Analysis and Design of Bandpass Single-Pole–Double-Throw FET Filter-Integrated Switches

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Abstract—This paper proposes a method to integrate a single-pole–double-throw (SPDT) switch and a quarter-wavelength bandpass filter. A 1-GHz SPDT hybrid switch and a 60-GHz pseudomorphic HEMT monolithic-microwave integrated-circuit SPDT switch with 30% fractional bandwidth are demonstrated. The 1-GHz SPDT switch achieves 1.5-dB insertion loss and 20-dB isolation at center frequency. For the 60-GHz SPDT switch, the measured insertion loss is lower than 2.5 dB and the isolation is higher than 27 dB. The low insertion loss and high isolation show that no performance is degraded when integrating the filter function. The analysis of the power performance is also described. Using the device dc- I/V curves, the power compression point can be predicted.

Index Terms—Bandpass filter, monolithic microwave integrated circuit (MMIC), power compression point, single-pole double-throw (SPDT).

I. INTRODUCTION

RADIO-FREQUENCY (RF) switches are important in time division duplex wireless communication systems. Since a filter is required in front of a switch when integrating a wireless system, the integration of filter and switch functions has the advantages of reduction of insertion loss, as well as complexity. To achieve good switching performance, the nonideal off-state capacitance and on-state resistance need to be taken into considerations. In [1], the nonideal effects of the device are first applied for the 2.69-GHz ring-filter integrated with a single-pole–single-throw (SPST) switch. In [2], the concept of the filter-integrated switch (FIS), which gives the systematic design approach to implement the switch using the filter synthesis of a 1-GHz quarter-wavelength bandpass filter, was proposed.

To design a single-pole–double-throw (SPDT) switch, an impedance transformer is usually required to combine two SPST switches so that the impedance of the off-state SPST switch will not influence the insertion loss of the on-state SPST switch [3], [4], [7], [8] in cases when the input/output

impedances of off-state switches are low. In [3], the image filter synthesis was applied in the design of an SPST traveling-wave switch and quarter-wavelength transmission lines applied as an impedance transformer. However, the frequency response of the entire SPDT switch cannot be synthesized. In [12], a narrowband hybrid SPDT switch using diodes with integration of a coupled-resonator filter was demonstrated by the shared resonator technique. The hybrid switch exhibits 5% fractional bandwidth (FBW) because of the filter topology. However, the topology in [12] is not suitable for wide-bandwidth application.

In this paper, we expand the concept of field-effect transistor (FET) FISs from SPST [2] to SPDT switches. The quarter-wavelength impedance transformers are integrated in the quarter-wavelength bandpass filter. A 1-GHz hybrid and a 60-GHz monolithic-microwave integrated-circuit (MMIC) SPDT switches are both implemented to verify the design concept. The SPDT switches exhibit 30% FBW and are suitable for the wide-bandwidth communication system. A simple and general analysis of the power compression is also described. The analysis in [2] is only derived for SPST switches. On the other hand, a more general analysis for n -port passive nonlinear circuits is proposed here. This analysis can not only be applied to this SPDT switch, but also to other n -port passive switches.

II. CONCEPT OF SPDT FIS

The concept of the SPST FIS has been proposed in [2]. To implement an SPDT FIS, the impedance of the off-state SPST switch and the impedance transformer that combines two SPST switches have to be designed as part of the filter. Fig. 1(a) illustrates the circuit block diagram of the conventional SPDT switch. It is composed of two SPST switches and one impedance transformer. For the shunt switch, quarter-wavelength transformers are used to transfer the low impedance of the off-state SPST switch to high impedance [4]. Although the filter analysis was used in the low-pass SPST switch design, the frequency responses of the entire SPDT switch were not synthesized due to the bandpass impedance transformer. However, in our proposed filter-integrated SPDT switch, the impedance transformer and the SPST switches are designed for bandpass frequency response, as shown in Fig. 1(b).

Fig. 2(a) shows the circuit schematic of the SPDT FIS. It is composed of five shunted resonators and four series quarter-wavelength transmission lines. One short stub shunted resonator directly connects to Port 1. Other four shunted resonators are constructed by parallel transistors and short stubs.

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Fig. 1. (a) Circuit block diagram of conventional SPDT switches. (b) Concept of the filter-integrated SPDT switch.

Fig. 2(b) illustrates the equivalent circuit of the SPDT switch when V_{c1} is below the threshold voltage and V_{c2} is higher than the threshold voltage. Since V_{c1} is below the threshold voltage, the transistors are turned off and can be modeled to capacitors (C_{off1} and C_{off2}). Similarly, the high V_{c2} turned on the transistors and the transistors are modeled as resistors (G_{on1} and G_{on2}). Comparing Fig. 2(b) and (c), we can treat the through path of the SPDT switch (Port 2 to Port 1) as the third-order quarter-wavelength short-circuited stubs bandpass filter shown in Fig. 2(c). To make the frequency response of the third-order quarter-wavelength short stubs bandpass filter equal to the through path of the SPDT switch, two series quarter-wavelength transmission lines between Port 1 and Port 2 in Fig. 2(b) correspond to the series transmission lines in Fig. 2(c). Furthermore, the susceptance and their frequency derivatives at the center frequency ω_0 of three shunt resonators (Y_{R1} , Y_{R2} , and Y_{R3}) of the switch have to be identical to those of the bandpass filter (Y_{Rf1} , Y_{Rf2} , and Y_{Rf3}). Thus, the relation of the design parameters between the switch and bandpass filter can be obtained as

$$\text{Im}(Y_{R1}) = \text{Im}(Y_{Rf1}) = \omega_0 C_{off1} - Y_1 \cot \theta_1 = 0 \quad (1)$$

$$\text{Im}(Y_{R2}) = \text{Im}(Y_{Rf2}) = \omega_0 C_{off2} - Y_2 \cot \theta_2 = 0 \quad (2)$$

$$\text{Im}(Y_{R3}) = \text{Im}(Y_{Rf3}) = \text{Im} \left(Y_{12}^2 / Y_{iso} - jY_3 \cot \theta_3 \right) = 0. \quad (3)$$

In (3), Y_{iso} is the admittance from the turned-on transistor to the isolation port and is expressed with the system characteristic admittance Y_0 . Since the transistors are turned on, G_{on2} and G_{on1} are large conductances, and thus, it is reasonable to approximate Y_{iso} to G_{on2} as follows:

$$Y_{iso} = G_{on2} - jY_2 \cot \theta_2 + \frac{Y_{12}^2}{Y_0 + G_{on1} - jY_1 \cot \theta_1} \cong G_{on2}. \quad (4)$$

Fig. 2. (a) Proposed SPDT FIS. (b) Equivalent circuit of the SPDT switch. (c) Circuit schematic of a third-order quarter-wavelength short-circuited stubs bandpass filter.

Since the frequency derivatives of the susceptances should be equal, we can obtain

$$Y_{Rf1} \frac{\pi}{4\omega_0} = C_{off1} + \frac{Y_1 \theta_1}{\omega_0} \csc^2 \theta_1 \quad (5)$$

$$Y_{Rf2} \frac{\pi}{4\omega_0} = C_{off2} + \frac{Y_2 \theta_2}{\omega_0} \csc^2 \theta_2 \quad (6)$$

$$Y_{Rf3} \frac{\pi}{4\omega_0} = \frac{Y_3 \theta_3}{\omega_0} \csc^2 \theta_3 - \frac{\pi Y_{12} (G_{on2}^2 - Y_{12}^2)}{2\omega_0 G_{on2}^2}. \quad (7)$$

Applying the filter synthesis, the design parameters of the bandpass filter (Y_{12} , Y_{Rf1} , Y_{Rf2} , and Y_{Rf3}) can be obtained. With given device size, C_{off1} , C_{off2} , C_{off3} , G_{on1} , and G_{on2} are determined. Consequently, the design parameters of the switch (Y_{12} , Y_1 , Y_2 , Y_3 , θ_1 , θ_2 , and θ_3) are obtained from (1)–(7).

For the insertion loss of the through path at ω_0 , only one lossy component G_{on2} is required to be taken into consideration. From (1) to (3), Y_{R1} , Y_{R2} , and Y_{R3} are all zero at ω_0 , the

Fig. 3. Circuit schematic for the calculation of isolation.

insertion loss S_{21} is expressed in terms of Y_{12} , Y_0 , and G_{on2} as follows:

$$S_{21} = -\frac{Y_0}{Y_0 + \frac{Y_{12}^2}{2G_{on2}} - j\frac{Y_3 \cot \theta_3}{2}}. \quad (8)$$

Similarly, the isolation of the isolation path can be calculated. From Fig. 3, the input admittance from the off-state transistor (Y_{thru}) to Port 2 is expressed as

$$Y_{thru} = -Y_{12}^2/Y_0. \quad (9)$$

From (9), the isolation S_{31} is (10), shown at the bottom of this page.

From (8) and (10), it is observed that the insertion loss and isolation can be improved by increasing G_{on1} and G_{on2} . It is also noticed that both the insertion loss and isolation can be improved at the same time. One constraint is that C_{off1} and C_{off2} have to be in a reasonable range to make (1)–(7) solvable. Since G_{on1} , G_{on2} , C_{off1} , and C_{off2} are the on-state channel conductances and off-state capacitances of the transistor and are all proportional to the gatewidth of the device, the device sizes in the switch need to be selected to make (1)–(7) solvable.

It is worth mentioning that in the conventional switch design, there are tradeoffs between insertion loss and the isolation for the device size. Due to the off-state capacitance, the insertion loss will be higher if the device size is increasing. However, in our new approach, the off-state capacitance is one of the design parameters and is to be resonated with the short stubs. Furthermore, by adjusting the impedance and length of the short stubs,

both the susceptance and frequency differentiation of the susceptance can be equal to those of the quarter-wavelength bandpass filter and, thus, the frequency response of the switch may be synthesized.

III. POWER ANALYSIS

There are two steps to analyze the power performance. The first step is to describe the nonlinear device, and the second is to define the relation between power compression and the nonlinear conductance. In [2], the relation between power compression and the nonlinear conductance was derived only for the SPST FIS, but is not suitable for SPDT switches. To analyze our SPDT switches, a more general analysis is performed for an arbitrary n -port network. Thus, this method is not only suitable for this SPDT switches, but also for n -port passive switches.

It is noticed that the piecewise linear approximation is used for the power analysis, which can estimate the power handling of the circuit fairly accurately. However, the piecewise linear approximation cannot be used to estimate the intermodulation since the third-order nonlinear effect is needed in the calculation of the intermodulation.

A. Piecewise Linear Approximation

Fig. 4(a) illustrates the measured IV curves of the off-state passive HEMT [13]. By approximating the IV curves with two segments, I_{ds} can be expressed as

$$I_{ds} = \begin{cases} 0, & \text{if } V_{ds} > -V_x \\ G_x(V_{ds} + V_x), & \text{if } V_{ds} \leq -V_x. \end{cases} \quad (11)$$

Assuming V_{ds} is a sinusoid voltage signal,

$$V_{ds} = V_p \cos \omega t. \quad (12)$$

The amplitude of the fundamental term of I_{ds} can be evaluated by the Fourier transformation

$$I_{ds} = \frac{G_x}{\pi} \left(\frac{V_p}{2} (2\phi + \sin 2\phi) - 2V_x \sin \phi \right). \quad (13)$$

In (13), ϕ is the conduction angle and is defined as

$$\phi = \cos^{-1} \left(\frac{V_x}{V_p} \right). \quad (14)$$

From (13), the nonlinear conductance of the off-state transistor (G_{off}^{nl}) can be expressed in terms of G_x , θ , and V_p as follows:

$$G_{off}^{nl} = \begin{cases} 0, & \text{if } V_x > V_p \\ G_x \left(\frac{(2\phi - \sin 2\phi)}{2\pi} \right), & \text{if } V_x \leq V_p. \end{cases} \quad (15)$$

$$S_{31} = -\frac{2Y_0}{\left(G_{on1} + Y_0 - jY_1 \cot \theta_1 \right) + \left(\frac{Y_{21}^2 + (G_{on1} + Y_0 - jY_1 \cot \theta_1)(G_{on2} - jY_2 \cot \theta_2)}{2Y_0 - jY_3 \cot \theta_3} \right) \frac{Y_{21}^2}{Y_{21}^2}} \quad (10)$$

Fig. 4. Measured and piecewise linear approximation of the FET (NE32584C [13]) in: (a) turned-off state and (b) turned-on state.

For the measured on-state passive HEMT shown in Fig. 4(b), the I/V curves of the on-state transistor can be expressed as

$$I_{ds} = \begin{cases} G_{y1}V_{ds}, & \text{if } -V_{z1} < V_{ds} < V_{y1} \\ G_{y2}V_{ds} + (G_{y1} - G_{y2})V_{y1}, & \text{if } V_{y1} \leq V_{ds} < V_{y2} \\ G_{y3}V_{ds} + (G_{y2} - G_{y3})V_{y2} + (G_{y1} - G_{y2})V_{y1}, & \text{if } V_{ds} > V_{y2} \\ G_{z2}V_{ds} + (G_{y1} - G_{z2})V_{z1}, & \text{if } -V_{z1} > V_{ds} \geq -V_{z2} \\ G_{z3}V_{ds} + (G_{z2} - G_{z3})V_{z2} + (G_{y1} - G_{z2})V_{z1}, & \text{if } V_{ds} < -V_{z2}. \end{cases} \quad (16)$$

Therefore, the nonlinear conductance of the on-state transistor (G_{on}^{nl}) is derived as

$$G_{on}^{nl} = \begin{cases} G_{y1} + \frac{1}{\pi} \left(\begin{aligned} &(G_{y2} - G_{y1})(\phi_{y1} - \cos \phi_{y1} \sin \phi_{y1}) \\ &+ (G_{y3} - G_{y2})(\phi_{y2} - \cos \phi_{y2} \sin \phi_{y2}) \\ &+ (G_{z2} - G_{y1})(\phi_{z1} - \cos \phi_{z1} \sin \phi_{z1}) \\ &+ (G_{z3} - G_{z2})(\phi_{z2} - \cos \phi_{z2} \sin \phi_{z2}) \end{aligned} \right) \end{cases} \quad (17)$$

Fig. 5. Schematic of a general n -port nonlinear circuit with driving power source.

Fig. 6. Simplified passive HEMT small-signal model of NE32584C.

Fig. 7. Circuit schematic of the 1-GHz hybrid SPDT switch.

In (14), ϕ_{y1} , ϕ_{y2} , ϕ_{z1} , and ϕ_{z2} are defined as

$$\begin{aligned} \phi_{y1} &= \begin{cases} \cos^{-1} |V_{y1}/V_p|, & \text{if } V_{y1} < V_p \\ 0, & \text{if } V_{y1} \geq V_p \end{cases} \\ \phi_{y2} &= \begin{cases} \cos^{-1} |V_{y2}/V_p|, & \text{if } V_{y2} < V_p \\ 0, & \text{if } V_{y2} \geq V_p \end{cases} \\ \phi_{z1} &= \begin{cases} \cos^{-1} |V_{z1}/V_p|, & \text{if } V_{z1} < V_p \\ 0, & \text{if } V_{z1} \geq V_p \end{cases} \\ \phi_{z2} &= \begin{cases} \cos^{-1} |V_{z2}/V_p|, & \text{if } V_{z2} < V_p \\ 0, & \text{if } V_{z2} \geq V_p. \end{cases} \end{aligned} \quad (18)$$

With the piecewise linear parameters and the implementation of (15) and (17), the relation between voltage amplitude and the equivalent admittance (G_{off}^{nl} and G_{on}^{nl}) can be analytically evaluated.

B. Analysis of the Power Compression

Fig. 5 illustrates the general schematic of an n -port nonlinear circuit with power source (P_{in}) driving to Port 1. The

Fig. 8. Admittance of the resonator of filter (Y_{RF1} , Y_{RF2} , and Y_{RF3}) and resonator of switch (Y_{R1} , Y_{R2} , and Y_{R3}).

Fig. 9. Comparison of the simulated frequency responses between the 1-GHz ideal third-order filter and the 1-GHz hybrid SPDT switch.

Fig. 10. 1-GHz hybrid SPDT switch.

schematic is composed of one linear network and several nonlinear conductances ($G_1^{nl} \sim G_n^{nl}$). The nonlinear conductance are assumed to be constants when the voltage swings ($v_1 \sim v_p$) are small, and after the voltage swings become large, the components are driven into the nonlinear region since ($G_1^{nl} \sim G_p^{nl}$) are weakly nonlinear at the input x -dB compression point ($P_{x\text{dB}}$, $0 < x \leq 1$). The effect of ($G_1^{nl} \sim G_p^{nl}$) can be discussed separately. Considering only G_1^{nl} , G_1^{nl} input,

(a)

(b)

(c)

Fig. 11. Simulated and measured results of the 1-GHz hybrid SPDT switch. (a) Insertion loss and output return loss of the on-state switch. (b) Isolation and output return loss of the off-state switch. (c) Input return of the switch.

$P_{x\text{dB}}$ can be obtained by sweeping G_1^{nl} for the n -port network parameters ($S_{mn}(G_1^{nl})$) since G_1^{nl} is the function of v_1 , while v_1 is also a function of P_{in} , P_{in} at x -dB compression dominated by G_1^{nl} can be determined. After investigating the conductance one by one, the input $P_{x\text{dB}}$ of the n -port nonlinear circuit can be determined by the lowest x -dB compression point and the dominant nonlinear conductance can be decided.

It is worth noticing that no nonlinear simulation tool is used to find the x -dB compression point in this method. For the passive HEMT, the nonlinear effect of the ($G_1^{nl} \sim G_p^{nl}$) is analyt-

TABLE I
DEVICE PARAMETERS OF THE DEVICE OBTAINED IN PIECEWISE
LINEAR APPROXIMATION (NE32584C)

Parameters	Value	Parameters	Value
V_x (V)	2.3	G_x (mS)	81.9
V_{y1} (V)	0.1	G_{y1} (mS)	175
V_{y2} (V)	0.4	G_{y2} (mS)	100
V_{y1} (V)	0.2	G_{y3} (mS)	12
V_{y2} (V)	0.7	G_{y2} (mS)	90
		G_{y3} (mS)	11

Fig. 12. Schematic of the 1-GHz SPDT hybrid switch for power analysis.

ically expressed in (15) and (17). Thus, only the IV curves and linear network analysis tool are required to obtain the power performance.

IV. EXPERIMENTAL RESULTS

A. 1-GHz Hybrid SPDT Switch

The devices used for this experiment are a packaged heterojunction field-effect transistor (HJFET) (NE32584C [13]). Fig. 6 illustrates the simplified passive HEMT small-signal model [4] of NE32584C. When the gate voltage changes from -3 to 0 V, the passive HEMT is switched from the off to the on state and has 0.52 -S channel conductance (G_{on}). Fig. 7 shows the complete circuit schematic of the 1-GHz SPDT switch. This switch is realized on an FR4 printed circuit board (PCB) board. From (8), it can be observed that the insertion loss is dominated by the turn-on conductance of M_3 . Therefore, the three-parallel devices are selected as M_2 and M_3 to reduce the turn-on conductance from 0.52 to 0.17 S.

The SPDT switch is designed for 1-GHz center frequency, 30% FBW, and 0.01-dB equal ripple frequency response. In the filter synthesis, the parameters of the filter prototype in Fig. 2(c) and the parameter Y_{12} are determined. Based on the design equations (1)–(3) and (5)–(7), the admittance and frequency derivatives of each shunt resonator (Y_{R1} , Y_{R2} , and Y_{R3}) have to be equal to those of a third-order quarter-wavelength bandpass filter (Y_{RF1} , Y_{RF2} , and Y_{RF3}) and, thus, the parameters Y_1 , Y_2 , and Y_3 can be determined. Fig. 8 shows the admittance of the resonators (Y_{RF1} , Y_{RF2} , and Y_{RF3}) and those of the switches (Y_{R1} , Y_{R2} , and Y_{R3}) versus frequency. It can be observed that admittance of the resonators of the switch are close to those of the filter when the frequency is lower than 1.3 GHz so that the conditions in (1)–(3) and (5)–(7) are satisfied.

(a)

(b)

Fig. 13. (a) Compression of Port 2 to Port 1 insertion loss versus nonlinear conductances of 1-GHz SPDT hybrid switch. (b) Calculated nonlinear conductance versus input power (P_{in}) at Port 2 of 1-GHz SPDT hybrid switch.

In this 1-GHz circuit design, the parasites in the package and layout discontinuities are taken into consideration. Fig. 9 shows the frequency responses of the ideal third-order bandpass filter and those of the simulated through path of the switch (Port 1 to Port 2 with $V_{c1} = -3$ V and $V_{c2} = 0$ V). The frequency responses of the switch are almost the same. The simulated results show 1-dB insertion loss at the center frequency and return losses better than 23 dB. Fig. 10 shows a chip photograph of the SPDT switch. Fig. 11(a) shows the measured and simulated results of the through path of the switch. At 1-GHz center frequency, the measured insertion loss is 1.5 dB and the return loss is better than 25 dB. Fig. 11(b) shows the performance of the isolation path of the switch (Port 1 to Port 3 with $V_{c1} = -3$ V and $V_{c2} = 0$ V). The measured isolation is better than 27 dB at the center frequency. Fig. 11(c) is the return loss of Port 1, which shows better than 25-dB measured return loss at the center frequency. The measured results are close to the simulated results when the frequency is below 1.5 GHz. The discrepancies between the simulation and measurement above 1.5 GHz may be due to the inaccuracy of the parasites of the packaged devices at higher frequencies.

The power compression of the SPDT switch in the center frequency (1 GHz) is analyzed. The device parameters of the piecewise linear approximation are obtained and listed in Table I. In

Fig. 14. (a) Compression of isolation from Port 3 to Port 1 isolation versus nonlinear conductances of 1-GHz SPDT hybrid switch. (b) Calculated nonlinear conductance versus input power (P_{in}) at Port 3 of 1-GHz SPDT hybrid switch.

TABLE II
SIMPLIFIED PASSIVE HEMT SMALL-SIGNAL MODEL AND THE PARAMETERS
OBTAINED FROM PIECEWISE LINEAR APPROXIMATION PARAMETERS
FOR A TWO-FINGER 150-nm pHEMT DEVICE

Parameters	Value	Parameters	Value
L_{d1} (pH)	35.7	V_x (V)	2.35
C_{off} (fF)	46.97	V_{y1} (V)	0.1
G_{on} (S)	0.145	V_{y2} (V)	0.35
G_{off} (S)	0	V_{y1} (V)	0.4
L_{s1} (pH)	1.35	V_{y2} (V)	1
R_d (Ω)	2.36	G_x (mS)	80
		G_{y1} (mS)	140
		G_{y2} (mS)	80
		G_{y3} (mS)	12
		G_{z2} (mS)	69
		G_{z3} (mS)	30

the schematic for power analysis, shown in Fig. 12, M_1 and M_2 are off-state transistors and M_3 and M_4 are on-state transistors.

The nonlinear channel conductance ($G_1^{nl}-G_4^{nl}$) of each transistors are analyzed to predict the power compression. To find the relations between the insertion loss from Port 2 to Port 1 and each nonlinear conductance, the insertion losses are simulated with sweeping $G_1^{nl}-G_4^{nl}$, and are illustrated in Fig. 13(a). It can be observed that G_1^{nl} , G_2^{nl} , and G_3^{nl} are possible nonlinear elements to affect input P_{1dB} (they are 5, 5, and 73 mS, respectively).

Fig. 15. Output power versus input power for insertion loss and isolation of the 1-GHz hybrid SPDT switch.

Fig. 16. Circuit schematic of the 60-GHz MMIC SPDT switch.

Fig. 17. 60-GHz MMIC SPDT switch with chip size of (2 mm × 1 mm).

The next step is to find an input power level such that the nonlinear conductance satisfies the condition of 1-dB compression. The branch voltage of different power levels can be simulated by the linear simulation tool with the conductance satisfying the condition of 1-dB compression. Using (15) and (17), the nonlinear conductance under different input power [$G_1^{nl}(v_1(P_{in}))-G_4^{nl}(v_4(P_{in}))$] are estimated. Fig. 13(b) shows the nonlinear conductance versus input power. It is observed that P_{1dB} is 19.9 and 22.7 dBm by only considering G_2^{nl} and G_1^{nl} , respectively. The nonlinear conductance of G_3^{nl} still does not decrease to 73 mS for 24-dBm input power. Thus, the actual input P_{1dB} of the SPDT switch is 19.9 dBm (lower one between 19.9–22.7 dBm), and is determined by G_2^{nl} . Therefore, the 1-dB compression of the insertion loss is dominated by G_2^{nl} .

Fig. 18. Simulated and measured results of the 60-GHz MMIC SPDT switch. (a) Insertion loss and output return loss of the on-state switch. (b) Isolation and output return loss of the off-state switch. (c) Input return of the switch.

Regarding the isolation from Port 3 to Port 1, the 1-dB compression occurs when G_2^{nl} , G_3^{nl} , and G_4^{nl} are 6.6, 166 and 57 mS, respectively, and are shown in Fig. 14(a). The relation between the input power and nonlinear conductances are illustrated in Fig. 14(b). It can be observed that only G_4^{nl} limits the isolation when the input power is below 24 dBm and the predicted P_{1dB} is 19 dBm. Fig. 15 is the measured output power versus input power for insertion loss and isolation at center frequency

Fig. 19. Output power versus input power for insertion loss and isolation of the 60-GHz MMIC SPDT switch.

TABLE III
COMPARISON BETWEEN THE SIMULATED AND MEASURED RESULTS

Parameters	Simu. of 1-GHz Switch	Meas. of 1-GHz Switch	Simu. of 60-GHz Switch	Meas. of 60-GHz Switch
Insertion Loss (dB) @ center frequency	1	1.5	2	2
Isolation (dB) @ center frequency	26	27	27	32
BWF (%)	30	30	30	30
Compression for Insertion Loss (dBm)	19.9 (P_{1dB})	20.5 (P_{1dB})	16 ($P_{0.5dB}$)	15.3 ($P_{0.5dB}$)
Dominant Element for Insertion Loss	G_2^{nl}		G_1^{nl}	
Compression for isolation (dBm)	19 (P_{1dB})	17 (P_{1dB})	19.2 (P_{1dB})	N.A.
Dominant Element for Isolation	G_4^{nl}		G_4^{nl}	

(1 GHz). The measured insertion loss P_{1dB} is 20.5 dBm and isolation P_{1dB} is 187 dBm. Both of them are close to the predicted 19.9-dBm insertion loss P_{1dB} and 19-dBm isolation P_{1dB} , respectively. The input third order intercept point (IIP₃) at 1 GHz with two tones of 100-kHz offset was measured to be 25 dBm.

B. 60-GHz MMIC SPDT Switch

The process used in this design is WIN's 0.15- μ m-high linearity AlGaAs/InGaAs/GaAs pseudomorphic HEMT (pHEMT) MMIC process [5]. In this design, the HEMT device of two fingers with total periphery of 150 μ m (2f150) was used. Following the simplified passive HEMT small-signal model in Fig. 6 and the piecewise linear approximation, the parameters of the HEMT device are listed in Table II.

The target frequency response of this switch is 60-GHz center frequency, 30% FBW, and 0.01-dB third-order equal ripple bandpass filter responses. With the target frequency response, the systematic synthesis approach in Section II is applied to determine the characteristic impedance and length of the transmission lines for the device with a total gatewidth of

TABLE IV
 PREVIOUSLY PUBLISHED HEMT SPDT SWITCHES AROUND 60 GHz

Author	Device	Approach	Freq. Range (GHz)	Insertion Loss (dB)	Isolation (dB)	Chip Size (mm × mm)
Z. M. Tsai Error! Reference source not found.	0.15 μm GaAs HEMT	Traveling-wave concept	15 ~ 135	< 6	> 20	1.35 × 0.5
K.-Y. Lin <i>et al.</i> Error! Reference source not found.	0.15 μm GaAs HEMT	Impedance transformation	53 ~ 61	< 4	> 30	2 × 1
M. Madihian <i>et al.</i> , Error! Reference source not found.	HJFET	Series resonant, shunt	57-61	3.9 (Min)	41 (Max)	3.3 × 1.7
K.-Y. Lin <i>et al.</i> Error! Reference source not found.	0.1 μm GaAs HEMT	Traveling-wave concept	15 ~ 80	< 3.6	> 25	1.5 × 1.5
K.-Y. Lin <i>et al.</i> Error! Reference source not found.	0.15 μm GaAs HEMT	Traveling-wave concept	DC ~ 60	< 3	> 25	1 × 1
G. L. Lan <i>et al.</i> Error! Reference source not found.	0.15 μm GaAs HEMT	Resonant, series	54 ~ 64	< 3.2	> 23	0.8 × 2.45
J. Kim <i>et al.</i> , Error! Reference source not found.	0.15 μm GaAs HEMT	FET-integrated transmission line structure	40 ~ 85	< 2	> 30	1.45 × 1

150 μm. Fig. 16 shows the complete schematic diagram of the 60-GHz MMIC SPDT switch. The discontinuous junctions and parasitic effects of the transmission lines are all characterized with the Sonnet electromagnetic simulation tool [6]. Fig. 17 shows a chip photograph of the SPDT switch with a chip size of 2 mm × 1 mm.

The chip was measured via an on-wafer test. Since that two RF ground-signal-ground (GSG) probes cannot be placed on the same side of the chip for an on-wafer test, one of the output ports is terminated with an on-chip 50-Ω load. In order to measure the complete frequency-response shape of this SPDT switch, three different test sets for different bands (dc to 50 GHz, *V*-band, and *W*-band) are used. Fig. 18(a) illustrates the measured and simulated results of the on-state switch. At the center frequency of 60 GHz, the measured insertion loss is 2.5 dB and the return loss is better than 10 dB. Fig. 18(b) shows the off-state switch performance. The measured isolation is better than 30 dB at center frequency. Fig. 18(c) shows the return loss of the input port, which shows better than 10-dB measured return loss at center frequency. From 45 to 65 GHz, the MMIC SPDT switch has an insertion loss, isolation, and return loss better than 2.2, 30, and 8 dB, respectively. They all reasonably agree with the simulated results.

Fig. 19 illustrates the measured output power versus input power for insertion loss and isolation. Due to the power limitation of the 60-GHz signal source, we can only measure 0.5-dB compression of the insertion loss when the input power is 15.3 dBm. For the isolation, the power compression does not occur when the input power is 15.3 dBm. Using the same analysis procedure in the 1-GHz hybrid SPDT switch, the conductances for the 0.5-dB compression of insertion loss are 1.1 mS for G_1^{nl} , 1.35 mS for G_2^{nl} , or 43 mS for G_3^{nl} by sweeping the nonlinear conductance, and then the nonlinear conductance can be evaluated by (15) and (17) with the swing voltage of the input power at port 2. It is also observed that the dominant component is G_1^{nl} at input power of 16 dBm. Regarding isola-

tion, $G_2^{nl} = 1.22$ mS, $G_3^{nl} = 77$ mS, or $G_4^{nl} = 83$ mS are the conductances for the 0.5-dB compression point of isolation. Thus, the 0.5-dB compression of isolation is estimated to be 19.2 dBm and the dominate element is G_4^{nl} , although we cannot measure this power level.

Table III lists the comparison between the simulated and measured results of the 1-GHz hybrid and the 60-GHz MMIC SPDT switches. The dominate components for the power compression are determined. The simulated results are close to the measured results; therefore, the design approach of the FIS and the predictions of the power performance are verified.

Table IV lists the published HEMT switch around 60 GHz and compares it with our SPDT FIS. The 2.5-dB insertion loss in the passband is the best result, except for [11], which shows that no tradeoffs is required when integrating the filter function into a switch.

V. CONCLUSION

In this paper, the concept of the FIS, which integrates the quarter-wavelength bandpass filter into an SPDT, has been expanded from the SPST switch to the SPDT switch. The systematic design approach is presented so that the design parameters of the SPDT switch can be determined by applying the filter synthesis. A more general analysis to predict the power compression has also been described. The experimental circuits of hybrid 1- and 60-GHz MMIC SPDT switches both have 30% FBW, and a 0.1-dB equal ripple filter frequency response was built and measured. The 1-GHz switch exhibits 1.5-dB insertion loss and 27-dB isolation at center frequency, while the *V*-band switch has 2-dB insertion loss and 32-dB isolation at center frequency. For the 1-GHz SPDT switch, the measured P_{1dB} for the insertion loss is 20.5 dBm and the isolation P_{1dB} is 17 dBm. For the 60-GHz SPDT switch, the insertion loss $P_{0.5B}$ is 15.3 dBm. The measured data are close to the predicted results. The method also provides a simple and precise way to obtain the power compression point and the dominant components without generating

nonlinear models and nonlinear simulation tools, which is useful to design for the power performance of SPDT switches.

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