing, Validation, Supervision, Resources, Investigation, Data curation. **Milad Mohammadi:** Validation, Investigation. **Derek Abbott:** Review & editing, Validation.

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Declaration of competing interest

The authors declare that they have no known competing financial interests or personal relationships that could have appeared to influence the work reported in this paper.

Data availability

Data will be made available on request.

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