Received: 15 November 2021 | Revised: 20 February 2022 | Accepted: 12 April 2022

DOI: 10.1002/mmce.23202

INTERNATIONAL JOURNAL OF RF AND MICROWAVE COMPUTER-AIDED ENGINEERING WILEY

ORIGINAL ARTICLE

A design of impedance transformer with tunable center frequencies using filter synthesis approach

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Funding information
National Research Foundation of Korea (NRF) Grant Funded by Korean Government (MSIT), Grant/Award Number: 2020R1A2C2012057; Basic Science Research Program through NRF Funded by Ministry of Education, Grant/Award Number: 2019R1A6A1A09031717; Research Base Construction Fund Support Program Funded by Jeonbuk National University in 2021

Abstract

Microwave impedance transformer (IT) with center frequency tunable capability is one of great research issue for better utilization of frequency spectrum resources and system miniaturization of next generation wireless communication systems. This article presents a filter synthesis approach for designing IT with center frequency tunable capability using single dc-bias controlled tunable resonators. The proposed IT provides bandpass filtering response and is designed by properly coupling quarter-wave tunable resonators with alternate *J*- and *K*-inverters. The *J*-inverters are realized using a parallel-coupled line, whereas *K*-inverter is implemented with T-type short-circuited TL. Using filter theory, analytical design equations are derived to calculate the circuit parameters. The center frequency of the proposed IT is tuned over a wide frequency tuning range (FTR) by changing the varactor diode capacitance of quarter-wave resonators. For experimental demonstration, a prototype of 25-to-50 Ω IT is designed, simulated, and measured. The measured results confirmed that the center frequency of the proposed IT is tuned from 2.05 to 2.91 GHz (860 MHz or 34.68% FTR) with an insertion loss variation of 2.57 to 1.76 dB.

KEYWORDS

bandpass filter response, frequency tunable, impedance transformer, quarter-wave transmission line resonator

1 | INTRODUCTION

Impedance transformer (IT), which convert certain impedance into another one; is one of key components in radio frequency and microwave systems. These ITs are widely used for power dividers/combiner, power amplifiers, and antenna feeding networks.[1–7] In general, IT can be easily realized using quarter-wave ($\lambda_g/4$) transmission line (TL), however, it cannot provide bandpass filtering response. More recently, various designs of filtering ITs have been presented with different topologies including parallel coupled lines with shunt open/short stubs, multi-section parallel coupled lines, and stepped impedance resonators.[8–11] Likewise, wideband filtering ITs with out-of-band suppression are presented in References 12,13, using coupled line and shunt half-wavelength TL. In References 14,15, ITs with enhanced impedance transforming ratio and wide out-of-band suppression are demonstrated. Using multi-section coupled line and modified coupled line, the wideband ITs are presented in References 16–18. In References 19–21, ITs with extremely low or high impedance transforming are presented. More recently, the multi-band ITs are presented in References 22–23. In Reference 24, tunable IT is demonstrated at fixed center frequency using split strip lines. Despite significant research, conventional ITs are designed for fixed center frequency and none of the previously reported ITs can provide the center frequency tunability.

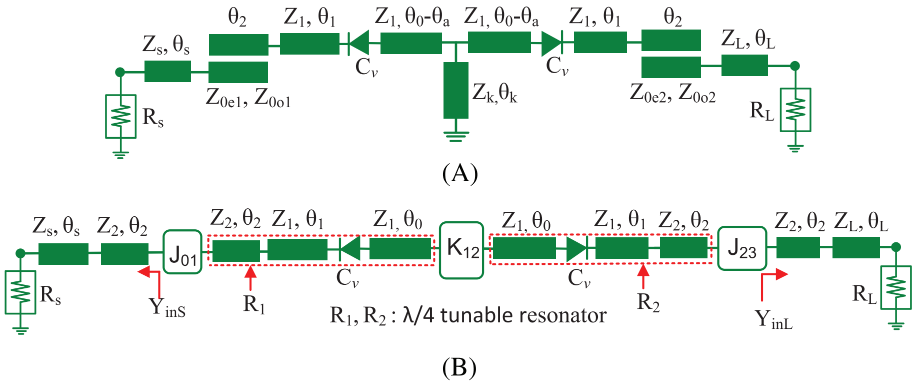


FIGURE 1 (A) Proposed structure of impedance transformer with tunable center frequency and (B) equivalent circuit of proposed impedance transformer

Simultaneous impedance transformation and center frequency tunable capability of IT are of great interest to future wireless communication for better utilization of limited frequency spectrum resources and system miniaturization.

In this article, we present a filter synthesis approach to design filtering IT with center frequency turnability capability over a wide frequency tuning range (FTR) using $\lambda_g/4$ TL resonators. The frequency tunability can be achieved by tuning the bias voltage of a varactor diode. We derive general design equations for designing IT with center frequency tunability by using filter theory.

2 | ANALYTICAL ANALYSIS

2.1 | Proposed structure of impedance transformer

Figure 1A shows the proposed structure of the IT with tunable center frequency. The source and load impedances are terminated with R_s and R_L, respectively. The proposed IT consists of source/load connecting transmission lines (TLs), parallel-coupled lines, tunable resonators, and T-type TLs. The TL with $Z_s = 1/Y_s$ and θ_s and the TL with $Z_L = 1/Y_L$ and θ_L are connected at source and load, respectively. The tunable resonators (R_1, R_2) are coupled to source/load connecting TLs through coupled lines. The equivalent circuit of the proposed IT is shown in Figure 1B. The parallel-coupled lines are equivalent to J-inverters, whereas the T-type TLs are equivalent to K-inverters. Using filter theory, the values of J- and K-inverters are determined[25] as (1).

$$J_{01} = \sqrt{\frac{\mathrm{Re}(Y_{inS})\Delta b}{g_0 g_1}} \tag{1a}$$

$$K_{12} = \frac{\Delta x}{\sqrt{g_1 g_2}} \tag{1b}$$

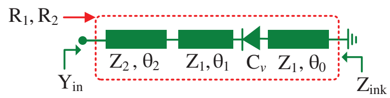


FIGURE 2 The proposed structure of a tunable quarter-wave resonator

$$J_{23} = \sqrt{\frac{\mathrm{Re}(Y_{inL})\Delta b}{g_2 g_3}} \tag{1c}$$

where g_i ($i = 0, 1, 2, 3$), Δ, b, and x are low-pass prototype-element values, fractional bandwidth, susceptance slope parameter and reactance slope parameter of tunable resonator, respectively. Similarly, $\mathrm{Re}(Y_{inS})$ and $\mathrm{Re}(Y_{inL})$ are the real part of input impedance looking at the first and last inverter as shown in Figure 1B, respectively.

2.2 | Resonant frequency and slope parameter of resonator

Figure 2 shows the proposed structure of a tunable $\lambda_g/4$ resonator. The resonator consists of three TLs with characteristic impedance and electrical lengths of Z_1, Z_2, θ_0, θ_1, θ_2, and a varactor diode capacitance C_v. The input admittance looking toward short-circuit TL is given as (2).

$$Y_{in} = jY_2 \frac{Y_A + Y_2 \tan\theta_2}{Y_2 - Y_A \tan\theta_2} = jB, \tag{2}$$

where

$$Y_A = Y_1 \frac{Y_B + Y_1 \tan\theta_1}{Y_1 - Y_B \tan\theta_1}, Y_B = \frac{2\pi f C_v Y_1 \cot\theta_0}{Y_1 \cot\theta_0 - 2\pi f C_v}, Y_1 = \frac{1}{Z_1}, Y_2 = \frac{1}{Z_2} \tag{3}$$

Similarly, the input impedance toward open circuited TL shown in Figure 2, is given as (4).

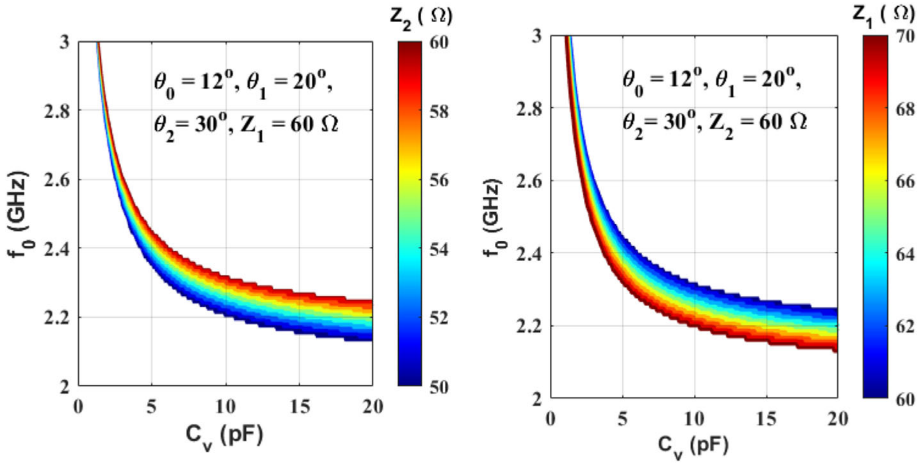


FIGURE 3 Resonant frequency of the proposed resonator according to C_v, Z_1, and Z_2. Electrical lengths of TLs (θ_0, θ_1, θ_2) are defined at 1.50 GHz. Color bar represents value of Z_1 and Z_2

$$Z_{ink} = jZ_1 \frac{Z_M + Z_1 \tan\theta_0}{Z_1 - Z_M \tan\theta_0} = jX, \tag{4}$$

where

$$Z_M = Z_1 \frac{Z_1 \tan\theta_1 - Z_2 \cot\theta_2}{Z_1 + Z_2 \cot\theta_2 \tan\theta_1} - \frac{1}{2\pi f C_v}. \tag{5}$$

The resonant frequency (f_0) can be solved by setting $B = im(Y_{in}) = 0$ or $X = im(Z_{ink}) = 0$. Similarly, the susceptance slope parameter (b) and reactance slope parameter (x) of the resonator at resonant frequency f_0 can be found as (6).

$$b = \frac{f_0}{2} \frac{dB}{df}\bigg|_{f=f_0} \tag{6a}$$

$$x = \frac{f_0}{2} \frac{dX}{df}\bigg|_{f=f_0} \tag{6b}$$

Figure 3 shows the calculated resonant frequencies of the proposed resonator. The resonant frequency decreases as C_v increases. From this figure, we also note that resonant frequency is high when Z_1 is low and Z_2 is high for the same C_v.

2.3 | Analysis of source and load connecting transmission lines

The source/load connecting TLs (Z_S, θ_S, and Z_L, θ_L) used to match source/load impedances to another real impedance at the first and last inverters. Using Figure 1B, input admittances (Y_{inS}) at first J-inverter looking toward source terminated impedance (R_S) is given as (7) assuming the impedance transforming ratio $r = R_L/R_S$ providing $R_L > R_S$.

$$Y_{inS} = \mathrm{Re}(Y_{inS}) + j\mathrm{Im}(Y_{inS}), \tag{7}$$

where

$$\mathrm{Re}(Y_{inS}) = Y_2 \frac{\alpha_1\alpha_3 + \alpha_2\alpha_4}{\alpha_3^2 + \alpha_4^2} \tag{8a}$$

$$\mathrm{Im}(Y_{inS}) = Y_2 \frac{\alpha_2\alpha_3 - \alpha_1\alpha_4}{\alpha_3^2 + \alpha_4^2} \tag{8b}$$

$$\alpha_1 = \frac{1}{rR_L}(Y_s - Y_2 \tan\theta_2 \tan\theta_s) \tag{8c}$$

$$\alpha_2 = Y_s(Y_s \tan\theta_s + Y_2 \tan\theta_2) \tag{8d}$$

$$\alpha_3 = Y_s Y_2 - Y_s^2 \tan\theta_s \tan\theta_2 \tag{8e}$$

$$\alpha_4 = \frac{1}{rR_L}(Y_2 \tan\theta_s + Y_s \tan\theta_2). \tag{8f}$$

The electrical length θ_S of source connecting TL that can transform R_S to another real impedance at f_0 looking at the first J-inverter, can be derived as (9) by equating (8b) to zero.

$$\theta_s = \tan^{-1}\left(\frac{-\alpha_6 \pm \sqrt{\alpha_6^2 - 4\alpha_5\alpha_7}}{2\alpha_5}\right), \tag{9}$$

where

$$\alpha_5 = \left(R_L^2 Y_2^2/r^2 - Y_s^4\right)\tan\theta_2 \tag{10a}$$

$$\alpha_6 = Y_2 Y_s\left(Y_s^2 - R_L^2/r^2\right)\left(1 - \tan^2\theta_2\right) \tag{10b}$$

$$\alpha_7 = Y_s^2\left(Y_2^2 - R_L^2/r^2\right)\tan\theta_2. \tag{10c}$$

Once θ_S is calculated, the value of $\mathrm{Re}(Y_{inS})$ looking at first J-inverter is determined as (8a). Similarly, using Figure 1B, input admittances (Y_{inL}) at last J-inverter looking toward load terminated impedance (R_L) is given as (11).

$$Y_{inL} = \mathrm{Re}(Y_{inL}) + j\mathrm{Im}(Y_{inL}), \tag{11}$$

where

$$\mathrm{Re}(Y_{inL}) = Y_2 \frac{k_1k_3 + k_2k_4}{k_3^2 + k_4^2} \tag{12a}$$

$$\mathrm{Im}(Y_{inL}) = Y_2 \frac{k_2k_3 - k_1k_4}{k_3^2 + k_4^2} \tag{12b}$$

$$k_1 = 1/R_L(Y_L - Y_2\tan\theta_2\tan\theta_L) \tag{12c}$$

$$k_2 = Y_L(Y_L\tan\theta_L + Y_2\tan\theta_2) \tag{12d}$$

$$k_3 = Y_LY_2 - Y_L^2\tan\theta_L\tan\theta_2 \tag{12e}$$

$$k_4 = 1/R_L(Y_2\tan\theta_L + Y_L\tan\theta_2). \tag{12f}$$

Electrical length θ_L of the load connecting the TL that can transform R_L to another real impedance at f_0 looking at the last J-inverter is given as (13) by equating (12b) to zero.

$$\theta_L = \tan^{-1}\left(\frac{-k_6 \pm \sqrt{k_6^2 - 4k_5k_7}}{2k_5}\right), \tag{13}$$

where

$$k_5 = \left(Y_2^2/R_L^2 - Y_L^4\right)\tan\theta_2 \tag{14a}$$

$$k_6 = Y_2Y_L\left(Y_L^2 - 1/R_L^2\right)\left(1 - \tan^2\theta_2\right) \tag{14b}$$

$$k_7 = Y_L^2\left(Y_2^2 - 1/R_L^2\right)\tan\theta_2. \tag{14c}$$

Once θ_L is obtained, the value of $\mathrm{Re}(Y_{inL})$ looking at the last J-inverter is determined by using (12a).

2.4 | Implementation of J- and K-inverters

Figure 4A shows the parallel-coupled lines and their equivalent J-inverter. The even- and odd-mode impedances (Z_{0ei}, Z_{0oi}) of parallel-coupled lines with arbitrary electrical length θ_2 and characteristic impedance Z_2 are given as (15) in terms of J-inverter.

$$Z_{0ei} = Z_2 \frac{1 + J_{i,i+1}Z_2\csc\theta_2 + J_{i,i+1}^2Z_2^2}{1 - J_{i,i+1}^2Z_2^2\cot^2\theta_2} \tag{15a}$$

$$Z_{0oi} = Z_2 \frac{1 - J_{i,i+1}Z_2\csc\theta_2 + J_{i,i+1}^2Z_2^2}{1 - J_{i,i+1}^2Z_2^2\cot^2\theta_2} \tag{15b}$$

Figure 4B shows the K-inverter implementation with T-type TLs. The T-type K-inverter consists of a series TL with a characteristic impedance of Z_1 and electrical length of θ_a and a shunt short-circuited TL with a characteristic impedance of Z_k and electrical length of θ_k. The circuit parameters of the T-type K-inverter can be derived by equating the $ABCD$-parameters of the K-inverter and T-type TLs and are given as (16).

$$\theta_a = -\tan^{-1}(K_{12}/Z_1) \tag{16a}$$

$$\theta_k = \tan^{-1}\left\{\frac{K_{12}Z_1^2}{Z_1^2Z_k - K_{12}^2Z_k}\right\} \tag{16b}$$

The negative electrical length θ_a will be absorbed by resonator electrical length θ_0. Therefore, the electrical length of the resonator will be shorter than the original length as shown in Figure 1A.

2.5 | Step by step design method

Based on the analytical design equations described in previous sections, the step-by-step design guidelines for the proposed IT with center frequency tunability can be summarized as follows.

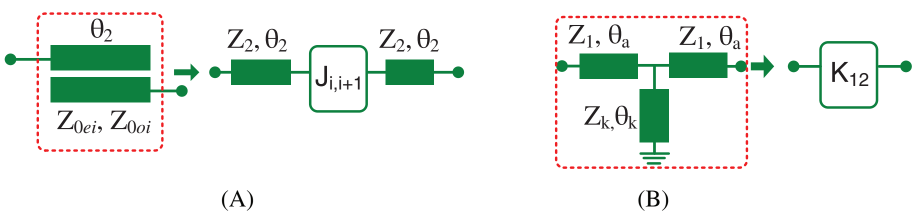


FIGURE 4 Circuit implementation of (A) J-inverter and (B) K-inverter

1. The design process starts with setting IT specifications such as the passband ripple, Δ, R_S, R_L.
2. After IT specification is determined, choose resonator parameters such as Z_2, Z_1, θ_2, θ_1, θ_0, C_v, and f_{ref}.
3. Calculate the resonant frequency (f_0) by setting $B = 0$ in (2) or $X = 0$ in (4). Similarly, calculate the susceptance and react slope parameters b and x using (6).
4. Calculate θ_S and θ_L at f_{ref} using (9) and (13) by assuming the value of Z_S and Z_L, respectively. Once θ_S and θ_L are obtained, then calculate the values of $\mathrm{Re}(Y_{inS})$ and $\mathrm{Re}(Y_{inL})$ from (8b) and (12b), respectively.
5. Calculate the J and K-inverter values using (1). After obtaining J-inverter values, calculate the even- and odd-mode impedances of the parallel-coupled line using (15). For practical implementation of the

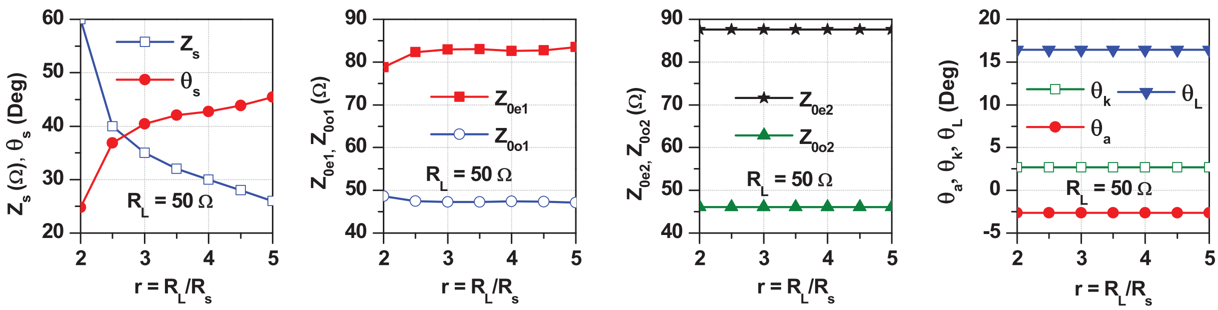
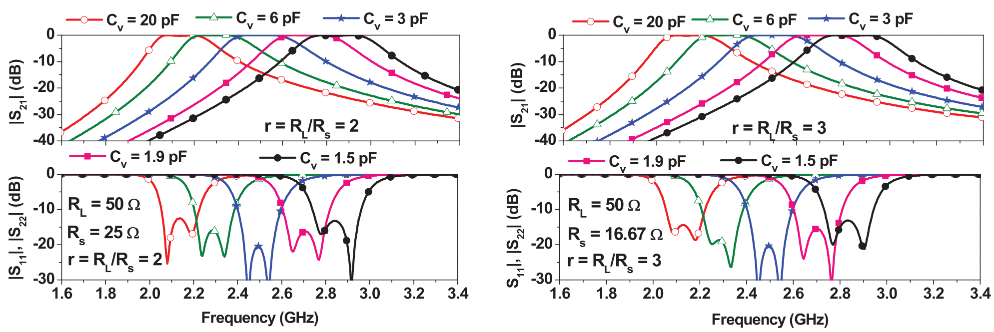


FIGURE 5 Calculated circuit parameters of proposed impedance transformer with $R_L = 50$, Ω, $\Delta = 5\%$, Chebyshev passband ripple of 0.043 dB, $f_{ref} = 1.5$ GHz, $C_v = 3$ pF, $Z_1 = Z_k = Z_L = 70\ \Omega$, $Z_2 = 60\ \Omega$, $\theta_0 = 12°$, $\theta_1 = 20°$, $\theta_2 = 29.4°$

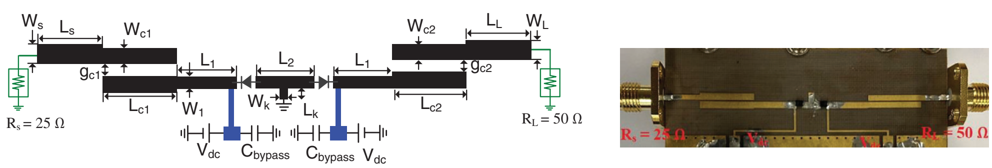


(A) (B)

FIGURE 6 Simulation results of the proposed impedance transformer with tunable center frequency: (A) $r = 2$ and (B) $r = 3$

TABLE 1 Calculated circuit parameters

$R_L = 50\ \Omega$, $\Delta = 5\%$, Chebyshev passband ripple $= 0.043$ dB, $f_{ref} = 1.5$ GHz, $C_v = 3$ pF, $Z_1 = Z_k = Z_L = 70\ \Omega$, $Z_2 = 60\ \Omega$, $\theta_0 = 12°$, $\theta_1 = 20°$, $\theta_2 = 29.4°$ $\theta_k = 2.67°$, $\theta_a = -2.65°$, $\theta_L = 16.45°$, $Z_{0e2} = 87.63\ \Omega$, $Z_{0o2} = 46.08\ \Omega$					
$r = R_L/R_s$	Z_s (Ω)	θ_0 (°)	Z_{0e1} (Ω)	Z_{0o1} (Ω)	C_v (pF)
2	60	24.81	78.77	48.61	1.5 ~ 20
3	35	40.43	82.97	47.27	1.5 ~ 20



(A) (B)

FIGURE 7 (A) Physical layout and (B) photograph of fabricated impedance transformer. Physical dimensions: $W_s = 1.2$, $W_L = 1.06$, $W_{c1} = 1.28$, $W_1 = W_k = 1.36$, $W_{c2} = 1.50$, $L_S = 5.8$, $L_{c1} = L_{c2} = 12$, $L_1 = 9.5$, $L_2 = 4.36$, $L_k = 2.04$, $L_L = 7.50$, $g_{c1} = 0.18$, $g_{c2} = 0.3$. Unit: millimeter

K-inverter, calculate θ_a and θ_k using (16). The negative electrical length θ_{ai} will be compensated within θ_0, with the result that θ_0 is shorter than the original value.

6. Once all the circuit parameters are determined, the center frequency is tuned by changing the value of C_v.

Using the step-by-step design guidelines described above, the calculated circuit parameters of the proposed IT are shown in Figure 5. As seen in this figure, Z_S decreases and θ_S increases as the impedance-transforming ratio (r) increases. Similarly, Z_{0e1} slightly increases but Z_{0o1} decreases as r increases. However, Z_{0e2} and Z_{0o2} remain the same for all r.

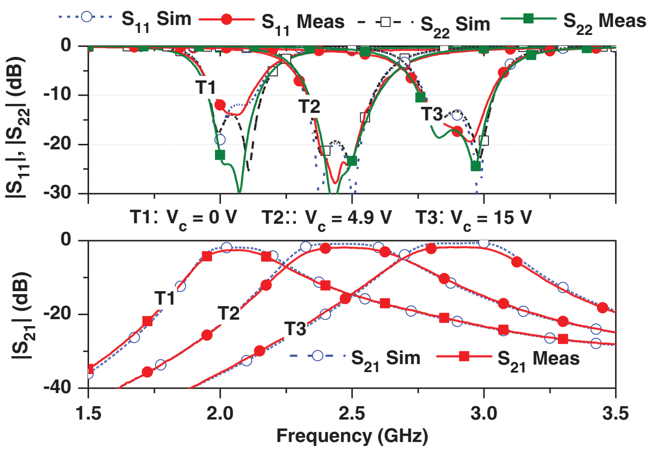


FIGURE 8 Simulation and measurement results of 25-to-50 Ω ($r = 2$) impedance transformer with tunable center frequency

TABLE 2 Measurement results

V_{dc} (V)	f_0 (GHz)	IL (dB)	RL (dB)	3-dB BW (MHz)
15	2.91	1.76	17.2	400
8	2.45	1.88	24.2	400
0	2.05	2.57	13.5	290

2.6 | Design examples

To validate the analytical analysis, Figure 6 shows the frequency response of IT with $r = 2$ and 3. The circuit parameters are shown in Table 1. These results show that the proposed IT can provide the bandpass filtering response. Besides, the center frequency of IT can be tuned from 2.13 to 2.86 GHz by adjusting capacitance from 20 to 1.5 pF, respectively. These results also confirm that the proposed circuit can integrate IT and the center-frequency tunable bandpass filter within a single device, which is essential for future wireless communication system miniaturization.

3 | SIMULATION AND MEASUREMENT RESULTS

For experimental validation, a 25-to-50 Ω ($R_S = 25\ \Omega$ and $R_L = 50\ \Omega$ with $r = 2$) IT was designed, fabricated, and measured with Chebyshev passband ripple of 0.043 dB, $\Delta = 5\%$ and $f_{ref} = 1.5$ GHz on Taconic substrate with a dielectric constant of 2.2 and thickness of 0.787 mm. The calculated circuit parameters are given as $C_v = 3$ pF, $Z_S = 60\ \Omega$, $Z_L = Z_1 = Z_k = 70\ \Omega$, $Z_{0e1} = 78.77\ \Omega$, $Z_{0o1} = 48.61\ \Omega$, $Z_{0e2} = 87.63\ \Omega$, $Z_{0o2} = 46.08\ \Omega$, $\theta_2 = 29.4$, $\theta_1 = 20°$, $\theta_0 = 12$, $\theta_S = 24.8\ °$, $\theta_L = 16.45°$, $\theta_a = -1.64°$, and $\theta_k = 2.67°$. The electrical lengths of TL are defined at $f_{ref} = 1.5$ GHz. The skyworks SMV-1233 varactors are used in this work, which can provide capacitance of 0.9 pF to 20 pF by varying DC bias voltage from 15 to 0 V. Figure 7 shows the physical layout and photograph of fabricated circuit.

Figure 8 shows the simulation and measurement results of fabricated IT. The measured results are consistent with

TABLE 3 Comparison between this work and other state-of-arts

	Frequency (GHz)	IL (dB)	RL_{min} (dB)	$r = R_L/R_s$	A
8	1.0	0.45	NA	2	No
10	1.0	0.82	20	2	No
11–15	2.60	<1	20	10/2/5	No
16	1	0.60	26	5	No
17	2	NA	18	2	No
18	1	NA	20	5	No
19	2.40	0.72	24.35	10	No
20	3.50	0.72	24.97	25	No
21	2.50	0.88	NA	20	No
This work	2.05 ~ 2.91	2.57 ~ 1.76	13 ~ 18	2	Yes

Note: A: Center frequency tunability, RL_{min}: Minimum input/output return losses.

simulated results. The measurement results are summarized in Table 2. These results show that the center frequency of IT can be tuned from 2.05 to 2.91 GHz (860 MHz or 34.68% FTR) with insertion loss variation of 2.57 to 1.76 dB by controlling DC bias voltage of 0 to 15 V. The measured 3-dB insertion loss bandwidth varies from 290 to 400 MHz. In addition, the measured input/output return losses are higher than 14 dB within the overall frequency tuning range.

The performance comparison of the proposed IT with previously reported works is shown in Table 3. As observed from this table, the proposed IT provides bandpass filtering response as well as center-frequency tunability capability, which was not possible in previously reported works.[8–23]

4 | CONCLUSION

This article demonstrated impedance transformer with tunable center frequency based on filter synthesis approach by using quarter-wave transmission-line resonators. Both analytical analyses and experimental results are shown for validation. For experimental demonstration, a prototype of a 25-to-50 Ω impedance transformer is designed and fabricated. The fabricated impedance transformer provides not only a wide range of center-frequency tunability but also arbitrary impedance-transforming characteristics. The proposed method can be easily extended to design a higher-order filtering impedance transformer. Therefore, the proposed circuit can integrate the impedance transformer and tunable filter within a single device and play an important role in the circuit miniaturization of next-generation microwave communication systems.

ACKNOWLEDGMENTS

This work was supported by National Research Foundation of Korea (NRF) grant funded by the Korean government (MSIT) under grant no. 2020R1A2C2012057, in part by the Basic Science Research Program through NRF funded by Ministry of Education under grant no. 2019R1A6A1A09031717, and in part by "Research Base Construction Fund Support Program" funded by Jeonbuk National University 2021.

DATA AVAILABILITY STATEMENT

The data that support the findings of this study are available on request from the corresponding author. The data are not publicly available due to privacy or ethical restrictions.

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How to cite this article: Chaudhary G, Jeong Y. A design of impedance transformer with tunable center frequencies using filter synthesis approach. *Int J RF Microw Comput Aided Eng*. 2022;e23202. doi:10.1002/mmce.23202