Radio Frequency Magnet-Free Circulators Based on Spatiotemporal Modulation of Surface Acoustic Wave Filters

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Abstract—In this article, a new generation of magnet-free circulators with high performance is proposed. Circulators are crucial devices in modern communication systems due to their ability to enable full-duplexing and double the spectral efficiency directly in the physical layer of the radio-frequency (RF) front end. Traditionally, the Lorentz reciprocity is broken by applying the magnetic bias to ferrite materials; therefore, conventional circulators are bulky and expensive. In this article, this problem is addressed by replacing the magnetic bias with periodic spatiotemporal modulation. Compared to previous works, the proposed circulator is constructed using surface acoustic wave (SAW) filters instead of transmission lines (TLs), which reduces the modulation frequency by at least a factor of 20 and ensures ultra-low power consumption and high linearity. The miniaturized high quality (Q) factor SAW filters also lead to a low-loss nonreciprocal band with strong isolation (IX) and broad bandwidth (BW) on a chip scale, therefore addressing such limitations in previous magnet-free demonstrations. Furthermore, compared to the conventional differential circuit configuration, a novel quad configuration is developed, which doubles the intermodulation-free BW.

Index Terms—Circulators, magnet free, nonreciprocity, surface acoustic wave (SAW).

I. INTRODUCTION

IN MODERN communications, full-duplex operation can, in principle, double the spectrum efficiency of traditional half-duplex systems [1]–[7]. In such systems, circulators are crucial to transmit the RF signal in one direction while blocking it in the opposite direction. Traditionally, these nonreciprocal devices require a strong magnetic bias in order to break time-reversal symmetry, leading to bulky and expensive device, which partially contribute to the long-held false assumption that full-duplex systems are impractical. Transistors have been considered as potential candidates to build magnet-free circulators [8]–[10]; however, they lead to fundamental limitations in terms of linearity and noise performance. Recently, a new class of magnet-free circulators based on time-varying circuits was proposed [11]–[38], reviving the hope to realize fully integrated full-duplex systems in the near future. In time-varying circulators, the magnetic bias is replaced by periodic spatiotemporal modulation, thus having the potential to achieve nonreciprocity at a much smaller form factor. Varactors are considered as suitable components to apply periodic modulation [11], [12]. Nevertheless, the use of varactors comes with some fundamental challenges, such as poor linearity and complicated modulation networks; therefore, varactors-based circulators are limited in real applications, which require an integrated circulator system with high power handling. On the other hand, radio frequency (RF) switches can provide a better linearity and alleviate the necessity of using complicated modulation networks; therefore, they can overcome the limitations that varactors suffer from. However, a tradeoff between the switching speed and the linearity imposes a stringent requirement on the modulation frequency. In [15] and [16], TLs have been periodically modulated to break reciprocity, however, due to the limited loaded Q-factor, large modulation frequencies are required, thus small RF switches with fast switching speed are used, resulting in challenges in terms of linearity. Even though [15] improves the linearity by suppressing the voltage swing at receiver port, this results in an asymmetrical circulator response, leaving the network sensitive to impedance mismatches. In fact, all the aforementioned demonstrations of magnet-free circulators, either based on varactors or RF switches, face the challenge of requiring large modulation frequencies or narrow BW of operation. Besides poor linearity and leakage of RF power into the modulation network, the large modulation frequency also leads to large power consumption, which is another critical metric for communication systems.

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In this article, we consider microacoustic filters periodically modulated in time to break reciprocity. Fig. 1 shows an example of microacoustic filters based on a ladder configuration of microacoustic resonators. Unlike TLs, microacoustic filters convert electromagnetic waves to acoustic waves using piezoelectric materials. Due to lower loss in the mechanical domain, high $Q$-factors are guaranteed. For a simple resonator circuit governed by a second-order differential equation [Fig. 1 (right inset)] with low damping factor [e.g., surface acoustic wave (SAW) resonators/bulk acoustic wave (BAW) resonators/contour-mode resonators (CMRs) [38]/cross-sectional lame mode resonators (CLMRs) [39]–[41], etc.], the group delay can be approximated as

$$t_d \approx \frac{2Q}{\omega_0}$$

where $\omega_0$ is the center frequency. Therefore, a large group delay is guaranteed by a large $Q$-factor. In the experimental demonstration, commercially available SAW filters are selected for the delay line, due to their superior performance at the targeted frequency range ($\sim 1$ GHz). The large group delay ($>15$ ns) of the SAW filters results in at least 20 times smaller modulation frequency and more than 850 times reduction in power consumption compared to other implementations working at similar central frequency [24].

The proposed circuit architecture is shown in Fig. 2(a). Two SAW filters are periodically modulated through RF switches, controlled using square waves $M_1$ and $M_2$ with a 50% duty cycle and a phase of $\varphi = 0^\circ$ and $90^\circ$, respectively. The modulation period $T_m$ of the control signals is set to be four times the group delay of the SAW filters, i.e., $T_m = 4T_2$, where $T_2$ is the group delay, so that the time delay between $M_1$ and $M_2$ is equal to the group delay. Each port is connected to a pair of complementary switches $SW_i$, where $i$ is the port number, forming a differential configuration. The operation of the circuit can be explained through the time diagram shown in Fig. 2(b) and (c). When port 1 is excited as shown in Fig. 2(b), $SW_{11}$ ($\varphi = 0^\circ$) is turned on from $t = 0$ to $t = (T_m/2)$, allowing the incident signal to be transmitted to the upper SAW filter during this time window. As the signal compared to all previous demonstrations of time-varying circulators. As a result, we present a time-varying circulator with ultra-low modulation frequency and power consumption, showing, at the same time, low IL, strong IX, broad BW, and high linearity, thus addressing the major challenges that currently prevent the realization of magnet-free integrated full-duplex components with high performance.

II. DESIGN

A. Time Domain Analysis

The proposed circuit architecture is shown in Fig. 2(a). Two SAW filters are periodically modulated through RF switches, controlled using square waves $M_1$ and $M_2$ with a 50% duty cycle and a phase of $\varphi = 0^\circ$ and $90^\circ$, respectively. The modulation period $T_m$ of the control signals is set to be four times the group delay of the SAW filters, i.e., $T_m = 4T_2$, where $T_2$ is the group delay, so that the time delay between $M_1$ and $M_2$ is equal to the group delay. Each port is connected to a pair of complementary switches $SW_{11}$ and $SW_{12}$, where $i$ is the port number, forming a differential configuration. The operation of the circuit can be explained through the time diagram shown in Fig. 2(b) and (c). When port 1 is excited as shown in Fig. 2(b), $SW_{11}$ ($\varphi = 0^\circ$) is turned on from $t = 0$ to $t = (T_m/2)$, allowing the incident signal to be transmitted to the upper SAW filter during this time window. As the signal
passes through the SAW filter, it exhibits a delay of $T_g$, then it reaches SW$_{21}$ ($\varphi = 90^\circ$), which has also a time delay of $(T_m/4) = T_g$ with respect to SW$_{11}$. Therefore, the signal is transmitted through SW$_{21}$ as well, and it is delivered to port 2. Similarly, from $t = (T_m/2)$ to $t = T_m$, the signal is transmitted to port 2 through the lower branch. Assuming that the filters have infinite BW and dispersion-free group delay, all the power excited from port 1 is delivered to port 2. On the other hand, if the signal is excited from port 2, it can be shown that all the incident power is transmitted to port 3, as depicted in Fig. 2(c). The same time diagram can also be drawn for the signal transmission from port 3 to port 4. Therefore, the input signal will circulate in the following order 1 $\rightarrow$ 2 $\rightarrow$ 3 $\rightarrow$ 4 $\rightarrow$ 1, which is the operation of a circulator.

### B. Frequency Domain Analysis

The time diagram in Fig. 2 is based on the assumption that the SAW filters have infinite BW, which is physically unrealistic. After the input signal is mixed with the clock of the switch SW$_{11}$ or SW$_{12}$ ($i = 1, 2, 3$ or $4$), infinite numbers of IMPs will be generated at points B and B’, depicted in Fig. 3. Due to the finite BW of the SAW filters, only a finite number of IMPs will be transmitted. As a simple approximation, we again assume a simple resonator governed by a second-order differential equation [Fig. 1 (right inset)]. Therefore, by definition

$$Q = \frac{f_0}{BW_{3\text{dB}}}.$$  \hspace{1cm} (2)

Substituting (2) into (1), the relationship between the group delay and BW of a microacoustic resonator can be derived as

$$BW_{3\text{dB}} \approx \frac{1}{\pi T_g}.$$  \hspace{1cm} (3)

Multiple microacoustic resonators can be coupled to form a microacoustic filter (e.g., the ladder-type filter in Fig. 1) to improve the BW. As a rule of thumb, the relationship between the BW and the group delay of a microacoustic filter can be expressed as

$$BW_{3\text{dB}} \approx \frac{1}{t_g}.$$  \hspace{1cm} (4)

Since the modulation frequency is $(1/4t_g)$, only the fundamental tone and the first IMPs fall inside the filter BW. However, as will be shown next, most of the signal power is, in fact, carried out by the fundamental tone and the first-order IMPs; therefore, the filtering process described above will cause only a small IL to the system.

Fig. 3 depicts the generation of IMPs from the input to the output port, i.e., $i = 1, 2, 3, 4$ and $j = 2, 3, 4$ or 1, respectively. As shown in Fig. 3, the incident voltage at the input has a frequency $\omega_{RF}$ and its spectrum can be written as

$$V_A(\omega) = \delta(\omega - \omega_{RF})$$ \hspace{1cm} (5)

where $\delta$ denotes the Dirac delta function. Note that for the frequency domain analysis in this article, only the incident voltage is considered and derived, while the reflected voltage (e.g., the reflected higher order IMPs from the delay line) is not considered for simplicity. After mixing with the switching signal of SW$_{11}$ ($M_1(t)$), the incident voltage at point B becomes the convolution of $V_{in}$ and a square wave with 50% of duty cycle

$$V_B(\omega, \varphi) = V_{in}(\omega) \times \tilde{M}_1(\omega, \varphi)$$ \hspace{1cm} (6)

where $\tilde{M}_1(\omega, \varphi)$ is the Fourier transformation of the square-wave clock signal $M_1(t)$ and $\varphi$ is the phase of the modulation signal. For a 50% duty cycle square wave, $\tilde{M}_1(\omega, \varphi)$ can be written as follows:

$$\tilde{M}_1(\omega, \varphi) = \tilde{M}_1(\omega, \varphi = 0) \times e^{-i\varphi} \left(\frac{n\pi}{\omega_m}\right)$$ \hspace{1cm} (7)

where

$$\tilde{M}_1(\omega, \varphi = 0) = \frac{1}{2} \delta(\omega) + \sum_{n=1}^{\infty} \frac{i}{n\pi} \left| \sin \left(\frac{n\pi}{2}\right) \right| \times [\delta(\omega + n \times \omega_m) - (\delta(\omega - n \times \omega_m))]$$  \hspace{1cm} (8)

where $\omega_m$ is the radius modulation frequency. Therefore, point B will see an infinite number of IMPs. Assuming a transfer function $H_{\text{line}}(\omega)$ for the delay line, the incident voltage at point C can be written as

$$V_C(\omega, \varphi) = V_B(\omega, \varphi) \times H_{\text{line}}(\omega).$$  \hspace{1cm} (9)

As mentioned above, the BW of the SAW filters can only transmit the fundamental tone and the first IMPs. Assuming $H_{\text{delay}}(\omega) = e^{j\delta}$ inside the BW and, by definition, $(d\delta/d\omega) = -T_d$, at the frequency $\omega_{RF} \pm \omega_m$, the phase $\delta$ of the transfer function is $\mp90^\circ$ with respect to the phase at the input frequency (remember that the modulation frequency is set to be $(1/4t_g)$). Therefore, (9) can be rewritten as

$$V_C(\omega, \varphi) = V_B(\omega, \varphi) \times \begin{cases} e^{j90^\circ}, & \omega = \omega_{RF} - \omega_m \\ 1, & \omega = \omega_{RF} \\ e^{-j90^\circ}, & \omega = \omega_{RF} + \omega_m \\ 0, & \text{elsewhere} \end{cases}$$  \hspace{1cm} (10)
These three frequency components are then mixed with the switching signal of the second switch, which exhibits a phase of $\varphi + \pi/2$, and the voltage at point D can be derived

$$V_D(\omega, \varphi) = V_C(\omega, \varphi) \times \hat{M}_{f1}(\omega, \varphi + \pi/2)$$

$$= \sum_{n=0}^{\infty} b_n \delta(\omega - \omega_{RF} - n \times \omega_m) \times e^{-i\omega_{RF} T_d}$$  \hspace{1cm} (11)

where

$$b_n = \begin{cases} \frac{1}{4} + \frac{2}{\pi^2}, & n = 0 \\ -\frac{1}{\pi} e^{-\text{in}_{\omega_{RF}}}, & n = \pm 1 \\ (-1)^n \cdot \frac{1}{\pi} \left( \frac{1}{n+1} - \frac{1}{n-1} \right) \times e^{-\text{in}_{\omega_{RF}}}, & n \text{ is even}, \ n \neq 0 \\ -\frac{1}{2n\pi} \sin \left( \frac{n\pi}{2} \right) e^{-\text{in}_{\omega_{RF}}}, & n \text{ is odd}, \ n \neq \pm 1. \end{cases}$$

Similarly, it can be shown that the voltage $V_D$ has the same form as $V_D$ by replacing $\varphi$ with $\varphi'$. From (11), we observe that all the IMPs have a phase term of $e^{-i\omega m_{\text{out}}}$. Therefore, by using a differential configuration with two complementary switches, i.e., $\varphi' - \varphi = 180^\circ$, the output spectrum will have zero odd-order IMPs, since all the odd-order IMPs will have different signs in the output from the two differential branches, i.e., $e^{-i\omega m_{\text{out}}} = -e^{-i(\varphi + \pi)}$ when $n$ is odd, and they destructively interfere with each other. Meanwhile, even-order IMPs will have the same sign from the two branches, i.e., $e^{-i\omega m_{\text{out}}} = e^{-i(\varphi + \pi)}$ when $n$ is even, and they constructively interfere with each other, showing up at the output port. Furthermore, the fundamental tone at the output can be calculated from $V_{\text{out}}(\omega = \omega_{RF}) = V_D(n = 0) + V_D(n = 0) \approx 0.905 \delta(\omega - \omega_{RF}) \times e^{-i\omega_{RF} T_d}$. Hence, the theoretical IL, assuming zero IL and constant group delay for the SAW filter, is $-20 \times \log(\vert V_{\text{out}} \vert / \vert V_{\text{in}} \vert) = 0.86$ dB. Therefore, even though higher order IMPs are filtered out by the SAW filter after the first switch, the filtering process will only cause an IL of 0.86 dB since most of the power is carried by the fundamental tone and the first IMPs.

The output spectrum of Fig. 3 is simulated using ADS harmonic balance (Fig. 5) with an input frequency of 895 MHz and an input power of 0 dBm. The S-parameters of the SAW filter Qorvo 856671 (Fig. 4) are used in obtaining these results. Even though the group delay of the SAW filter has frequency dispersion for a single-tone input, the group delay dispersion does not affect the performance of the circuit, as long as the phase of the SAW filter transfer function $(H_{d\text{line}}(\omega))$ at the first IMPs is equal to $\mp 90^\circ$ with respect to the input frequency, in which case (10) will be exactly the same as in the ideal case (i.e., no group delay dispersion). Therefore, a modulation frequency of 14 MHz is chosen to achieve this condition. The simulated results are compared to the theoretical ones calculated using (12), showing good agreement. The larger loss of the simulated fundamental tone, i.e., 1.9 dB versus 0.86 dB, is due to the nonzero IL of the SAW filter. The larger simulated values of the higher order IMPs ($n > 4$) can be explained by the fact that the SAW filter has a finite out-of-band rejection so higher order IMPs at points B and B’ can leak into the output of the filter (C and C’).

### C. Group Delay Dispersion

The above spectrum analysis is based on the use of a single-tone input signal. Therefore, the modulation frequency is chosen such that $\mp 90^\circ$ (with respect to the phase at the input frequency) phase are inserted into the first IMPs from the delay line to satisfy (10). However, this relationship does not hold for other frequencies, due to the group delay dispersion of the SAW filter (Fig. 4), causing additional IL. For a given input frequency, the phase at the left and right IMPs is $\varphi_1$ and $\varphi_2$, respectively. As a result, the transfer function at (10) should be modified based on $\varphi_1$ and $\varphi_2$ (i.e., replacing $\mp 90^\circ$ by $\varphi_1$ and $\varphi_2$). Substituting the modified (10) into (11), the new fundamental voltage amplitude $b_0$ (12) can be calculated as

$$b_0' = \frac{1}{4} + \frac{1}{\pi} \cos \left( \frac{\pi}{2} - \varphi_1 \right) + \frac{1}{\pi} \cos \left( \frac{\pi}{2} - \varphi_2 \right).$$  \hspace{1cm} (13)

Based on (13), the IL versus input frequency of the circulator based on the use of Qorvo 856671 SAW filter can be plotted [Fig. 6(a)]. Even though the group delay has
noticeable dispersion, the IL is nearly constant over a frequency range of approximately 890–905 MHz and starts to degrade outside this frequency range.

D. Timing Error

Timing error caused by nonideal modulation frequency is also worth discussing. For simplicity, constant group delay without any dispersion is again assumed. Assuming a nonperfect modulation frequency, i.e., \( f_m = \alpha \times \left( \frac{1}{4} T_d \right) \), where \( \alpha \) is the mismatch coefficient, the phase of the transfer function \( H_{d,\text{line}}(\omega) \) at \( \omega_{RF} \pm \omega_m \) is no longer \( \pm 90^\circ \) with respect to the input frequency. Instead, it can be derived that coefficients of \( e^{\pm i\omega \delta t} \) are introduced to the frequency components at \( \omega_{RF} \pm \omega_m \). Therefore, (10) can be rewritten as

\[
V_C(\omega, \varphi) = V_B(\omega, \varphi) \times \begin{cases} 
eq 0, & \omega = \omega_{RF} \pm \omega_m \\
1, & \omega = \omega_{RF} \\
e^{-i\omega \delta t}, & \omega = \omega_{RF} + \omega_m \\
e^{-i\omega \delta t}, & \omega = \omega_{RF} - \omega_m \\
0, & \text{elsewhere} \end{cases}
\]

(14)

where \( \delta t = T_d - \left( \frac{1}{4} f_m \cdot \alpha \right) \). Therefore, the new fundamental voltage amplitude at D can be calculated by substituting (14) into (11)

\[
b''_0 = \frac{1}{4} + \frac{2}{\pi^2} \times \cos \left[ \frac{\pi}{2} (a - 1) \right].
\]

(15)

Fig. 6(b) plots the additional IL caused by the timing error when the modulation frequency is mismatched from 14 MHz. The additional IL caused by timing error is minimal (e.g., only 0.2 dB when the modulation frequency is off by 3 MHz), showing reasonably large tolerance for modulation frequency mismatch.

E. S-Parameter Simulations

The S-parameters of the circuit in Fig. 2(a) are simulated using ADS harmonic balance. The S-parameters of the SAW filter from Fig. 4 are used in the simulation. The switches are assumed to be resistive at both ON and OFF states (\( R_{ON} = 5 \, \Omega \) and \( R_{OFF} = 1 \, M\Omega \)), with zero rise and fall transition times. The simulated results are shown in Fig. 7, depicting a strong nonreciprocity with an IL of 2.1 dB. The simulated BW, defined as the 20-dB IX of \( |S_{31}| \), is 30 MHz. The slightly higher IL compared to the result of the output spectrum in Fig. 5 is due to the small power leakage from port 1 to ports 3 and 4 resulting from the finite dispersion of the in-band group delay.

F. Quad Configuration

As explained above, a differential configuration with a modulation phase \( \varphi \) of 0° and 180° for the constituent single-ended circuits will cancel the odd-order IMPs yet maintain the even-order tones. In order to further suppress the latter, a new quad configuration is proposed, as shown in Fig. 8. At each port, four switches with a modulation phase difference of 90° are used. Assuming that the modulation phase of the four switches...
Fig. 9. Simulated S-parameters of the quad configuration. The switches used in simulation are based on resistive switching model with ON- and OFF-resistance of 5 and 1 MΩ, respectively.

Fig. 10. Simulated output spectrum of quad configuration.

is 0°, 90°, 180°, and 270°, then the phase term $e^{-in\varphi}$ when $n$ equals 2 will be +1 for $\varphi = 0°$ or 180°, and −1 for $\varphi = 90°$ or $\varphi = 270°$. Therefore, the second-order IMPs from the four branches will destructively interfere with each other, and the output spectrum will show zero second-order IMPs. The quad architecture is also simulated using ADS harmonic balance. The same S-parameters of the SAW filter and the resistive switch model are used in the simulation. The characteristic impedance is tuned to 25 Ω, since at each port, the number of SAW filters is doubled. The simulated S-parameters show a strong nonreciprocity with an IL of 2.1 dB (Fig. 9). The output spectrum is also simulated with an input power of 0 dBm and an input frequency of 895 MHz (Fig. 10). As expected, all the first three IMPs are suppressed, and the closest IMPs to the fundamental tone are the fourth-order ones.

III. MEASUREMENT

A. Implementation

In order to test the proposed circuit architecture, a printed circuit board (PCB) prototype was designed and implemented.

Fig. 11 shows a picture of the fabricated board. Commercially available SAW filters Qorvo 856671 were selected for the experimental demonstration, due to their lowest IL and broad BW at the targeted center frequency of around 1 GHz. The SAW filters and the RF switches were wire-bonded to the PCB. The RF switches are GaAs DPDT engineering samples from Qorvo (IL $\sim$ 0.3 dB; isolation $\sim$ 26 dB; rise/fall time = 0.5–1 ns; P1dB $\sim$ 27 dBm). Since no inductors or TLs are required in the circuit, the device area (SAW filters and switches) is only 6 × 6 mm².

The testing setup is shown in Fig. 12. Two dual-channel function generators were synchronized together to provide the control signals $M_1(t)$, $\overline{M}_1(t)$, $M_2(t)$, and $\overline{M}_2(t)$. The S-parameters were measured using a four-port vector network analyzer (VNA).

B. S-Parameters

The measured and simulated S-parameters are shown in Figs. 13 and 14. The dashed results represent for the
Fig. 13. Measured and simulated IL and IX of the differential configurations. The simulated results are based on the measured parasitic of the switches.

Fig. 14. Measured and simulated RL of the differential configurations. The simulated results are based on the measured parasitic of the switches.

simulated results with the measured parasitic from the RF switches. By assigning ports 1–3 to transmitter (Tx), antenna (ANT), and receiver (Rx), respectively, the IL of Tx-to-ANT and ANT-to-Rx and the IX of Tx-to-Rx, which are the most important metrics for the operation of full-duplex radio, are shown in Fig. 13. The measured IL is 2.9 dB, while the simulated one is 2.7 dB. The 20-dB IX-BW is 23 MHz, i.e., ~2.6% of the center frequency. The RL (Fig. 14) is larger than 15 dB over the entire BW and can be further improved by adding a matching network or improving the RL of the filters. The slight difference between simulated and measured results is attributed to the parasitic introduced by the PCB.

In order to further suppress the IMPs close to the BW, the quad configuration with schematically shown in Fig. 8 was experimentally implemented by combining two differential circuits, as shown in Fig. 15. SMA tees were used to combine the pairs of ports on two PCBs. The S-parameters of the quad configuration were then measured, using the same testing setup shown in Fig. 12, with a terminal impedance of 25 Ω (postprocessing). The measurement shows a similar result of S-parameters, depicted in Figs. 16 and 17, as the differential configuration. The measured IL is 3.3 dB, which is slightly higher than the differential case. The IX is also slightly worse than the differential results. These are due to the parasitic coming from the PCB-level connection of the two boards, such as the extra IL from the SMA cables and connectors and the slight mismatch between the two boards. The measured RL
Fig. 18. Measured and simulated output spectrum for the differential configuration with an input signal of 0 dBm at 895 MHz.

Fig. 19. Measured and simulated output spectrum for the quad configuration with an input signal of 0 dBm at 895 MHz.

is more than 10 dB for both the Tx and the ANT ports. It is worth mentioning that even though the quad configuration is matched to 25 $\Omega$, a 50-$\Omega$ matched quad architecture can be designed by customizing the characteristic impedance of the SAW filters.

C. Output Spectrum

The output spectrum of both the differential and quad configurations is measured by sending an input signal of 895 MHz with an amplitude of 0 dBm, plotted in Figs. 18 and 19, respectively. As expected, the output spectrum of the differential configuration shows the suppression of only odd IMPs, while the quad configuration shows large suppression of all the first three IMPs, i.e., the suppression is more than 32 dBc. The finite IMP suppression and the slight mismatch between the measurement and simulation are attributed to the nonperfect modulation phase provided by the function generator, the slight mismatch between the two PCBs and the finite switching speed of the RF switches. In real implementations, the worst output spectrum typically happens at the edge of the BW, instead of the center. Therefore, in order to evaluate the IM distortion within the entire 20-dB IX nonreciprocal band, the output spectrum with the input frequency close to both of the two band edges (885 and 905 MHz) and the center (895 MHz) is measured and plotted in the same figure (Fig. 20). In all of these three spectrums, the IMP suppression is more than 32 dBc. Therefore, a pseudolinear output spectrum within the entire 20-dB IX nonreciprocal band is demonstrated, with a large IMP suppression.

D. Linearity

The linearity of the differential configuration was tested by measuring the 1-dB input compression point (P1dB) and the input-referred third-order intercept point (IIP3). Due to the 50-$\Omega$ termination of the linearity measurement setup, the linearity of the quad configuration cannot be measured. The input signal for P1dB measurement is a single tone signal centered at 895 MHz. Fig. 21 shows that a P1dB of 29.5 dBm was achieved. The P1dB of the circuit is limited by the RF switches, since SAW filters have higher P1dB (P1dB of a single switch is 27 dBm, while that of a single SAW filter is 28 dBm). Since the signal power is divided into two paths, a $\sim$3 dB more power handling is expected for the circulator circuit compared to a single switch, consistent with the measured results. The IIP3 measurement was done by sending a two-tone signal with frequency spacing of 1 MHz (895 and 896 MHz). The measured IIP3 is 41 dBm (Fig. 22) and is limited by the measurement setup (mostly by the IIP3 of the power combiner to combine the two-tone signal) and can be higher by employing a better setup. The measured linearity is among the highest for all the magnet-free circulators, thanks to the use of large RF switches with high power handling, enabled by the low modulation frequency associated with the employed SAW filter technology.

E. Power Consumption

The power consumption of time-varying circulators originates from the dissipation of energy from the modulation
TABLE I
SUMMARY OF MAGNET-FREE CIRCULATOR PERFORMANCES

| Technology | Center freq. | Mod. Freq. | BW | IX | IL | P1dB | IIP3 | Power Consum. |
|------------|--------------|------------|-----|----|----|------|------|--------------|
| [15] TL    | 25 GHz       | 33%        | 18.4% | 18.3 dB | 3.3/3.2 dB | 21.5/21 dBm | N/A | N/A          |
| [26] TL    | DC-3GHz*     | 83%*       | 93.3%* | 20 dB | 4.3 dB | N/A | N/A | N/A          |
| [24] LC    | 950 MHz      | 33%        | 17%  | 25 dB | 2.1/2.9 dB | 21/31 dBm | 37/50 dBm | 170 mW      |
| [12] LC    | 1000 MHz     | 19%        | 2.4%  | 20 dB | 3.3 dB | 29 dBm | 34 dBm | N/A          |
| [11] LC    | 1000 MHz     | 10%        | 2.3%  | 20 dB | 0.8 dB | 29 dBm | 32 dBm | N/A          |
| [18] MEMS  | 155 MHz      | 0.6%       | 5.8%  | 20 dB | 6.6 dB | N/A | 30 dBm | N/A          |
| [19] MEMS  | 2500 MHz     | 0.1%       | 0.02% | 20 dB | 11 dB | N/A | N/A | N/A          |
| [14] MEMS  | 146 MHz      | 0.1%       | 0.2%  | 15 dB | 8 dB | -8 dBm | N/A | N/A          |
| [27] MEMS  | 1165 MHz     | 0.1%       | 0.3%  | 15 dB | 12 dB | N/A | N/A | N/A          |

This work SAW 900 MHz 1.4% 2.6% 20 dB 2.9 dB 29.5 dBm 41 dBm 0.2 mW

* Defined by the ratio with center frequency.
*b Defined by the IX value.
*c Results are broadband measured from DC to 3 GHz.
*d Assuming center frequency is 1.5 GHz.
*e Baluns are de-embed.

Fig. 21. Measured P1dB of the differential configuration.

Fig. 22. Measured IIP3 of the differential configuration.

signal through the charging and discharging current of the MOSFET, since there is no exchange of power between modulation and signal network [42]. Thanks to the use of acoustic filters, the modulation frequency is lower than the one required in an equivalent LC network. For this reason, its power consumption from driving the switch gates is inherently lower. The power consumption of the device is calculated by measuring the rms voltage and current at the modulation ports. An ultralow power consumption of 202.2 μW is measured in our prototype, showing 850 times reduction compared to previous demonstrations based on LC [25].

IV. CONCLUSION

In this article, we demonstrated a magnet-free circulator based on the spatiotemporal modulation of SAW filters. Thanks to the high Q-factor of SAW filters, the circulator shows low IL, large IX, broad BW, and ultralow modulation frequency all at the same time. Table I summarizes the presented results in comparison to other works on magnet-free circulators. Working at a center frequency of 900 MHz with one of the lowest IL of only 2.9 dB, this article achieves a broad BW of 2.6% with a low modulation frequency of only 1.4%. Compared to [11], [12], [15], [25], and [26] based on the spatiotemporal modulation of either TLs or LCs, the modulation frequency and power consumption are significantly reduced. For example, compared to [25] based on LC delay networks with similar center frequency, the modulation frequency is more than 20 times smaller, which translates to a reduction of power consumption by a factor of 850. Therefore, compared to demonstrations based on TLs/LCs, this article overcomes the problems of large modulation frequency and large power consumption while does not show tradeoffs in other performances such as IL and BW like [14], [18], [19], and [27]. Furthermore, thanks to the low modulation frequency, RF switches with high power handling are used; therefore, the demonstrated circulator shows one of the highest linearity (P1dB of 29.5 dBm and IIP3 of 41 dBm). The demonstrated high-performance magnet-free circulator is an important progress toward full-duplex communication systems in the near future.

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