Design of Wilkinson Power Dividers with SITL Compensated Microstrip Bandpass Filters

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Abstract – This paper presents a simple technique for improving performances of a conventional Wilkinson power divider. The technique is achieved by replacing bulky quarter-wave transmission lines with stepped impedance transmission lines (SITL) compensated coupled lines. With the internal function of bandpass filter integrated with the proposed coupled lines, the spurious response at 2^{nd} harmonics frequencies that normally exists in the conventional divider is considerably reduced. Simulated and measured results at 2.1 GHz operating frequency of the proposed and conventional Wilkinson power divider were compared. The proposed divider achieves -3.8 dB insertion loss (S_{21}, S_{31}) and more than 25 dB return loss (S_{11}, S_{22}, S_{33}) across 10% fractional bandwidth. Based on this measurement, the proposed circuit achieves more than 34.5 dB suppression at the 2^{nd} harmonic frequency.

Index Terms – Harmonic suppression, parallel-coupled line, step impedance transmission line, Wilkinson power divider.

I. INTRODUCTION

Microwave circuits have many applications such as filter circuit [1], reflectometer [2], diplexer [3], Wilkinson power divider/combiner [4–6], microstrip add-drop multiplexer [7, 8], and microwave sensor application [9, 10]. Since the microwave circuit is implemented and developed with the microstrip substrate material, the Wilkinson power divider is a general microwave and millimeter wave communication circuit [11].

The Wilkinson power divider was first introduced by Wilkinson [4] and is an essential component for microwave and millimeter-wave applications. It is widely used because of its helpful property of being perfectly matched at all ports and sound isolation between the output ports [11, 12]. Moreover, the Wilkinson power divider is widely used to design and implement microstrip coupled lines. The non-homogeneous nature of physical parallel coupled lines microstrip gives rise to spurious frequencies, particularly beyond the second harmonic frequency. To address this issue, compensation techniques such as inductive [13, 14] and capacitive [15, 16] methods are commonly employed. These techniques play a crucial role in mitigating the effects of non-idealities, thereby enhancing the performance and reliability of microstrip-based circuits. Notwithstanding, researchers have presented Wilkinson power divider techniques for harmonic suppression frequencies such as an anti-coupled line [17], parallel-coupled line structure [18], looped coupled-line [19], inductively loaded microstrip line [20], and bandpass or band stop filter [21–23]. The utilization of power divider applications needing a narrowband divider and bandpass filter designs are prevalent in the construction of I-Q demodulator topologies, leading to the creation of simplified circuits [24–27]. For RF and microwave circuit design, including balanced amplifiers, phase shifters, image rejection mixers, I/Q modulators, and circularly polarized antenna polarizers, the techniques mentioned above can produce spurious harmonic frequencies. However, those techniques have limitations on the complexity of the circuit.

The Wilkinson power divider consists of two quarter-wavelength branches of transmission lines with a characteristic impedance and a termination resistor, as

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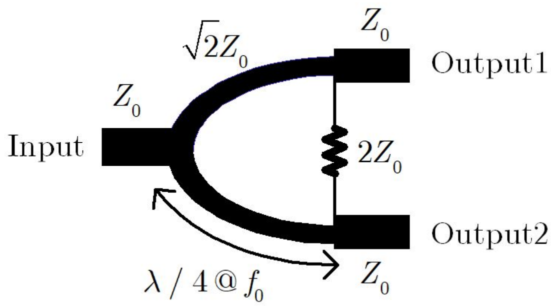

Fig. 1. Schematics of conventional Wilkinson power divider.

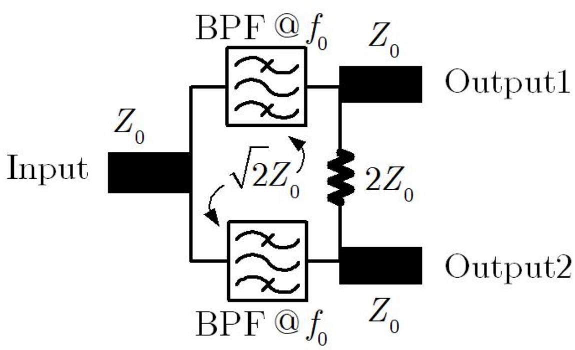

Fig. 2. Schematics of the proposed integrated bandpass filter Wilkinson power divider.

shown in Fig. 1. The power divider is designed based on the electrical length of a desired fundamental frequency. Still, high-order harmonic signals usually appear at the output ports due to the periodic characteristics of the transmission lines. Using the quarter-wavelength branches of transmission lines will unavoidably lead to poor selectivity in each transmission path. Intuitively, the selectivity can be improved by adding an extra high-order bandpass filter before the input or after each output port of the power divider. Still, the sizeable total circuit area may be a terrible problem. The concept of this research work is to add the BPF function in the divider without circuit size trade-off as shown in Figs. 1 and 2, a quarter-wave transformer is replaced with a transformer with integrated BPF. The basic theory of the proposed topology is described in Section II. Design and experimental results will be described in Section III. The paper is finally concluded in Section IV.

II. THEORY

A. The step impedance transmission lines resonator

Step impedance transmission lines (SITL) are non-uniform transmission lines, which were used in the filter design to reduce the circuit sizes [28, 29], to shift the spurious to higher band, and to suppress multiple spurious passband [29]. The SITL employed in this paper is shown in Fig. 1. It consists of two cascaded transmission lines with characteristic impedances Z_a and Z_b and electrical lengths θ_1 and θ_2, respectively. In this technique,$Z_a > Z_b$, so the impedance ratio $R_z = Z_a/Z_b$ is aligned between $0 \leq R_z \leq 1$. As shown in Fig. 4, the SITL is proposed to connect at the coupled lines port to suppress signal transmission between port 1 and port 3 of the coupled lines. To achieve this condition, the electrical parameters of coupled lines $(Z_0, Z_{0e}, Z_{0o}, \varepsilon_{effe}, \varepsilon_{effo})$ should be chosen optimally.

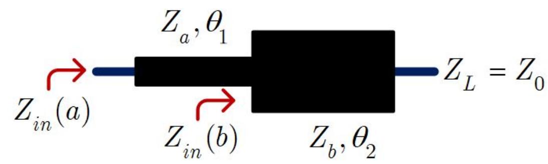

Fig. 3. Schematic of the proposed step impedance transmission line.

Based on the above-mentioned condition, the optimum directivity at operating frequency (f_0) of coupled-line occurs when the signal transmission of ports 1 and 3 is enforced to be nearly zero or $S_{31}(f_0) \approx 0$, which will occur as the driving point impedance from the coupled port to the SITL for this design $Z_{in}(a)$ is also enforced to [30] :

$$Z_{ZT}(f_0) = -j\frac{(Z_{0o}{}^2 Z_{0e} sin\theta_0 - Z_{0e}{}^2 Z_{0o} sin\theta_e) + 2Z_0Z_\beta + Z_0Z_\alpha}{Z_\alpha + Z_\beta}, \quad (1)$$

where $Z_\alpha = jZ_0(sin\theta_0 - sin\theta_e)$ and $Z_\beta = Z_0^2(cos\theta_0 - cos\theta_e)$.

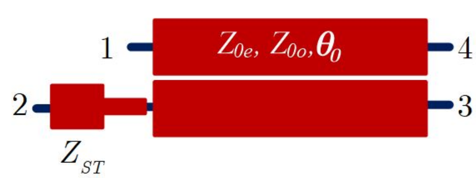

Fig. 4. The proposed coupled lines based SITL compensated coupled lines.

The input impedance $Z_{in}(a)$ of the SITL section at operating frequency (f_0) is related to the coupled lines electrical parameters. To preserve or minimize the change of the coupling coefficient of the proposed coupled lines, its electrical length $\theta_S(f_0)$ should be shortened and computed from:

$$\theta_S = \frac{1}{\Theta} cot^{-1}\left(\frac{2\pi f_0(Z_{in}(a) - Z_0)}{Z_{0o}}\right). \quad (2)$$

B. Proposed BPF resonator based SITL compensated coupled lines

Based on the impedance parameters [11], the relationship between voltages and currents at each port shown in Fig. 5, are given as follows:

$$V_1 = Z_{11}I_{11} + Z_{12}I_{12} + Z_{13}I_{13} + Z_{14}I_{14}, \quad (3)$$

$$V_2 = Z_{21}I_{21} + Z_{22}I_{22} + Z_{23}I_{23} + Z_{24}I_{24}, \quad (4)$$

$$V_3 = Z_{31}I_{31} + Z_{32}I_{32} + Z_{33}I_{33} + Z_{34}I_{34}, \quad (5)$$

$$V_4 = Z_{41}I_{41} + Z_{42}I_{42} + Z_{43}I_{43} + Z_{44}I_{44}, \quad (6)$$

where V_n and I_n (n = 1..4) are each ports voltage and current. From the initial condition in Fig. 5, where $V_4 = 0$ and $V_2 = -I_2 Z_{ST}$, then 4-port network is transformed to 2-port bandpass filter, and the relationship between voltage, current and impedance of 2-port network are as in equation (7):

$$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} Z_{11T} & Z_{12T} \\ Z_{21T} & Z_{22T} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}, \quad (7)$$

where

$$Z_{11T} = \left(Z_{11} - \frac{Z_{12}^2}{Z_{ST}+Z_{11}}\right) + \left(Z_{14} - \frac{Z_{12}Z_{13}}{Z_{ST}+Z_{11}}\right) Z_{\Theta}, \quad (8)$$

$$Z_{12T} = \left(Z_{13} - \frac{Z_{12}Z_{14}}{Z_{ST}+Z_{11}}\right) + Z_{\Phi}, \quad (9)$$

$$Z_{21T} = \left(Z_{13} - \frac{Z_{12}Z_{14}}{Z_{ST}+Z_{11}}\right) + \left(Z_{12} - \frac{Z_{13}Z_{12}}{Z_{ST}+Z_{11}}\right) Z_{\Theta}, \quad (10)$$

$$Z_{22T} = \left(Z_{11} - \frac{Z_{12}Z_{14}}{Z_{ST}+Z_{11}}\right) + Z_{\Phi}, \quad (11)$$

when

$$Z_{\Theta} = -\left(\frac{Z_{14}(Z_{ST}+Z_{11}-Z_{12}Z_{13})}{Z_{11}(Z_{ST}+Z11)-Z_{13}^2}\right), \quad (12)$$

and

$$Z_{\phi} = -\left(\frac{Z_{12}(Z_{ST}+Z_{11}-Z_{13}Z_{14})}{Z_{11}(Z_{ST}+Z11)-Z_{13}^2}\right). \quad (13)$$

From equation (7), the input reflection coefficient S_{11} and the forward transmission coefficient S_{21} of the synthesized BPF resonator are as shown in equations (14) and (15) :

$$S_{11} = \frac{(Z_{11T}^2 - Z_0^2) - Z_{12T}^2}{(Z_{11T}+Z_0)(Z_{22T}+Z_0) - Z_{11T}^2}, \quad (14)$$

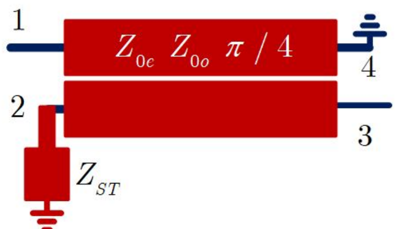

Fig. 5. The proposed BPF resonator based SITL compensated coupled lines.

$$S_{21} = \frac{2Z_0 Z_{12T}}{(Z_{11T}+Z_0)^2 - Z_{12T}^2}. \quad (15)$$

These concepts, theories, and equations were employed to design a BPF resonator based SITL coupled lines at an operating frequency (f_0) of 2.1 GHz. In this paper,-10 dB coupled lines with even and odd-mode impedances of $Z_{0e} = 69.37\ \Omega$, $Z_{0o} = 36.03\ \Omega$, for a 50 Ω system is chosen to design the proposed BPF resonator. $Z_{ST}(f_0)$ and $\theta_S(f_0)$ are calculated from equations (1) and (2). After that the impedance $Z_{ST}(f_0)$ and $\theta_S(f_0)$ electrical degree were used to calculate the electrical parameters Z_a, Z_b and θ_a, θ_b of SITL, respectively. Table 1 lists the electrical parameters of the conventional and the proposed BPF resonator based SITL compensation design for -10 dB while the electrical length of the conventional Wilkinson circuit is 0.50π rad and the proposed Wilkinson circuit is 0.42π rad, with the circuit size reduced by 0.08π rad in electrical length. The parameters of the SITL have an electrical length of 8 rad, respectively.

The frequency response (S_{21}) vs. return loss (S_{11}) curve of the conventional and the proposed BPF resonator based SITL compensated coupled lines simulated in this study are shown in Fig. 6. The simulation results of S_{21} and S_{11} presented frequency response at the same

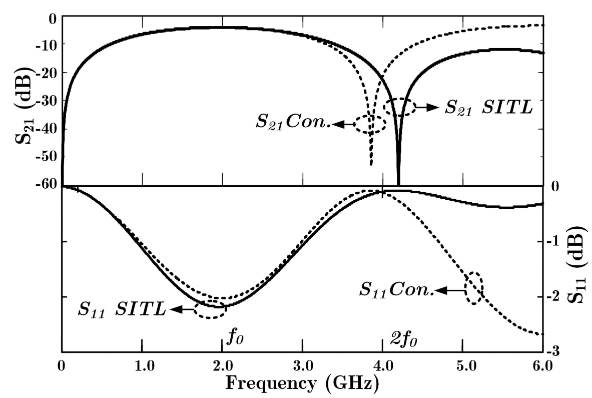

Fig. 6. S-parameters of 2.1 GHz conventional and the proposed BPF resonators.

Table 1: Parameters of the -10 dB coupled lines for BPF resonator at 2.1 GHz

Sections	Impedances (Ω)	Coupler's Length (θ, rad)	W,S,L (mm)
Cpl-SITL	$Z_{0e1} = 69.37$ $Z_{0o1} = 36.03$	0.42π	2.20, 0.17 , 21.00
Con-Cpl	$Z_{0e1} = 69.37$ $Z_{0o1} = 36.03$	0.50π	2.20, 0.17 , 25.00
SITL	$Z_{in}(a)$ = j22.20	$Z_{a1} = 100\Omega$, $Z_{b1} = 62\Omega$	$\theta(Z_a) = \theta(Z_b)$ = 8°

level compensated by STIL. The resulting frequency at $2f_0$ is higher than in the conventional.

C. Transmission lines modified for second-order BPF frequency response characterization

In the paper presented here, the filter, as shown in Fig. 7 is designed using the Chebyshev bandpass prototype technique [32, 33]. The element values for the considered prototype were proposed to determine the even- and odd-mode impedances by using the admittance inverter. The method of using a higher-order parallel coupled-line bandpass filter to replace transformer [32, 33] can be utilized in the proposed equal-split Wilkinson power divider with favorable selectivity. The SITLs compensation coupled lines BPF design equations of the proposed divider can be employed [33], where N is the filter's order, z_{0ei} and z_{0oi} are the even- and odd-mode characteristic impedances of each coupled-line section, Z_0 and Δ are the system impedance and the 3-dB fractional bandwidth of the filter,J_i,i=1,2,3,...,N+1, are the admittance inverters (J-inverters),g_i,i=1,2,3,...,N+1 are the lumped element values for low-pass filter prototype. In this conventional Wilkinson power divider, the characteristic impedance of each branch of quarter-wavelength transmission lines in Fig. 1 is $\sqrt{2}Z_0$. In the proposed technique, each quarter-wavelength branch of transmission lines in the Wilkinson power divider can be replaced

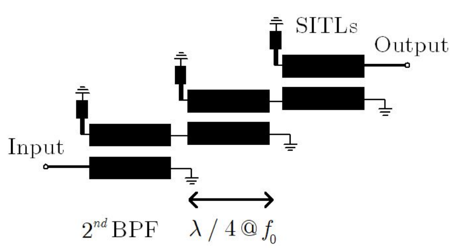

Fig. 7. Schematic of the proposed 2^{nd} order BPF SITL compensated coupled lines [31].

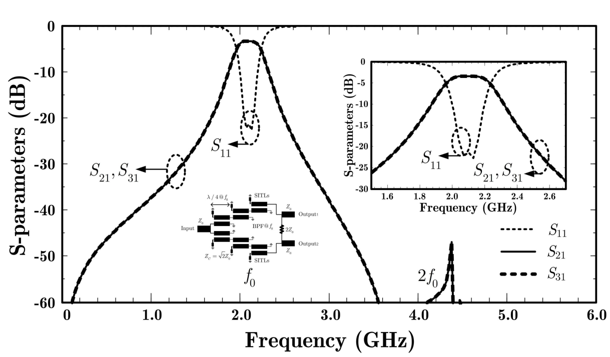

Fig. 8. Simulation results in S-parameters of the proposed Wilkinson power divider.

by the even-order coupled lines bandpass filter where the system characteristic impedance of the filter is $\sqrt{2}Z_0$ as shown in Fig. 8 and the proposed power divider has employed 2^{nd} order BPF based SITL compensated coupled lines instead of an ordinary quarter-wavelength branch of transmission lines to improve the desired selectivity in each output signal.

III. DESIGN AND EXPERIMENTAL RESULTS

To enhance the selectivity performance of the proposed Wilkinson power divider, the circuit prototype has been designed and fabricated on AD260 microwave substrate with the following design parameters: ε_r=2.60, h=1.00 mm, t=35μm and $tan\delta$=0.0017. The centre frequency and 3-dB fractional bandwidth of each filter transformer (both upper- and lower-BPF) are around 2.1 GHz and 5% fractional bandwidth. The corresponding design parameters of each 0.5 dB equal-ripple Chebysheve filter transformer are g_1=1.4029, g_2=0.7071, g_3=1.9841, Z_0J_1=0.2366, Z_0J_2=0.0792, and Z_0J_3=0.2366. The even- and odd-mode characteristic impedances of each coupled-line section are Z_{0e1}=91.40 Ω,Z_{0o1}=57.93 Ω,Z_{0e2}=76.75 Ω, Z_{0o2}=65.55 Ω, Z_{0e3}=60.89 Ω, and Z_{0o3}=87.93 Ω, respectively. The physical and electrical design parameters of SITLs compensated BPF are shown in Table 2.

In Fig. 9, the simulated result of the proposed Wilkinson power divider shows the suppression performances at 4.2 GHz below the desired frequency of 2.1 GHz more than 43 dB. The measurement was performed with the E5071C Vector Network Analyzer as shown in Fig. 10, which is in good agreement with the simulated result. The implemented circuit has the measured results

Table 2: Parameters of the Wilkinson power divider with integrated BPF at 2.1 GHz

Sections	Impedances (Ω)	Coupler's Length (θ, rad)	W,S,L (mm)
Coupled lines 1 $Z_{0o1}=57.93$	$Z_{0e1}=91.40$	0.42π , 21.00	$1.30, 0.59$
Coupled lines 2	$Z_{0e2}=76.75$ $Z_{0o2}=65.55$	0.39π	$1.50, 1.90$, 19.60
Coupled lines 3	$Z_{0e3}=91.40$ $Z_{0o3}=57.93$	0.42π	$1.30, 0.59,$ 21.00
SITLs	$Z_{in}(a1)$ = j28.42 $Z_{in}(a2)$ = j36.32 $Z_{in}(a3)$ = j23.42	$Z_{a1}=100\Omega$ $Z_{b1}=68\Omega$ $Z_{a2}=100\Omega$ $Z_{b2}=74\Omega$ $Z_{a3}=100\Omega$ $Z_{b3}=68\Omega$	$\theta(Z_a)=\theta(Z_b)$ $=8^\circ$

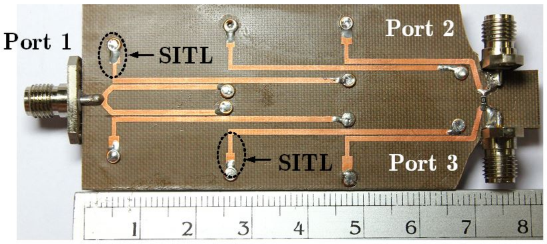

Fig. 9. PCB photograph of the proposed Wilkinson power divider.

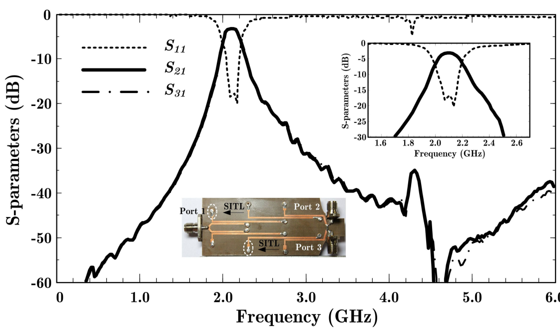

Fig. 10. Measured results s-parameters of the proposed Wilkinson power divider.

S_{21} -4.13 dB, S_{31} -4.12 dB, S_{11} -19.5 dB at the desired frequency 2.1 GHz and presents the suppression performances at 4.2 GHz more than 34.5 dB. The photograph of the PCB of the proposed frequency doubler based on a square ring resonator is shown in Fig. 11. The circuit size excluding the input and output SMA connectors of the proposed Wilkinson power divider is around 22.27 cm^2. Table 3 presents a summary of previous studies alongside

Table 3: The comparison of Wilkinson power dividers

Ref.	Techniques	Harmonic Suppression
Ref [4]	Open stub	> -70.00 dB
Ref [6]	LPFs	> -40.55 dB
Ref [12]	Lumped-distributed	NA
Ref [17]	Anti-coupled line	> -20.00 dB
Ref [18]	Parallel coupled line	> -30.00 dB
Ref [19]	Looped coupled-line	NA
Ref [21]	Open-short stubs	> -50.00 dB
Ref [23]	BSF	> -20.00 dB
This work	BPF	> -20.00 dB

our proposed circuit. Our design employs BPF coupled lines, which offers a simpler approach compared to the more intricate designs employed by other researchers. As a result, the proposed Wilkinson power divider proves to be more suitable for practical implementation.

IV. CONCLUSION

A modified Wilkinson power divider with integrated SITL compensated coupled lines bandpass filters has been proposed. The proposed topology has many attractive characteristics such as low insertion loss, good selectivity in each output signal path, wide stop band, and ease of design and implementation. The level of spurious response at a harmonic of the proposed Wilkinson power divider is then 34.5 dB at the 2nd harmonic frequency. It is believed that the proposed power divider can be simply modified for many microwave systems.

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