A Wideband Filtering Switch: Its Synthesis and Design

Citation for published version:
https://doi.org/10.1109/tmtt.2023.3297720

Digital Object Identifier (DOI):
10.1109/tmtt.2023.3297720

Link:
Link to publication record in Heriot-Watt Research Portal

Document Version:
Peer reviewed version

Published In:
IEEE Transactions on Microwave Theory and Techniques

Publisher Rights Statement:
© 2023 IEEE. Personal use of this material is permitted. Permission from IEEE must be obtained for all other uses, in any current or future media, including reprinting/republishing this material for advertising or promotional purposes, creating new collective works, for resale or redistribution to servers or lists, or reuse of any copyrighted component of this work in other works.

General rights
Copyright for the publications made accessible via Heriot-Watt Research Portal is retained by the author(s) and / or other copyright owners and it is a condition of accessing these publications that users recognise and abide by the legal requirements associated with these rights.

Take down policy
Heriot-Watt University has made every reasonable effort to ensure that the content in Heriot-Watt Research Portal complies with UK legislation. If you believe that the public display of this file breaches copyright please contact open.access@hw.ac.uk providing details, and we will remove access to the work immediately and investigate your claim.
A Wideband Filtering Switch: Its Synthesis and Design

Kai-Da Xu®, Senior Member, IEEE, Xiaoyu Weng, Qiang Chen®, Senior Member, IEEE, and Haijun Fan®, Member, IEEE

Abstract—This article presents a direct synthesis method for the design of a wideband filtering switch, where p-i-n diodes are incorporated into each resonator. As parasitic parameters are embedded in the p-i-n diodes, an ideal transformer is introduced in each resonator to provide the freedom to designate the proper values for inductances and capacitances in the resonators. According to the rejection specification in the ON state, an Nth order bandpass filter (BPF) network with N transmission zeros (TZs) is synthesized, followed by a new circuit model extraction procedure to extract lumped circuits in the BPF network. To simplify the network topology, two ideal transformers at input-output port remain after absorbing the ideal transformers between resonators. Then, the lumped circuit model is converted into a distributed circuit model in the synthesis process, maintaining the filtering response in a wide frequency range when the switch is in the ON state. Besides, the N TZs of the filtering switch in the OFF state can be introduced and arranged to suitable positions. Consequently, high isolation within the stopband of the switch in the OFF state can be achieved. For demonstration purposes, two filtering switches with fractional bandwidths (FBWs) of 30% and 50% are fabricated and measured. Good agreement can be observed between the simulated and measured results, which validates the feasibility of the proposed method.

Index Terms—p-i-n diode, single-pole-single-throw (SPST), transmission zeros (TZs), wideband response.

I. INTRODUCTION

A wideband filtering switch is an essential component in modern wireless communication systems, where microwave switches are utilized to control the communication channel and manage the signal pathway. Due to the poor selectivity of the switch, a filter is usually required in front of it to pass through the useful signal and suppress the interference. In recent years, the co-design of switches and filters has attracted much more attention compared with conventional independent components since the filtering switch structure has the advantages of improved ON-state selectivity, high off-state suppression, low loss of interconnections, and compact size.

To obtain high-performance filtering switches, the devices with inherent switching characteristics are widely utilized to design different switches, such as field-effect transistors (FETs) [1], [2], [3], [4], [5], [6], [7], [8], [9] and p-i-n diodes [10], [11], [12], [13], [14], [15], [16], [17]. Due to the negligible dc power consumption and the ease of integration to amplifiers, FET is mostly used to design monolithic microwave integrated circuit (MMIC) switches. In [1], a single-pole-double-throw (SPDT) switch was reported based on a composite resonator, which consists of a transmission line (TL) and an FET in series. When the FETs are biased below the pinch-off voltage, the frequency response of the switch in the ON state can be designed with filter synthesis methods. In [2], an FET-based resonator was proposed by combining the shorted TL with FET in parallel. By selecting appropriate coupling parameters, the MMIC switches can be easily designed based on the coupled resonator theory. In [5], a filtering switch was presented by replacing the capacitances in the lumped bandpass filter (BPF) with FET-state transistors. Besides, the FET can be utilized to provide the coupling between adjacent resonators [6] and control ON-state and OFF-state modes of resonators [8].

Similarly, the p-i-n diode is another option with two inherent states based on different bias voltages. The characteristics of p-i-n diode could be used to design switches with several technologies, such as dielectric resonator [10], coaxial cavity [11], low-temperature cofired ceramic (LTCC) [12], and printed circuit board (PCB) [13], [14], [15], [16], [17]. In [10], two p-i-n diodes were connected to two output feedlines, where the switching between two channels was achieved by controlling the coupling between the output feedlines and the dielectric resonator. In [13], the stepped-impedance resonators with p-i-n diodes loaded at one end were used to construct SPDT filtering switches. Excellent isolation in the OFF state could be obtained by properly adjusting the stepped length ratio of the stepped-impedance resonator. In [15], switches consisting of two distributed coupled tri-mode resonators loaded with opposite-polarity p-i-n diodes were presented, where the out-of-band rejection could be improved with the absorptive OFF-state property.

It is noted that most filtering switches integrated with FETs and p-i-n diodes are analyzed and designed by the coupled resonator theory, where the coupling matrix synthesis method is employed. The dimensions of coupling structures could be determined by the coupling matrix generated from...
the specifications of the switch in the ON state. However, it is challenging to design a wideband filtering switch using the coupling matrix synthesis method due to its narrowband nature. In addition, the characteristics of switches in the OFF state are hardly exploited in the existing synthesis process.

In this article, therefore, a direct synthesis method for the design of a wideband filtering switch has been proposed. A wideband filtering switch with \( N \) transmission zeros (TZs) is attained by embedding a p-i-n diode into each resonator and incorporating the p-i-n diode parasitic parameters into the wideband filter synthesis. High isolation within the switch stopband in the OFF state can be achieved by arranging the \( N \) TZs in the OFF state properly when converting the lumped circuit model into a distributed circuit model. For demonstration, two filtering switches with fractional bandwidths (FBWs) of 30\% and 50\% are designed and measured, validating the feasibility of the proposed method.

II. ANALYSIS OF WIDEBAND FILTERING SWITCH EMBEDDED WITH P-I-N DIODES

The proposed wideband filtering single-pole-single-throw (SPST) switch structure is shown in Fig. 1(a), which consists of lumped inductors, capacitors, p-i-n diodes, and two ideal transformers at the input and output ports. Different from the BPF topology in [18], as shown in Fig. 1(b), p-i-n diodes are embedded into each resonator to perform the switch function, and two ideal transformers are introduced at the input and output ports which stem from the synthesis process of the switch ON and OFF states and will be presented later in Section III. Due to parasitic parameters embedded in the p-i-n diodes, an ideal transformer will be introduced in each resonator to provide the freedom to designate the proper values for inductances and capacitances in the resonators. Two ideal transformers at the input and output ports remain after absorbing the ideal transformers between resonators. To simplify the synthesis process, the equivalent circuits of the p-i-n diode under reverse bias and forward bias conditions given in Fig. 2 are adopted [19], [20]. Since \( R_p \) in reverse bias is very large (usually in k\( \Omega \)) and \( R_0 \) in forward bias is very small (about or less than 1 \( \Omega \)), they can be seen as open-circuited and short-circuited, respectively. For convenience, they will both be ignored in the following analysis.

Fig. 3 shows the equivalent circuits of resonator \( i \) with p-i-n diode under forward bias and reverse bias, where \( L_{ri} \), \( C_{ri} \), and \( L_{pi} \) denote the circuit elements of resonator \( i \). Under reverse bias, whose circuit model is shown in Fig. 3(a), the transmission pole (TP) \( f_{p,R} \) and TZ \( f_{z,R} \) of the resonator can be found as

\[
f_{p,R} = \frac{1}{2\pi \sqrt{C_{ri} + C_0 - \frac{C_{ri} C_0}{L_{ri} + L_p + L_{pi}}}} \tag{1a}
\]

\[
f_{z,R} = \frac{1}{2\pi \sqrt{C_{ri} + C_0 - \frac{C_{ri} C_0}{L_{ri} + L_p + L_{pi}}}} \tag{1b}
\]

Under forward bias in Fig. 3(b), the TP \( f_{p,F} \) and TZ \( f_{z,F} \) can be expressed as

\[
f_{p,F} = \frac{1}{2\pi \sqrt{C_{ri} - \frac{1}{L_{ri} + L_p + L_{pi}}}} \tag{2a}
\]

\[
f_{z,F} = \frac{1}{2\pi \sqrt{C_{ri} - \frac{1}{L_{ri} + L_p + L_{pi}}}} \tag{2b}
\]

It is noted that \( f_{p,F} < f_{p,R} \) and \( f_{z,F} < f_{z,R} \), which means the response of resonator \( i \) will shift toward lower frequency under forward bias. Moreover, since \( f_{p,F} < f_{z,F} \) and \( f_{p,F} < f_{z,F} \), the TZ of resonator \( i \) under both states will be located on the right side of TP. Thus, \( f_{z,F} \) in each resonator is to be arranged in the interested frequency band, and no TP under forward bias is in the interested frequency band. This feature gives the feasibility of realizing the filtering switch function and featuring high isolation in the OFF state with such a resonator under forward bias. When the p-i-n diodes are under reverse bias condition, the filtering switch is in the ON state, while with forward bias, it is in the OFF state.

By replacing p-i-n diodes with the equivalent circuit under reverse bias, the filtering switch shown in Fig. 1(a) in the ON state can be transformed into a BPF network, as illustrated in Fig. 4. The BPF network with two ideal transformers can then
be synthesized and extracted based on the given specifications of filtering switch in the ON state. Next, the circuit elements $L_{ri}, C_{ri}, L_{pi}$, and $L_{pi}$ of the filtering switch in Fig. 1(a) can be determined by circuit elements $L'_{ri}, C_{ri}, L_{ri}'$, and $L_{pi}'$ in Fig. 4. Therefore, the circuit elements of the filtering switch can be calculated by

$$L_{ri} = L_{ri}' - L_0, \ C_{ri} = \frac{C_0 C_{ri}}{C_0 - C_{ri}}, \ L_{pi} = L_{pi}' (i = 1, 2, 3, \ldots, N)$$

$$L_{ri} = L_{ri}' (i = 1, 2, 3, \ldots, N, N + 1)$$

where $L_0$ and $C_0$ are parameters of p-i-n diode under reverse bias. To ensure $L_{ri}, C_{ri}$ is greater than zero, $L_{ri}' > L_0$, and $0 < C_{ri}' \leq C_0$ should be taken into consideration while extracting the circuit model of the BPF network.

III. CIRCUIT MODEL EXTRACTION

With the given specifications of the filtering switch in the ON state, the filtering function can be synthesized straightforwardly by the process presented in [21]. Based on the filtering function, the overall ABCD matrix can be derived [22], from which the circuit elements of the BPF topology can be sequentially extracted. Finally, the elements of the SPST switch are calculated based on (3), which are associated with the parameters of the p-i-n diode under reverse bias. In this work, the extraction process of an $N$th-order BPF network with $(N + 1)$ ideal transformers given in Fig. 5 is proposed. By introducing an ideal transformer with transforming ratio $n_i, L_{ri}'$, and $C_{ri}'$ can be designated with proper values meeting the requirements of $L_{ri}' > L_0$ and $0 < C_{ri}' \leq C_0$. Then, the BPF with two ideal transformers in Fig. 4 can be obtained by absorbing the ideal transformers with numbers 2 to $N$.

As illustrated in Fig. 6, the proposed circuit model extraction procedure for the circuit in Fig. 5 consists of five basic operations: 1) extract a series inductor $L_{ri}'$; 2) extract an ideal transformer; 3) extract a shunt branch of series $L_{ri}'$ and $C_{ri}'$; 4) extract a shunt inductor $L_{pi}'$; and 5) extract the remainder circuit with a shunt inductor, an ideal transformer, and a series inductor. Compared with the extraction method in [18], an additional ideal transformer is extracted between series inductor $L_{ri}'$ and the shunt branch of series $L_{ri}'$ and $C_{ri}'$, i.e., operation 2) illustrated in Fig. 6(b). Besides, each shunt resonator of series $L_{ri}'$ and $C_{ri}'$ is responsible for the generation of a finite TZ in the ON state. Therefore, the TZs of BPF should be allocated to all resonators before the extraction procedure. Starting from the left-hand side of the network, elements in Fig. 5 will be extracted in sequence. The following equations can present the network transformation during the extraction operations 1)–4) accordingly:

$$\frac{1}{P_1} \begin{bmatrix} A_{n1} & B_{n1} \\ C_{n1} & D_{n1} \end{bmatrix} = \frac{1}{P_2} \begin{bmatrix} A_{n1} + sL_{n1} C_{n2} & B_{n2} + sL_{n1} D_{n2} \\ C_{n2} & D_{n2} \end{bmatrix}$$

(4)

$$\frac{1}{P_1} \begin{bmatrix} A_{n2} & B_{n2} \\ C_{n2} & D_{n2} \end{bmatrix} = \frac{1}{P_3} \begin{bmatrix} A_{n2} & B_{n3} \\ C_{n3} & D_{n3} \end{bmatrix}$$

(5)

$$\frac{1}{P_1} \begin{bmatrix} A_{n3} & B_{n3} \\ C_{n3} & D_{n3} \end{bmatrix} = \frac{1}{P_3} \begin{bmatrix} A_{n3} & B_{n5} \\ C_{n5} & D_{n5} \end{bmatrix}$$

(6)

$$\frac{1}{P_1} \begin{bmatrix} A_{n4} & B_{n4} \\ C_{n4} & D_{n4} \end{bmatrix} = \frac{1}{P_4} \begin{bmatrix} A_{n5} & B_{n5} \\ C_{n5} & D_{n5} \end{bmatrix}$$

(7)

where the subscript $n$ denotes the numerator polynomials, e.g., $A_{n1}$ is the numerator polynomial of $A_1$, $s_j$ is the TZ assigned to the series $L_{ri}'$ and $C_{ri}'$ resonator circuit in the $s$-plane.

The relationship between denominator polynomials of these $ABCD$ matrices $(P_1–P_5)$ can be expressed as

$$P_1 = P_2 = P_3 = P_4 (s^2 - s_j^2), \quad P_4 = P_5.$$

(8)
Based on (4), the series element \( L'_{tie} \) extracted in operation 1) can be calculated by

\[
L'_{tie} = \frac{A_{n1} - A_{n2}}{s_{zj}C_{n2}} \bigg|_{s_{zj},n} = \frac{A_{n1}}{s_{zj}C_{n1}} \bigg|_{s_{zj},n} \tag{9}
\]

where \( C_{n2} = C_{n1} \) and \( A_{n2}(s_{zj}) = A_{n3}(s_{zj})/n \). Then, for the extraction of the ideal transformer in operation 2), \( L'_{ri} \) should be designated under the constraint of \( L'_{ri} > L_0 \). Correspondingly, \( C'_{ri} \) is in terms of \( L'_{ri} \) based on the assigned TZ \( s_{zj} \)

\[
C'_{ri} = -\frac{1}{s_{zj}^2}L'_{ri}. \tag{10}
\]

Obviously, \( 0 < C'_{ri} \leq C_0 \) should be satisfied as well.

From (5) and (6), \( A_{n2} \) and \( C_{n2} \) can be expressed as

\[
A_{n2} = \frac{1}{n}A_{n3} = \frac{1}{n} \left( \frac{1}{L_{ri}} - s_{zj}^2 \right) A_{n4} \tag{11a}
\]

\[
C_{n2} = n_iC_{n3} = n_i \left( \frac{1}{L_{ri}} s A_{n4} + (s^2 - s_{zj}^2) C_{n4} \right) \tag{11b}
\]

by equaling \( s \) to \( s_{zj} \), the transforming ratio \( n_i \) can be derived

\[
n_i = \frac{C_{n2}}{L_{ri} s A_{n4}} \bigg|_{s_{zj},n} = \frac{C_{n2}(s^2 - s_{zj}^2)}{L_{ri} s A_{n2}} \bigg|_{s_{zj},n}. \tag{12}
\]

Through (7), it seems that the shunt inductor \( L'_{pi} \) can be determined by

\[
L'_{pi} = \frac{A_{n5}}{s(C_{n4} - C_{n5})} = \frac{A_{n4}}{s(C_{n4} - C_{n5})}. \tag{13}
\]

However, it is unable to calculate \( L'_{pi} \) by eliminating \( C_{n5} \) at \( s_{zj} \) as the method in (9). To continue the extraction procedure, \( L'_{pi} \) should be predefined with an initial value, making all the extracted circuit elements positive. After operation 4), the \( ABCD \) matrix with the subscript \( n5 \) can be obtained based on (7), which will be used in the next cycle of extraction operations 1)–4). In the above analysis, the elements are derived from numerator polynomials of \( A \) and \( C \) at each extraction stage. The same value can be extracted from \( B \) and \( D \).

For operation 5), the overall \( ABCD \) matrix of remainder elements is given by

\[
\begin{bmatrix}
A_{nR} & B_{nR} \\
C_{nR} & D_{nR}
\end{bmatrix}
\begin{bmatrix}
\frac{n_i(N+1)}{n_i(N+1)R_{i(N+1)}} & n_i(N+1)S_{i(N+1)} \\
\frac{n_i(N+1)R_{i(N+1)}}{n_i(N+1)S_{i(N+1)}} & n_i(N+1)S_{i(N+1)} + \frac{4}{n_i(N+1)}
\end{bmatrix}
\]

\[
\tag{14}
\]

Therefore, these circuit elements can be obtained

\[
L'_{p(N+1)} = \frac{A_{nR}}{S_{nR}} n_{(N+1)} = \frac{A_{nR}}{S_{nR}} L'_{i(N+1)} = B_{nR} \tag{15}
\]

Here, \( L'_{p(N+1)} \) can be absorbed by the shunt inductor \( L'_{pN} \) extracted in operation 4) in the final cycle.

By repeating the basic operations 1)–4) for \( N \) times, the circuit elements with numbers 1–\( N \) of the proposed BPF in Fig. 5 can be obtained. Then, the remainder elements \( L'_{i(N+1)} \) and \( n_{(N+1)} \) are extracted by operation 5) to complete the extraction procedure. However, it is difficult to realize the circuits with ideal transformers. Additional circuit transformations should be used to eliminate the ideal transformers. As shown in Fig. 7, the combination of a series inductor and an ideal transformer can be replaced by a \( \Pi \)-network of inductors. Equating the \( ABCD \) matrices of two circuits, the elements of the \( \Pi \)-network can be obtained as

\[
L'_{i(N+1)} = n_i L'_{i(N+1)} L'_{p(N+1)} = L'_{i(N+1)} n_i - 1, \quad L'_{pi} = n_i L'_{pi} n_i - 1 = 1-n_i \tag{16}
\]

where \( L'_{pi} \) and \( L'_{pi} \) have opposite signs. For the ideal interresonator transformers \( (i = 2, \ldots, N) \), shunt inductors \( L'_{p(i-1)} \) and \( L'_{pin} \) of the equivalent \( \Pi \)-network can be absorbed by the adjacent inductors \( L'_{p(i-1)} \) and \( L'_{pi} \).

After eliminating the interresonator ideal transformers of the BPF network in Fig. 5, the BPF with two ideal transformers given in Fig. 4 can be obtained, where the values of \( L'_{pi} \) \( (i = 1, \ldots, N) \) and \( L'_{ri} \) \( (i = 2, \ldots, N) \) will be updated due to the absorption of ideal transformers. Besides, the final values of \( L'_{pi} \) and \( L'_{ri} \) in Fig. 4 will remain unchanged if choosing different initial values of \( L'_{pi} \) in operation 4).

A. Approximation of \( I/O \) Coupling Network

To eliminate two ideal transformers at the input and output ports, the network at input–output port as shown in Fig. 8 can be analyzed in Section IV-B.
be approximated by a TL \((Z_{zi}, \theta_{zi})\) and a shunt capacitor \(C_{pia}\). By evaluating the \(ABCD\) matrix of the two circuits at the center frequency \(f_0\), the equations can be obtained as

\[
\theta_{ti} = \arccos(n_i), \quad Z_{ti} = \frac{\omega_0 L_{ti}}{\tan(\theta_{ti})}, \quad C_{pia} = -\frac{\sin(2\theta_{ti})}{2n_i^2\omega_0 Z_{zi}} \quad (17)
\]

where \(n_i\) is less than 1 to ensure \(\theta_{ti}\) rational. It can be achieved by assigning \(L_{ti}\) and \(C_{ri}\) properly in the extraction procedure of the BPF network. Besides, \(\theta_{ti}\) is less than \(\pi/2\) because the transforming ratio \(n_i\) is positive, as seen in (12). Obviously, the capacitance \(C_{pia}\) is always negative, which could be eliminated through the approximations of the resonant circuits.

When \(f_0 = 4\) GHz, \(L_{ti} = 1\) nH, and \(n_i = 0.95\), \(Z_{zi}\), \(\theta_{ti}\), and \(C_{pia}\) are calculated as 76.5 \(\Omega\), 18.2\(^\circ\), and \(-0.171\) pF, respectively. The \(S\)-parameters of these two networks are compared and shown in Fig. 9.

### B. Approximation of Resonant Circuit

As illustrated in Fig. 10, the lumped resonant circuit is approximated by a short-circuited stub in parallel with a branch, which is a series of a TL, a \(p-i-n\) diode, and an open-circuited stub. To eliminate the negative capacitors generated from approximations of interresonator coupling inductors [18] and I/O coupling networks in (17), a shunt capacitor \(C_{pia}\) is included in the approximation of lumped resonant circuit. For the \(p-i-n\) diode, its equivalent circuits under forward bias and reverse bias are different, which will influence the calculation of distributed elements. Under reverse bias (switch in the ON state), the admittances of two resonant circuits in Fig. 10 can be expressed as

\[
Y_{in} = j\omega C_{pi} + \frac{1}{j\omega L_{pi}} + \frac{j\omega(L_{ri} + L_0) + 1}{j\omega C_{ri} + 1}
\]

\( (18a) \)

\[
Y'_{in} = \frac{1}{j\omega C_{pi}} + 1 + Z_{ri} + j Z_L \tan(\theta_{ri} \frac{\omega}{\omega_0}) + Z_L + j Z_{ci} \tan(\theta_{ci} \frac{\omega}{\omega_0})
\]

\( (18b) \)

\[
Z_L = j\omega L_0 + 1 + j Z_{ci} \tan(\theta_{ci} \frac{\omega}{\omega_0})
\]

\( (18c) \)

where \(Z_{pi}, Z_{ri}, \) and \(Z_{ci}\) are the characteristic impedances of TLs at \(f_0\), and \(\theta_{pi}, \theta_{ri}, \) and \(\theta_{ci}\) are the corresponding electric lengths at \(f_0\).

Under forward bias (switch in the OFF state), the impedance of distributed branch with the \(p-i-n\) diode is expressed as

\[
Z_{in} = Z_{ri} \left( \frac{j\omega L_0 - j Z_{ci} \cot(\theta_{ci} \frac{\omega}{\omega_0})}{Z_{ri} + j(j\omega L_0 - j Z_{ci} \cot(\theta_{ci} \frac{\omega}{\omega_0}))} \right) \tan(\theta_{ci} \frac{\omega}{\omega_0}) \quad (19)
\]

By equating \(Z_{in}\) to zero, the relationship between circuit parameters and TZ of the distributed resonant circuit under forward bias \(f_{TZ,DF}\) can be obtained as

\[
\omega_{TZ,DF} L_0 - Z_{ci} \cot(\theta_{ci} \frac{\omega_{TZ,DF}}{\omega_0}) + Z_{ri} \tan(\theta_{ci} \frac{\omega_{TZ,DF}}{\omega_0}) = 0.
\]

(20)

By setting (18a) equal to (18b) at the center frequency \(f_0\), the lower band-edge frequency \(f_1\), and the upper band-edge frequency \(f_2\), a set of nonlinear equations with the constraint of (20) can be obtained. With the characteristic impedance \(Z_{pi}, Z_{ri}, Z_{ci}, \) and \(f_{TZ,DF}\) predefined, the parameters \(\theta_{pi}, \theta_{ri}, \) and \(\theta_{ci}\) can be numerically solved. Note that the parameters of distributed resonant circuits are not unique due to the nonlinearity equation.

For example, with \(f_1 = 3\) GHz, \(f_0 = 4\) GHz, \(f_2 = 5\) GHz, \(C_{pi} = -0.4\) pF, \(L_{pi} = 6\) nH, \(L_{ri} = 0.57\) nH, \(C_{ri} = 9\) pF, \(L_0 = 0.4\) nH, and \(C_0 = 0.78\) pF, choosing \(Z_{pi}, Z_{ri}, \) and \(Z_{ci}\) as 80, 80, and 20 \(\Omega\), respectively, \(\theta_{pi}, \theta_{ri}, \) and \(\theta_{ci}\) can be calculated with different \(f_{TZ,DF}\). For distributed circuit 1, \(f_{TZ,DF}\) is set as 3.4 GHz, and the values of \(\theta_{pi}, \theta_{ri}, \) and \(\theta_{ci}\) are 54.6\(^\circ\), 9.2\(^\circ\), and 53.8\(^\circ\), respectively; for distributed circuit 2, \(f_{TZ,DF}\) is 4 GHz, \(\theta_{pi} = 54.2\(^\circ\), \(\theta_{ri} = 15.6\(^\circ\), and \(\theta_{ci} = 42.8\(^\circ\).

Fig. 11 shows the input admittances of the lumped circuit and two distributed circuits. As illustrated in Fig. 11(a) and (b), under reverse bias, good agreement between lumped circuits and distributed circuits can be achieved within the passband while the TZs of distributed circuits will deviate from that of the lumped circuits. In Fig. 11(c), the TZs of two distributed circuits under the forward bias are exactly located at the designated position. It means that the arrangement of \(f_{TZ,DF}\) hardly affects the approximation within the passband under reverse bias.

However, there is a significant discrepancy in the input admittances under the forward bias of lumped resonant circuit and its distributed approximations. As presented in Fig. 11(c), the TZ of the lumped circuit under forward bias is around 1.7 GHz, which is far away from that of two distributed circuits. This observation can be addressed by analyzing circuits...
Fig. 11. Input admittances of the lumped resonant circuit and its two distributed approximation circuits under (a) reverse bias within the passband, (b) reverse bias from 5.8 to 6.4 GHz, and (c) forward bias. For distributed circuit 1, TZ under forward bias is set as 3.4 GHz. For distributed circuit 2, TZ under forward bias is set as 4 GHz.

in Fig. 10. The parameters of components in the lumped circuit are frequency-invariant, while the electric length of TLs in distributed circuits varies with the operating frequency. It will lead to the difference between lumped and distributed resonant circuits outside the frequency band bounded by $f_1$ and $f_2$.

Moreover, the nonlinear equations grouped to derive the elements of the distributed resonant circuit are attained by equating the input admittances of two circuits under reverse bias, where $f_{TZ,DF}$ is involved as the constraint. For the design of the filtering switch, it is not necessary to chase good approximations of two resonant circuits under forward bias. To achieve better isolation under forward bias, $f_{TZ,DF}$ is generally allocated within $f_1$ and $f_2$ instead of the same position with lumped resonant circuit. Therefore, the performance under the forward bias of distributed circuits, especially the location of TZ in the OFF state, will be different from that of the lumped circuits, even though the lumped circuit under reverse bias can be well approximated within the passband by the distributed circuit.

Due to the above characteristics of approximation for the resonant circuit, the frequency response of the proposed lumped switch in the OFF state is negligible in the design procedure of distributed filtering switch. Controlling the TZs of the distributed switch in the OFF state becomes an effective method to obtain high isolation in the desired frequency band. However, the options for $f_{TZ,DF}$ are limited by the parameters of the lumped resonant circuit. The electric length of the TLs in the distributed resonant circuit could be less than zero with some choices of $f_{TZ,DF}$. To avoid such a predicament in the filtering switch design, the parameters of extracted BPF network and the arrangement of TZs in the OFF state should be continuously adjusted by iterations to obtain the realizable parameters of the distributed circuit. Therefore, the distributed filtering switch may be obtained while the frequency response of its lumped prototype in the OFF state is totally different, as the case illustrated in Fig. 11(c).

V. DESIGN PROCESS OF FILTERING SWITCH

Based on the circuit approximations under wideband conditions, the lumped switch shown in Fig. 1(a) can be converted into a distributed circuit through three steps: 1) the I/O coupling inductors ($L_{t1}$ and $L_{t(N+1)}$) and ideal transformers are approximated by TLs and shunt capacitors using (17); 2) the interresonator inductors $L_{ti}$ ($i = 2, \ldots, N$) are replaced by their approximated distributed circuits introduced in [18]; and 3) all the shunt resonant circuits together with the shunt capacitors generated in steps 1) and 2) are approximated by the distributed circuit in Fig. 10. The topology of the distributed filtering switch is illustrated in Fig. 12.

Finally, a general synthesis and design procedure of the proposed wideband filter can be concluded by the flowchart outlined in Fig. 13. To achieve a satisfactory performance, the procedure may take several iterations of fine-tuning the $L'_{ri}$ and $C'_{ri}$, modifying the assignment of TZs in the ON state and adjusting the TZs in the OFF state.
TABLE I
LUMPED ELEMENT VALUES OF FOURTH-ORDER FILTERING SWITCH

<table>
<thead>
<tr>
<th>i</th>
<th>(L_1) (nH)</th>
<th>(L_2) (nH)</th>
<th>(L_3) (nH)</th>
<th>(C_e) (pF)</th>
<th>(n_i)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1.38</td>
<td>5.13</td>
<td>0.58</td>
<td>9.36</td>
<td>0.87</td>
</tr>
<tr>
<td>2</td>
<td>2.08</td>
<td>5.94</td>
<td>0.72</td>
<td>9.36</td>
<td>-</td>
</tr>
<tr>
<td>3</td>
<td>1.69</td>
<td>2.03</td>
<td>0.9</td>
<td>9.36</td>
<td>-</td>
</tr>
<tr>
<td>4</td>
<td>1.71</td>
<td>11.06</td>
<td>0.52</td>
<td>9.36</td>
<td>-</td>
</tr>
<tr>
<td>5</td>
<td>1.46</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>0.98</td>
</tr>
</tbody>
</table>

Fig. 14. S-parameters of lumped element circuit models of the filtering switch in (a) ON state and (b) OFF state.

VI. DESIGN EXAMPLES

A. Fourth-Order Filtering Switch With 30% FBW in the ON State

To design a filtering switch, the BPF network should be synthesized first based on the specifications of the switch. In this example, the BPF network is designed with a center frequency of 4 GHz, FBW of 30%, fourth-order, and 20-dB in-band return loss. The TZs of four resonators are located at 6, 5.6, 5.2, and 6.2 GHz. Subsequently, the polynomial of the filter network can be synthesized according to these specifications. In this work, the MADP-030015-1314 p-i-n diode is used with a forward current of 38 mA and reverse voltage of \(-5\) V, which exhibits \(C_0 = 0.78\) pF, \(L_0 = 0.4\) nH, and \(R_0 = 0.49\) Ω. With all \(C_{ij}\) as 0.72 pF, the lumped element values of the filter circuit model can be extracted. After incorporating the circuit models for the p-i-n diodes, the lumped element values of the filtering switch can be obtained and are listed in Table I. The frequency responses of the lumped filtering switch in the ON and OFF states are shown in Fig. 14.

To obtain the high isolation in the OFF state, four TZs in the OFF state are arranged at 3.6, 3.2, 3.1, and 4.3 GHz. With the proposed approximations, the parameters of the distributed circuit can be determined according to the predefined characteristic impedances, as listed in Table II. Fig. 15 illustrates the frequency responses of the ideal distributed circuit model of the filtering switch.

For the prototype of a fourth-order filtering switch, the Rogers RO4003 substrate with a dielectric constant of 2.2, loss tangent of 0.0027, and thickness of 0.813 mm, is used. Fig. 16 presents the layout of the designed fourth-order filtering switch with dimensions marked. The dc biasing circuit includes a high-impedance meander line, a 90° fan stub, and a chip resistor [19], [20]. The final dimensions are listed in Table III, and the photograph of the fabricated switch is shown in Fig. 17.

The simulated and measured results of the fourth-order filtering switch in the ON and OFF states are shown in Fig. 18(a) and (b), respectively. In the ON state, the measured return loss within the passband is greater than 15 dB. In addition, the values of measured insertion losses at the lower band-edge frequency 3.4 GHz, the center frequency 4 GHz, and the upper band-edge frequency 4.6 GHz are 1.51, 2.17, and 8.15 dB, which are 1.05, 1.62, and 4.94 dB more than the simulated results, respectively. The discrepancy of insertion losses can be attributed to the fabrication tolerances.
losses within the ON-state passband at high frequencies exists, which may be caused by the losses from the p-i-n diodes and the bias circuits. In the OFF state, it can be seen that the isolation of the switch is better than 45 dB from 3.4 to 4.6 GHz (i.e., 30% FBW). The overall size of the fourth-order filtering switch (including feedlines and bias circuits) is 62.8 mm × 54.18 mm, namely, 0.84λ0 × 0.722λ0, where λ0 denotes the guided wavelength at 4 GHz in free space.

**B. Fifth-Order Filtering Switch With 50% FBW in the ON State**

The second example is a fifth-order filtering switch with 50% FBW in the ON state. First, the BPF network is synthesized with the specifications of the center frequency of 4 GHz, 50% FBW, fifth-order, 20-dB in-band return loss, and TZs of five resonators located at 7, 6, 6, 6, and 6.7 GHz. By designating \( C_{ix} \) as 0.7, 0.74, 0.74, 0.7, and 0.66 pF, the lumped element values of the filter topology can be extracted.

---

**TABLE III**

<table>
<thead>
<tr>
<th>i</th>
<th>( w_i ) (mm)</th>
<th>( w_{oi} ) (mm)</th>
<th>( w_{ci} ) (mm)</th>
<th>( w_{pi} ) (mm)</th>
<th>( l_i ) (mm)</th>
<th>( l_{ci} ) (mm)</th>
<th>( A ) (mm)</th>
<th>( r_i ) (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1.27</td>
<td>0.75</td>
<td>0.75</td>
<td>0.75</td>
<td>0.8</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>0.75</td>
<td>0.75</td>
<td>0.75</td>
<td>0.75</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>6.39</td>
<td>6.39</td>
<td>3.8</td>
<td>6.39</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>0.43</td>
<td>0.43</td>
<td>0.43</td>
<td>0.43</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>3.8</td>
<td>6.84</td>
<td>5.23</td>
<td>3.88</td>
<td>3.05</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>0.6</td>
<td>1</td>
<td>1.87</td>
<td>2.72</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>7</td>
<td>7.25</td>
<td>6.8</td>
<td>8.16</td>
<td>2.05</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>8</td>
<td>3.97</td>
<td>6.78</td>
<td>2.93</td>
<td>9.3</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>9</td>
<td>0.2</td>
<td>0.8</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10</td>
<td>12.2</td>
<td>1.7</td>
<td>0.5</td>
<td>-</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>11</td>
<td>11.1</td>
<td>0.25</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**TABLE IV**

<table>
<thead>
<tr>
<th>( L_i ) (nH)</th>
<th>( L_{ci} ) (nH)</th>
<th>( C_L ) (pF)</th>
<th>( n_i )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1.16</td>
<td>15.98</td>
<td>0.34</td>
</tr>
<tr>
<td>2</td>
<td>1.71</td>
<td>5.19</td>
<td>0.55</td>
</tr>
<tr>
<td>3</td>
<td>1.57</td>
<td>6.12</td>
<td>0.55</td>
</tr>
<tr>
<td>4</td>
<td>1.61</td>
<td>7.24</td>
<td>0.61</td>
</tr>
<tr>
<td>5</td>
<td>1.7</td>
<td>12.44</td>
<td>0.45</td>
</tr>
<tr>
<td>6</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
</tbody>
</table>

**TABLE V**

<table>
<thead>
<tr>
<th>( Z_0 ) (Ω)</th>
<th>( \theta_{i0} ° )</th>
<th>( \theta_{pi} ° )</th>
<th>( Z_i ) (Ω)</th>
<th>( \theta_{ci} ° )</th>
<th>( Z_{ci} ) (Ω)</th>
<th>( \theta_{ci} ° )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>65.89</td>
<td>23.78</td>
<td>100</td>
<td>68.12</td>
<td>80</td>
<td>7.15</td>
</tr>
<tr>
<td>2</td>
<td>70</td>
<td>37.84</td>
<td>100</td>
<td>47.98</td>
<td>80</td>
<td>9.44</td>
</tr>
<tr>
<td>3</td>
<td>70</td>
<td>34.2</td>
<td>100</td>
<td>47.97</td>
<td>80</td>
<td>7.19</td>
</tr>
<tr>
<td>4</td>
<td>70</td>
<td>35.41</td>
<td>100</td>
<td>47.6</td>
<td>80</td>
<td>9.18</td>
</tr>
<tr>
<td>5</td>
<td>70</td>
<td>37.61</td>
<td>100</td>
<td>54.73</td>
<td>80</td>
<td>7.43</td>
</tr>
<tr>
<td>6</td>
<td>56.91</td>
<td>25.31</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
</tbody>
</table>

**TABLE VI**

<table>
<thead>
<tr>
<th>( w_i )</th>
<th>( w_{oi} )</th>
<th>( w_{ci} )</th>
<th>( w_{pi} )</th>
<th>( l_i )</th>
<th>( l_{ci} )</th>
<th>( l_{pi} )</th>
<th>( w_i )</th>
<th>( l_i )</th>
<th>( r_i )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1.11</td>
<td>0.99</td>
<td>0.99</td>
<td>0.99</td>
<td>0.99</td>
<td>1.45</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>0.75</td>
<td>0.75</td>
<td>0.75</td>
<td>0.75</td>
<td>0.75</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>3.8</td>
<td>3.8</td>
<td>6.39</td>
<td>6.39</td>
<td>3.8</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>0.43</td>
<td>0.43</td>
<td>0.43</td>
<td>0.43</td>
<td>0.43</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>2.41</td>
<td>5.67</td>
<td>4.65</td>
<td>5.18</td>
<td>5.31</td>
<td>3.02</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>1.33</td>
<td>1.75</td>
<td>0.33</td>
<td>0.67</td>
<td>1.47</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>7</td>
<td>3.98</td>
<td>5.41</td>
<td>10.07</td>
<td>7.86</td>
<td>4.75</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>8</td>
<td>9.07</td>
<td>9.11</td>
<td>5.48</td>
<td>7.19</td>
<td>6.86</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>9</td>
<td>0.2</td>
<td>0.2</td>
<td>0.8</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10</td>
<td>9.8</td>
<td>8</td>
<td>0.5</td>
<td>1.4</td>
<td>-</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>11</td>
<td>10</td>
<td>10</td>
<td>0.25</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
After incorporating the circuit models for the p-i-n diodes, the lumped element values of the filtering switch can be obtained as listed in Table IV. By arranging TZs in the OFF state at 4.8, 3.5, 3, 3.2, and 4.6 GHz, the parameters of distributed element circuit can be derived and listed in Table V. Fig. 19 illustrates the frequency responses of the lumped circuit and the ideal distributed circuit of the filtering switch.

Fig. 20 presents the layout of the designed fifth-order filtering switch, and the final dimensions are listed in Table VI. The photograph of the fabricated switch is shown in Fig. 21. The simulated and measured results of the fifth-order switch are shown in Fig. 22. In the ON state, as seen in Fig. 22(a), the measured return loss within the passband is greater than 14 dB. In addition, the measured insertion losses at 3, 4, and 5 GHz are 0.97, 1.45, and 7.27 dB, respectively. Similar to the fourth-order filtering switch, the larger discrepancy of insertion losses between simulated and measured results can be observed at higher frequencies in Fig. 22(a). In the OFF state, the isolation of the switch is better than 30 dB from 3 to 5 GHz (i.e., 50% FBW), as illustrated in Fig. 22(b).
TABLE VII

<table>
<thead>
<tr>
<th>Type</th>
<th>Technology</th>
<th>Frequency (GHz)</th>
<th>FBW</th>
<th>ON state IL(dB)</th>
<th>TZ</th>
<th>OFF state Isolation(dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[2]</td>
<td>SPST</td>
<td>On-chip + FET</td>
<td>24.2-36.3</td>
<td>40.8%</td>
<td>1.75-2.25</td>
<td>No</td>
</tr>
<tr>
<td>[8]</td>
<td>SPDT</td>
<td>On-chip + FET</td>
<td>40</td>
<td>8%</td>
<td>&gt;5.1</td>
<td>Yes</td>
</tr>
<tr>
<td>[10]</td>
<td>SPDT</td>
<td>DR + P-I-N Diode</td>
<td>1.861</td>
<td>1.16%</td>
<td>0.51*</td>
<td>No</td>
</tr>
<tr>
<td>[11]</td>
<td>SPQT</td>
<td>CR + P-I-N Diode</td>
<td>2.37</td>
<td>2.9%</td>
<td>0.93*</td>
<td>Yes</td>
</tr>
<tr>
<td>[13]</td>
<td>SPST</td>
<td>PC + P-I-N Diode</td>
<td>1.5</td>
<td>5%</td>
<td>3.9*</td>
<td>No</td>
</tr>
<tr>
<td>[14]</td>
<td>SPDT</td>
<td>PC + P-I-N Diode</td>
<td>0.612-1.088</td>
<td>19.5%</td>
<td>2.3-2.8</td>
<td>No</td>
</tr>
<tr>
<td>[15]</td>
<td>SPDT</td>
<td>PC + P-I-N Diode</td>
<td>0.89-1.82</td>
<td>8.3%</td>
<td>0.97*</td>
<td>Yes</td>
</tr>
<tr>
<td>This work</td>
<td>SPST</td>
<td>PC + P-I-N Diode</td>
<td>3.4-4.6</td>
<td>30%</td>
<td>1.51*</td>
<td>Yes</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>3-5</td>
<td>50%</td>
<td>0.97*</td>
<td>Yes</td>
</tr>
</tbody>
</table>

Notes: DR means dielectric resonator, CR means coaxial resonator, # denotes the insertion loss at center frequency, * denotes the minimum insertion loss within passband.

Fig. 22. Simulated and measured S-parameters of prototyped fifth-order filtering switch in (a) ON state and (b) OFF state.

size of the fifth-order filtering switch (including feedlines and bias circuits) is 66.24 mm × 53.9 mm (0.88λ₀ × 0.719λ₀). For further illustration, a table of comparisons between the reported filtering switches and the proposed works is given in Table VII.

VII. CONCLUSION

In this article, a novel wideband filtering switch is proposed together with the direct synthesis and design processes. By incorporating the circuit model of the p-i-n diodes in the resonators of the filtering network, the circuit models of the filtering switch operating in the ON states can be synthesized according to the specifications. Moreover, the ideal transformers are introduced to designate the capacitances and inductances of each resonator with proper values directly, which ensures all elements of the filtering switch are feasible. Then, the lumped filtering switch embedding p-i-n diodes will be transformed into distributed circuits in a wide frequency range. Due to the frequency-variant characteristic of TLs, the TZs in the OFF state can be adjusted to a suitable position after converting the lumped element circuit to a distributed element circuit. Therefore, high isolation in the desired frequency band can be obtained by controlling the TZs of the distributed switch in the OFF state. To validate the proposed filtering switch as well as the synthesis and design process, two filtering switches with FBWs of 30% and 50% are designed and measured. Good agreement between the simulated and measured results is obtained.

REFERENCES

Dr. Fan was the Japan Society for the Promotion of Science (JSPS) Fellow with the Department of Communications Engineering, Graduate School of Engineering, Tohoku University, from 2019 to 2021. He received the Fellowship from JSPS. He received the UESTC Outstanding Graduate Award in 2009 and 2015. He was a recipient of the National Graduate Student Scholarship in 2012, 2013, and 2014 from the Ministry of Education, China. He has been serving as an Editorial Board Member for the AEUInternational Journal of Electronics and Communications and Electronics (MDPI). Also, he has been serving as an Associate Editor for IEEE ACCESS, IET Microwaves, Antennas & Propagation, and Electronics Letters.

Xiaoyu Feng received the B.Eng. degree from Xi’an Jiaotong University, Xi’an, China, in 2020, where he is currently pursuing the M.Eng. degree with the School of Information and Communication Engineering.

His research interests include microwave and millimeter-wave filters and circuits.

Qiang Chen (Senior Member, IEEE) received the B.E. degree from Xidian University, Xi’an, China, in 1986, and the M.E. and D.E. degrees from Tohoku University, Sendai, Japan, in 1991 and 1994, respectively. He is currently the Chair Professor with the Electromagnetic Engineering Laboratory, Department of Communications Engineering, School of Engineering, Tohoku University. His primary research interests include antennas, microwave and millimeter waves, antenna measurements, and computational electromagnetics.

Dr. Chen is an IEICE Fellow. He received the Best Paper Award and the Zenichi Kiyasaku Award in 2009 from the Institute of Electronics, Information and Communication Engineers (IEICE). He served as the Chair for the IEICE Technical Committee on Photonics-Applied Electromagnetic Measurement from 2012 to 2014, the IEICE Technical Committee on Wireless Power Transfer from 2016 to 2018, and the Tokyo Chapter of the IEEE Antennas and Propagation Society from 2017 to 2018. He is now the Chair of the IEICE Technical Committee on Antennas and Propagation.

Authorized licensed use limited to: Heriot-Watt University. Downloaded on August 10, 2023 at 07:42:01 UTC from IEEE Xplore. Restrictions apply.