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3-D Metal Printed Compact High-$Q$ Folded Waveguide Filter With Folded Antenna

Jiayu Rao, Student Member, IEEE, Kenneth Nai, Povilas Vaitukaitis, Student Member, IEEE, Yuepei Li, Student Member, IEEE, and Jiasheng Hong, Fellow, IEEE

Abstract—A novel compact fan-shaped folded waveguide resonator (FSFWGR) is proposed in this paper. The resonator can reduce the size by 87.5% compared with the waveguide cavity while retaining a high unloaded quality ($Q_u$) value (> 1500). Owing to these advantages, a high-quality ($\sim Q$) fourth-order bandpass filter centered at 6.0 GHz with a 6.67% bandwidth is designed firstly. Novel slots cross-coupling is presented to produce a pair of transmission zeros (TZs) while avoiding support structures that traditional metal probes often used. Then compact deformed folded antenna is designed to integrate with the filter. Compared with published works where the last resonator of the reference filter often needs to be detuned, all parameters here do not need to change by introducing a metal probe between the feeding point of the filter and antenna. Therefore, the integrating process is more straightforward. The integrated filter/antenna achieves the same filtering function as the reference filter and a high gain of 6.7 dBi. For experimental demonstration, metal 3-D printing technology was used to fabricate models with only two layers that significantly reduced the assembly loss. The final measured results agreed well with simulated ones.

Index Terms— 3-D metal printing, deformed folded antenna, fan-shaped folded waveguide resonator, filtering antenna, high-$Q$ value, slots cross-coupling.

I. INTRODUCTION

The components miniaturization are urgently in demand for radio frequency (RF) front end with the evolution of the modern wireless communication systems. Antennas and filters are indispensable in any communication system but occupy a significant portion of the entire system footprint. For reducing the occupied spaces, the concept of the filtering antenna that has both radiating and filtering functions simultaneously was proposed and studied in the past few years [1]-[18]. Due to removed the traditional matching circuits between the antennas and filters, these works showed advantages at reducing the profiles and improving transmission efficiency. Among those cases, the prototype filters with a high-quality ($\sim Q$) factor are more attractive than low-$Q$ planar filter/antennas as they could present lower insertion loss and better frequency selectivity.

Cavity resonators, including substrate integrated waveguide (SIW) and metal waveguide resonators are often used to design high-$Q$ prototype filters to integrate with different antennas [5]-[18]. In [10] [11], SIW cavity filters with unloaded $Q$ ($Q_u$) value ~1000 integrated with slot antennas are proposed and tested. Benefiting from the high-$Q$, the efficiency of the filtering antenna was more than 95% while maintaining a good selectivity. Furthermore, Xun et al. proposed a synthesis approach to integrating high-$Q$ filter/antennas seamlessly at different frequencies and bandwidths in [12] and then applied this method to design various filter/antennas successfully in [13]-[18]. Among them, the 3-D vertical stacked SIW technique was adopted to reduce the transverse footprints of the filtering antenna to design high-$Q$ slot filtering antennas [17] [18]. As a result, near-zero transition loss between the antenna and filter was achieved with reduced transversal sizes. Although these works obtained desired results, it should be mentioned that they all were based on one wavelength-sized resonator which means the footprint of the prototype filter is still significant. Besides, the selectivity performance is not well due to transmission zeros (TZs) are not introduced in the filter parts. Meantime, a higher $Q_u$ than 1200 is hard to achieve because of the limitation of SIW [19].

It is well known that metal waveguide resonators can obtain the highest $Q_u$, and greater robustness but at the price of bulky volume. For obtaining compact waveguide cavity resonators while maintaining the high-$Q$ value, the folded-waveguide resonator (FWGR) was proposed in [20] and applied to design filters successfully in [21]-[23]. These works obtained similar high-$Q$ properties to conventional waveguide resonators while reducing the size to quarter wavelength. However, no work reported high-order quasi-elliptic bandpass filter design based on FWGR, which would also present some challenges for traditional machining because of the inside thin metal plate structure. In addition, the extra loss brought during fabrication and assembly process would need to be taken into account [22] [23]. An alternative manufacturing method, direct 3-D metal printing has demonstrated its superiority for those components with complex internal structure in our previous work [24], which will be exploited here to fabricate the models.

This paper is an extended version of our work presented at APMC 2020 [25]. The additional materials are the necessary details of the analysis, synthesis, and experimental demonstration of the filter and filter/antenna. Firstly, the proposed fan-shaped folded waveguide resonator (FSFWGR)
can achieve a more compact size while maintaining high-$Q$ compared to published folded resonators [21]-[23]. Then a four-pole bandpass filter with a pair of TZs is designed based on the resonator. Compared with the traditional method of using a metal probe to obtain cross-coupling to generate TZs, the self-supporting slots cross-coupling is adopted for integrated printing. Subsequently, a curved folded antenna is integrated with the filter to reduce the footprints further. Folded antennas have been proved can reduce the size to 0.06λ effectively by adding multiple layers folding but at the price of the narrower bandwidth ($\leq 3\%$) [26]-[28]. Herein, the integrated filter/antenna can obtain the same bandwidth of 6.67% as the filter while retaining a compact profile. The transmission efficiency is as high as 97% compared with 87% for a standalone folded antenna reported in [26]. Finally, the filter and filter/antenna are fabricated using 3-D metal printing technology for experimental validation. The measured results agreed well with simulated ones.

This article is organized as follows. Section II presents a comprehensive analysis of the resonator and its comparison with non-standard circular waveguide resonators. Section III gives the synthesis process of the 4-pole filter, which is used as a reference for the filter/antenna system. The filter/antenna synthesis procedure is described in Section IV. The equivalent circuit models are developed to synthesize the filter/antenna, which is validated by full-wave simulations. Section V details the practical implementation of 3-D metal manufacturing including the printing direction and sensitivity analysis. The measured results are also given and discussed in Section V. Section VI summarizes the main results of the work.

II. Resonant Analysis

The physical configuration of the proposed fan-shaped folded waveguide resonator (FSFWGR) is shown in Fig. 1. It consists of the inside metal plate (yellow part) and the outer metal box (blue part). The dimensions of the inside metal plate are determined by the thickness ($t$), length ($L_0$), and arc angle ($\theta$). And the $L_0$ and $H_0$ are the length and height of the surrounding fan-shaped box, respectively. The angle of the surrounding box is fixed as 45°. The two slots denoted by $g_1$ and $g_2$ are used to adjust the resonant frequency and $Q_u$ easily. The two slots functioned as mimic magnetic walls that will allow electromagnetic fields to continue from one half to the other of the folded waveguide resonator. Fig. 2 illustrates the electrical field distributions of the fundamental mode in FSFWGR and a TM$_{011}$ mode in a circular waveguide resonator using full-wave electromagnetic (EM) simulation software. As the resonant mode of FSFWGR resembles a portion of the TM$_{011}$ mode of circular waveguide, it can be expected to have some high-$Q$ property with a significant small size or volume. In order to validate this, Table I shows the comparison results of geometric size along with simulated $Q_u$ of a circular waveguide and FSFWGR at the same fundamental resonant frequency. The ratio of $Q_u$ of both resonators is almost constant at all heights. Although the $Q_u$ of the proposed FSFWGR is less than half of the circular waveguide, the volumes decreased by six times.

![Fig.1. The configuration of the Fan-Shaped Folded-waveguide resonator.](image1)

**Fig.2. Electrical Field distribution. (a) Circular waveguide. (b) FSFWGR.**

<table>
<thead>
<tr>
<th>TABLE I</th>
<th>COMPARISON OF THE CAVITY AND PROPOSED RESONATORS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Height</td>
<td>Fan shape folded waveguide resonator (radius = 22mm, $\theta = 45^\circ$)</td>
</tr>
<tr>
<td>7 mm</td>
<td>$f_0$(GHz) $Q_u$ $f_0$(GHz) $Q_u$</td>
</tr>
<tr>
<td>6.03</td>
<td>1885</td>
</tr>
<tr>
<td>11 mm</td>
<td>6.04</td>
</tr>
<tr>
<td>15 mm</td>
<td>6.04</td>
</tr>
</tbody>
</table>

For further investigation of the $Q_u$ of the FSFWGR, Fig. 3 plots the $Q$-factor and the resonance frequency versus physical parameters. Unless otherwise stated, the metal in this paper is aluminium ($\sigma = 3.56 \times 10^7$ S/m). All physical dimensions are fixed as follows except the one varying in the paper for simulation (unit: mm): $L_0 = 22.0$, $H_0 = 7.0$, $L_1 = 21$, $g_1 = 1.0$, $t = 1.0$. It should be noted that adjusting $L_1$ and $g_1$ is equivalent to adjusting the parameters of $g_2$ and $\theta$ but easier to control. From Fig. 3, it can be seen that $L_1$ ($g_2$) and $g_1$ ($\theta$) are two key parameters for controlling $Q_u$ and frequency. The shorter $L_1$ or larger $g_1$, the higher $Q_u$ and frequency. Hence, a compromise is made between the frequency and $Q_u$. The parameter $t$ has little effect on these two properties, therefore, that is fixed as 1.0 mm considering the tolerance and robustness of the fabricating.

III. Fourth-Order Prototype Filter Design

Following the analysis in Section II, the FSFWGR can be exploited to design compact high-$Q$ bandpass filters (BPFs). This section describes the design procedure in detail, with the filter centred at 6.0 GHz and a 400 MHz bandwidth. In addition, the $Q_u$ and narrowband rejection ($f_0 \pm 400$ MHz) are asked for larger than 1500 and -30 dB, respectively. With the considerat-
-ion of the above specifications, the design starts with the choice of the resonators. Referring to Fig. 3, the dimensions of the resonator are chosen as (units: mm) $L_0 = 22$, $L_1 = 20.5$, $H_0 = 7.0$, $t = 1.0$, $g_1 = 1.0$ with $Q_e$ more than 1700. Subsequently, the two critical steps are extracting internal and external coupling coefficients that will be given in the next subsection.

A. Internal and External Coupling

The coaxial cables with 50-ohm port impedance are utilized to obtain the required external coupling ($Q_e$) values as shown in Fig. 4. Moving the coaxial probe in the horizontal direction ($dport_H$) or vertical direction ($dport_V$), different $Q_e$ values can be extracted using the equation:

$$Q_e = \frac{0.5 \tau_{max}}{t}$$  \hspace{1cm} (1)

where $\omega_0$ is the resonant frequency while $\tau_{max}$ is the maximum value of group delay of $S_{11}$ at resonance. Fig. 5 illustrates the various simulated $Q_e$ values concerning the direction distance that effective control to the $Q_e$ can be observed in the vertical direction. Also, it should be noted that the probe feeding will shift the resonant frequency a little higher.

In order to achieve the compact sizes of the high-order filter, two types of internal coupling are adopted here: horizontal intersect coupling and vertical inductive window coupling, as shown in Fig. 6 (a) and Fig. 6 (b), respectively. Both couplings are inductive that the inductive window coupling behaved as the conventional magnetic window coupling. The intersect coupling allows the current flow through the slots between two adjacent resonators, as shown in Fig. 6 (a). Furthermore, the entire footprint of the filter can be reduced further due to the intersection. Fig. 7 gives the coupling coefficients ($k$) between the adjacent cavities as a function of the intersect depth ($dport_{H/V}$) and aperture sizes ($L_1, C, W_C$). The curves of $k$ have been obtained by considering the frequencies $f_{even}$ and $f_{odd}$ of a 2-pole filter with weak feeding and applying the equation [29]:

$$k = \frac{f_{even}^2 - f_{odd}^2}{f_{even}^2 + f_{odd}^2}$$  \hspace{1cm} (2)

where $f_{even}$ and $f_{odd}$ are even mode and odd modes, respectively. Fig. 7 shows the intersect coupling can obtain an extensive range of $k$ as the conventional inductive window coupling.

B. Transmission Zeros

For meeting the requirements of the narrowband rejection, TZs were introduced. Cross-coupling between the first and last resonator can achieve symmetrical TZs at each side of the passband often used in the filter design. The traditional physical realization is shown in Fig. 8 (a), where a metal probe is used to produce the required negative or electrical coupling. However, the metal probe cannot be printed together with the remaining parts because it is overhanging. The slots negative coupling is shown in Fig. 8 (b). To validate this, Fig. 9 compares
Fig. 6. Front and side view of inner coupling: (a) intersect coupling. (b) inductive window coupling.

Fig. 7. Extracted $k$ versus the physical parameters.

EM results of the slots coupling ($g_3$) and inductive window coupling ($W_C$) of a 2-pole filter with weak feed coupling. As shown, assuming the unchanged mode is even mode, the sign of $k$ of inductive coupling would be positive while the sign of $k$ of slots coupling would be negative according to (2). This verified that slots coupling can function as the negative coupling, and TZs should be expected.

C. Fourth-Order Filter Realization

For demonstration, a four-order bandpass filter prototype with the listed specifications is designed based on the above analysis and design curves. A pair of TZs located at 5.65 GHz and 6.37 GHz was introduced for improving the selectivity of the stopband. The layout of the filter is shown in Fig. 10 (a), where the green metal cylinders are frequency tuning screws and the blue ones are bandwidth tuning screws. Additionally, the purple metal screws can tune frequency and bandwidth simultaneously. It should be noted that the purple bandwidth tuning screws have a more extensive tuning range than the blue ones. The topology is shown in Fig. 10 (b), where $S/L$ denotes the input and output port, and each node represents a resonator.

Fig. 8. Side and front view of the negative coupling between resonators 1-4. (a) metal probe negative coupling. (b) slots negative coupling.

Fig. 9. Simulated $S_{21}$ of a two-order FSFWGR filter with weak feeding.

The solid line represents the main path coupling based on the magnetic coupling and the dashed line means cross electrical coupling (slots coupling). This creates a negative coupling that produces TZs at both sides of the passband. According to [29], the design values of the inner and external coupling coefficients can be given

$$Q_{51} = Q_{44} = 15.3066$$

$$k_{g_3} = k_{W} = 0.0523, k_{g_3} = 0.0523, k_{W} = -0.0144$$

Referring to the above analysis curves and combining the optimization of CST, the final sizes of the filter can be extracted as shown in Fig. 10. The final simulated results of the filter and the metal probe coupling designed filter are plotted together for comparison in Fig. 11. As shown, the TZ that appeared at 5.66 GHz is very close to the conventional method, while another one is not as obvious. This might be attributed to the fact that the electrical coupling produced by the slots is not strong as the metal probe. Nonetheless, the filter still meets the requirements...
the ground while the part area of the left sidewall (denoted by dc) can be regarded as the shorting wall. The curved surface of the patch radiation element is defined by (5)

\[ k_s \times (W_a^2 + h_s^2) \]  

where \( W_a \), \( h_s \), and \( k_s \) are length, height, and curve degree factors of the curved patch. The \( k_s \) can adjust the radiation beam pattern due to the inclined ground. In addition, a long metal probe inside the cavity is used as the antenna feeding (feed-end). The rotation angle (\( \theta_k \)) and radius of the metal probe (\( R_k \)) can control the feeding position and quality factor of the antenna (\( Q_k \)). Another end of the metal probe is connected to the feeding point of the prototype filter (connect-end). It should be noted that to avoid affecting the performance of the filter, a gap (>1.0 mm) between the metal probe and metal plate is necessary. With the above arrangements, the integrated filter/antenna can be designed which will be validated in the next subsection.

### B. Equivalent Circuit Model and Physical Realization

For synthesizing the filtering antenna and extracting the circuit components, the equivalent circuit is shown in Fig. 13. In Fig. 13 (a), the \( \theta \) represents a matching transmission line (TL) between the curved patch surface and ground denoted by \( \text{dc} \). In addition, the \( X_c \) express the resonant reactance, and \( R_s \) corresponds to the antenna radiation resistance. Assuming the TL is lossless, the TL and antenna can be incorporated together as a new antenna load, and with the equivalent circuit shown in Fig. 13 (b). Noted that the same circuit part as Fig. 13 (a) was removed to highlight the change. The left part of Fig. 13 (b) can also be transformed into the TL equivalent circuit as shown in the right part. The input impedance is fond by

\[ Z_m = jX_p + Z_i \]  

where \( X_p \) is the metal probe reactance that is given by

\[ X_p = \frac{\sigma p h_s}{2\pi} \left[ \ln \left( \frac{2}{\beta p} \right) - 0.57721 \right] \]  

with \( \beta = 2\pi / \lambda_0 \). \( Z_i \) or 1/\( Y_i \) is obtained from the TL equivalent circuit, which is

\[ Y_i = \frac{Y_0}{\tan(\beta h_a) + jY_0} + Y_0 \frac{Y_0 \tan(\beta h_a - h_s)}{\tan(\beta h_a) + jY_0} \]  

where \( Y_0 \) is the characteristic impedance and \( Y_i \) is the reactance associated with power near the radiating edge. According to (5) in [26], \( Y_0 \) is strongly dependent on the ratio of \( W_a / \text{dc} \). Herein, it is not necessary to calculate the accurate value of \( Z_m \), but to know that the reactance of \( Z_m \) might be capacitive or Specifically, the sign of the reactance of the new antenna load would be defined as follows.

IV. FILTER /ANTENNA SYNTHESIS

Although the last order of a reference filter could be used as the radiator and resonator simultaneously to achieve the filter / antenna, it results in a low equivalent \( Q \) for the filter and a small gain for the antenna. In order to maintain the integrity of the filter, an extra curved folded antenna is introduced in this design. Since the footprint of both the filter and antenna are very compact, and space is utilized effectively, the entire footprint of the filter/antenna is minimal.

#### A. Configuration of The Filter /Antenna

The basic configuration of the proposed filtering antenna is shown in Fig. 12, consists of an above reference filter, an inclined metal probe and a curved patch radiation element that is located at the top of the metal box. Compared with traditional folded antennas [20], the right inclined sidewall can be seen as
also depends on the TL \((W_a, dc)\) with the condition of \(X_p\) is always larger than zero \((R_0 < 2.0 \text{ mm})\).

To confirm this, Fig. 14 plots the \(LC\) values of the standalone curved folded antenna with different \(dc\). All dimensions are fixed as following except the one that is varying (units: mm): \(W_a = 22, h_a = 10.1, dc = 6.4, R_p = 1.27, \theta_a = 45^\circ\). As shown, with \(dc\) increasing from 5.5 mm to 9 mm, the \(L\) and \(C\) can be switched between negative and positive. Also, the sign of \(L\) and \(C\) is always opposite, so when \(L\) is positive, the equivalent of the new antenna load can be modelled as series circuits. And when \(L\) is negative, the equivalent of the new antenna load can be modelled as parallel circuits. Hence, with the existence of the TL, the equivalent circuit of the new antenna load can be modelled as series or parallel \(RLC\) circuits, which means any type of antenna can be integrated regardless of the circuit of the antenna itself. Assume the new antenna load is the series \(RLC\) circuit, then the \(L_a, C_a\) of the last resonator can be incorporated into the reactance part of another new antenna load as shown in Fig. 13 (c) with \(L_a'' = L_a + L_a', C_a'' = C_a + C_a'.\) The circuit in the right part of Fig. 13 (c) can then be used to synthesize the filter/antenna. The antenna, TL and the fourth resonator become the last stage of the filter/antenna and work as a radiator.

For designing a filter/antenna, the quality factor \(Q_a\) and centre frequency \(f_0\) of the bandpass filter are two critical factors for the synthesis. According to Fig. 13 (c), the \(f_0\) is

\[
f_0 = \frac{1}{2\pi \sqrt{L_a''C_a''}}
\]

which may be slightly different from the antenna centre frequency \(f_a = 1 / (2\pi \sqrt{L_aC_a})\) due to the existence of the TL. The \(f_a\) can be determined by \(h_a\) and \(d_c\) \((h_a + d_c \approx \lambda / 4)\) that \(h_a\) is smaller than normally folded antenna due to the curvature. Then to achieve the same \(S_{11}\) responses as the reference filter, the \(Q_a\) must have the same value as the \(Q_{cn}\) of the filter, which is defined as:

\[
Q_a = \frac{2\pi f_a L_a''}{R_a} = \frac{2\pi f_a L_a}{R_a} + \frac{2\pi f_a L_a'}{R_a} = \frac{2\pi f_a L_a}{R_0}
\]

where \(R_0\) is the input impedance of the reference filter.

Firstly, assume \(R_a = R_0\), then \(L_a' = 0\) refer to (10) which means a perfect matching between filter and antenna. If \(R_a \neq R_0\), the desired \(Q_a\) can be obtained only by adjusting \(L_a\) which depends on \(TL\) and antenna parameters such as \(dc, h_a, w_a, \theta_a, k_a\). At this stage, one thing needs to be noted that the \(f_a\) does not need to be exactly fixed to \(f_0\) and a rough frequency is enough due to the \(Q_a\) varies relatively slowly with frequency. For example, the parameter \(dc\) can change \(f_a\) quickly but affects \(Q_a\) little as shown in Fig. 15. After meeting the requirements of \(Q_a\), the \(f_a\) can be easily adjusted using other parameters \((h_a, w_a)\) when connected to the filter. Fig. 16 gives the EM simulated results of the \(S_{11}\) response as the function of \(dc\), which showed that adjusting the \(dc\) has a negligible impact on \(f_0\) (although \(f_a\) did change) but can control \(Q_a\) effectively. This also proved that again the \(f_a\) could be different from \(f_0\) when extracting the \(Q_a\).
To this end, the complete design process is summarized as follows. First, select the initial length of the curved antenna according to the length of $h_a$ and $dc$, then extract $Q_a$ based on (8) and Fig. 15. Although many factors can control the $Q_a$, only three fundamental parameters are needed to be tuned ($dc$, $h_a$, and $\theta_a$) which is a straightforward job with the optimization function of CST. Subsequently, reflector and director radiation elements are added to improve the gain as shown in Fig. 17. It should be noted a metal cylinder support (green colour) is added compared with Fig. 12 to make sure the inclined metal probe can be printed successfully. The added support will not affect the in-band responses with selected place and radius. The details will be given in part A of Section V. Finally, in addition to the above three parameters, another three ($h_{a1}$, $h_{a2}$, $dc_1$) are also aimed to be optimized with the goal of $S_{11}$ below -20 dB. The dynamic range of each parameter was set to be ± 20% away from the initial values. The final dimensions are soon obtained and given in Fig. 17. The simulated results of the reference filter and filter/antenna are given in Fig. 18. As it is shown, the filter/antenna system exhibits the same filtering function as the reference filter which proved the effectiveness of the above synthesis process.

In general, the selectivity of the filter/antenna is excellent due to the introduction of the TZs. The centre frequency is 6.0 GHz with a bandwidth of 6.67% desired. The overall efficiency across the passband is around 97% which is calculated using gain/directivity. The simulated average gain and directivity using CST are found to be 6.92 and 6.71 dBi, respectively.

V. 3-D PRINTED PROTOTYPES

A. Printing Discussion

For experimental demonstration, all designed models were fabricated by Renishaw PLC using an AM500Q printer with a layer thickness of 30 μm. The metal powder, ALSi10Mg-0403 ($\sigma = 2.56 \times 10^7$ S/m), is composed of aluminium alloyed with 10% silicon, along with other minor elements. The powder allows printing the models that maintain high conductivity while obtaining other better features such as lighter weight and stronger robustness than pure aluminium [30] [31].

Fig. 18. EM and measured responses of the filter and filter/antenna.

Fig. 19 illustrates the status check before printing using the software called “Renishaw QuantAM”, where the colour (blue and red) cylinder supports are generated by the software automatically [30]. The blue means the supports in this area might be needed, while the red means supports are necessary. As it is shown, referring to the printing direction, supports are necessary only in those screw holes which are easily removed after printing. The $\theta_a$ is called safe build angle that one part could be printed successfully with $\theta_a$ is no more than 55°. Following this rule, even with complex inside structures and the inclined metal probe, only one support might be needed to print filter-part as shown in Fig. 19 (a). Here, the support is might be needed because the angle of the incline ($\theta_a$) is 45° refer in Fig. 17. It should be noted that the place or the radius of the support generated by the software is random during the printing process that could affect the in-band responses. For in case, we put cylinder support manually before printing to ensure printing the inclined probe will be successful as shown in Fig. 17. The cylinder support with a selected place and a diameter of 1.0 mm does not affect the in-band responses. Hence, the added support does not need to be removed after manufacturing which simplified the post-processing. In addition, for the antenna piece (Fig. 19 (b)), it can be seen that the curved degree of the antenna elements ($\theta_a'$) is much less than 55°, which means it also can be printed well in the printing direction.

In conclusion, the models designed in this paper achieved self-supported structures that could be printed from bottom to top directly without any inclination angle or supports.

B. Sensitivity Analysis

For high-$Q$ cavity filters, tolerance analysis is often required before fabrication. For the metal 3-D printing process, the manufacturing error often comes from two aspects. One is the tolerance of the printer which is ± 0.1mm [30], and another is the top surface roughness which is around 15μm [30]. It should be noted here that the resolution could be much smaller than this that depends on the model’s direction and geometry. The models in this paper would have smaller tolerances because of their self-supported structures as stated above. However, to examine the worst-case scenario, we would set the threshold value as 0.15mm which is quite enough for sensitive analysis.
The -0.1 mm tolerance analysis is ignored because this error usually does not appear for the AM500Q printer according to our experiences.

Referring to Fig. 10 and Fig. 19, there are five critical parameters for the filter to analyze in the printing direction which are \( g_1 \), \( g_2 \), \( g_3 \), \( L_C \) and \( W_C \). In addition, in the direction that is normal to the printing direction (denoted by purple arrow), only parameter \( t \) needs to be considered. Fig. 20 shows the simulated curves for the sensitivity analysis. As it is shown, only \( g_1 \) shifts the centre frequency a little while, others are less sensitive to the manufacturing tolerance even when the worst situation is considered. Furthermore, the RF properties of the low frequency (compared with millimetre band) and wideband ( \( > 5\% \) ) can also explain the reduced sensitivity of the filter. In summary, considering the low sensitivity and without postwork after manufacturing the filter, we did not add tuning screws.

C. Experiment Results

The fabricated prototypes of the filter and filter antenna are shown in Fig. 21 which finished well without supports. After measurement using the digital ruler, the average fabrication error was only around 50 \( \mu m \). The measured comparison results of \( S \)-parameters are shown in Fig. 18. As shown, the measured results of the filter and filter/antennas are agreed well with simulated ones. The same filtering function as the filter of the filter/antenna can be observed. The TZs highly improved the selectivity of the stopband of both components. The measured center frequency is 5.99 GHz which is very close compared with simulated 6.0 GHz. The insertion loss of the filter is only 0.45 dB due to the high-\( Q \) of this type of resonator. The measured \( Q \) value is around 1640. In addition, the gain of the filtering antenna was measured and plotted versus frequency as shown in Fig. 20. The measured bandwidth (\( S_{11} < -15 \) dB) of the filtering antenna is 6.43%. The measured average gain across the passband is around 5.9 dB with around 85.8% efficiency. The measured radiation patterns of both E- and H- planes of the filter/antenna are shown in Fig. 21. This demonstrates that the integrated folded antenna can function with a 7.35% FBW (\( S_{11} < -10\)dB), which is larger than the bandwidth ( \( \approx 4.5\% \) ) of a probe–fed standalone patch antenna with similar dimensions.

VI. CONCLUSION

A compact high-\( Q \) FSFWGR has been presented and analyzed in this paper. A fourth-order BPF with a pair of TZs has been designed then as the reference filter of the high-\( Q \) filter/antenna based on the resonator. The design process has been detailed including obtaining inner coupling, external coupling and TZs. For keeping the integrity of the filter, an extra curved folded antenna was utilized due to its compact size. The synthesis process for extracting the antenna centre \( f_a \)
and $Q_a$ is illustrated and employed to integrate the filter and antenna. As a result, the integrated filter/antenna achieved the same filtering selectivity as the filter. For experimental validation, 3-D metal printing technology was used to print the models, which showed promising results for further R&D exploiting advanced design and manufacturing using 3-D metal printing techniques for 5G/6G applications.

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