

ORIGINAL ARTICLE

Co-Design of Class-F⁻¹ Microwave Amplifier and Bandpass Filtering Matching Network With Harmonics Control Functionality

Phanam Pech¹  | Yongchae Jeong^{2,3}

¹JIAIT-IT Human Resource Development Center, Jeonbuk National University, Jeonju, Republic of Korea | ²IT Convergence Research Center, Jeonbuk National University, Jeonju, Republic of Korea | ³Division of Electronics and Information Engineering, Jeonbuk National University, Jeonju, Republic of Korea

Correspondence: Yongchae Jeong (ycjeong@jbnu.ac.kr)

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ABSTRACT

This paper demonstrates the co-design of class-F⁻¹ microwave amplifier (MA) and bandpass filtering (BPF) matching network (MN) with harmonics control functionality (HCF). The proposed BPF MN can match arbitrary termination impedances (ATIs) to 50 Ω at fundamental frequency (f_0) and simultaneously control the second harmonic ($2f_0$) and third harmonic ($3f_0$) impedances to be open- and short-circuited, respectively. The BPF MN with HCF is realized with the alternative J/K inverters cascaded via quarter wavelength ($\lambda/4$) transmission line resonators and the additional shunt-series LC resonators, which produces the resonances at harmonic frequencies. In the realization process, the circuit is implemented using microstrip lines and applied as the output MN (OMN) of the proposed class-F⁻¹ MA. To validate the proposed design method, three class-F⁻¹ MAs with different OMNs are designed with f_0 of 6 GHz using CGHV1F006S GaN HEMT. With the same input power (P_{in}) level, the output power (P_{out}) and drain efficiency of the proposed class-F⁻¹ MA with BPF OMN are better than those of the conventional class-F⁻¹ MA (CMA) cascaded with conventional BPF (CBPF) about 0.42 dB and 5.1%, respectively. Moreover, the circuit size is reduced by about 35.68%.

1 | Introduction

The increasing popularity of wireless communication systems underscores the importance of high-efficiency microwave amplifiers (MAs). Among these, the class-F and class-F⁻¹ MAs are significant components. In the class-F MA design, short-circuited termination at even-order harmonics is required for current peaking, while open-circuited termination at odd-order harmonics is needed for voltage peaking. The harmonics control configuration has become representative of high-efficiency MAs. There has been a growing interest in the class-F⁻¹ MA due to its superior performances. In contrast to the class-F MA, the class-F⁻¹ MA exhibits an open circuit at even-order harmonics,

resulting in voltage peaking, and a short circuit at odd-order harmonics, resulting in current peaking. It is widely recognized that this configuration allows class-F⁻¹ MA to achieve higher efficiency than class-F operation. In [1–4], the quantitative and clear explanation of better efficiency and the design guide for optimum output power or efficiency of class-F⁻¹ MA in comparison to the class-F MA have been discussed.

The briefly described difference between class-F and class-F⁻¹ can be seen with ideal current and voltage waveforms in Figure 1. For class-F⁻¹ MA, the half-period rectangular of current and sine wave voltage must not be overlapped ideally while these current and voltage waveforms can be expressed by

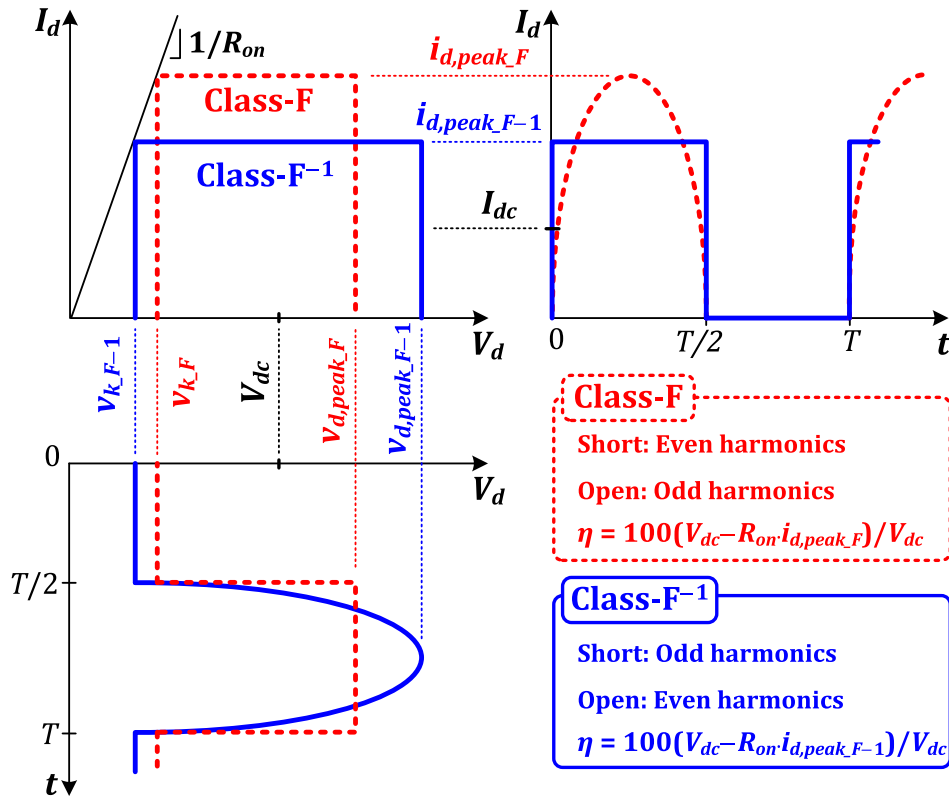


FIGURE 1 | Ideal current and voltage waveforms of class-F and class-F⁻¹ MAs.

Fourier series as (1) and (2). The current and voltage waveforms consist of odd- and even-order harmonics in addition to fundamental and DC components, respectively. The dissipation of harmonic power may be eliminated if the even-order harmonics are terminated with open circuit and the odd-order harmonics are terminated with short circuit.

$$I_d = i_{d,peak_F-1} \left(\frac{1}{2} + \frac{2}{\pi} \sin(\omega_0 t) + \frac{2}{\pi} \sum_{n=3,5,7,\dots} \frac{1}{n} \sin(n\omega_0 t) \right) \quad (1)$$

$$V_d = v_{d,peak_F-1} \left(\frac{1}{\pi} - \frac{1}{2} \sin(\omega_0 t) + \frac{2}{\pi} \sum_{n=2,4,6,\dots} \frac{1}{n^2 - 1} \cos(n\omega_0 t) \right) \quad (2)$$

Ideally, the class-F⁻¹ MA can be provided with an efficiency of approximately close to 100%. It can be obtained if there is no overlapping between the current and voltage waveforms, resulting in approximately zero internal dissipation power. In literature, various design approaches of class-F⁻¹ MA have been presented [5–9]. Mostly, only the second and third harmonics are considered for realizing convenience. Once the number of controlled harmonics increases, the complexity of the harmonics control networks (HCNs) is also increased. Based on the measurement results of class-F and class-F⁻¹ MA presented in [1–9], 100% efficiency cannot be obtained, and the efficiency is decreased as the operating fundamental frequency (f_0) increases to a higher band. The efficiency reduction may also relate to the non-ideality of the transistor, conduction angle limitations, mismatch losses, and parasitic losses due to the parasitic components. It is important to note that those class-F⁻¹ MAs are

designed using the typical HCNs and matching networks (MNs), but the filtering characteristics are not considered.

In radio frequency (RF) front-end transmitting systems, MA is typically followed by conventional bandpass filter (CBPF) for harmonics rejection and securing the quality of the desired signal as can be seen in Figure 2. Those devices are designed separately with the port termination impedances (TIs) equal to the system impedance (Z_0) of 50 Ω . Cascading MA with CBPF can reduce the electrical performances of MA because of the connection mismatch that causes signal reflection between them and insertion loss (IL). Moreover, it may require a cable or adaptor to connect MA with CBPF. In real application, the ILs of cable and adaptor cannot be ignored. As a result, the output power of MA will be degraded from P_{out} to P'_{out} . Once the output power is degraded, the other related parameters such as gain and efficiency of transmitter will be also degraded.

To reduce the complexity of cascading method, the co-design of MA with BPF is investigated. The co-design of MAs and cavity resonators of the complex termination impedance (CTI) BPFs are introduced in [10–12] with f_0 of 1.22, 3, and 3.1 GHz, respectively. These CTI BPFs are designed with the coupling matrix and transformed into the physical design with substrate-integrated evanescent-mode cavities. Recently, the co-design of MAs and substrate-integrated waveguide (SIW) BPF MNs are presented in [13–15]. The MAs are designed at X-band frequency while the SIW BPF MNs are analyzed based on the detuning resonant frequency of the SIW resonator to match the reactance components of CTIs. However, controlling the frequency locations of harmonic components is not considered in those designs and focuses only the fundamental matching.

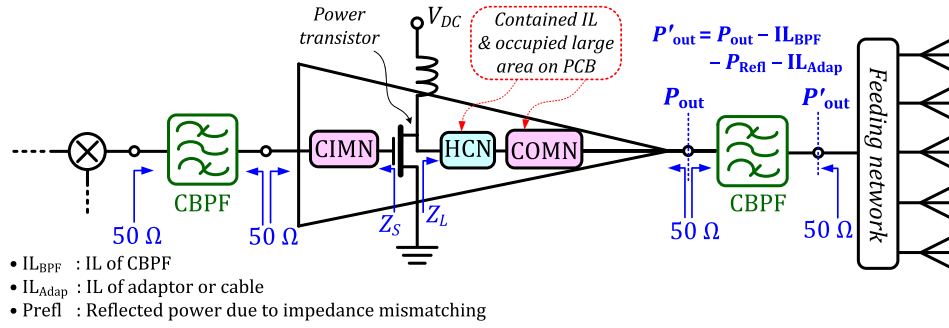


FIGURE 2 | RF front-end transmitter of a typical wireless communication system.

This paper presents the co-design method of class- F^{-1} MA and BPF output matching network (OMN) with harmonic control functionality (HCF). The proposed BPF OMN with HCF is a multifunctional circuit that provides filtering, impedance matching, and harmonics control functions. The BPF OMN with HCF can directly match the optimum load impedance (Z_L) of transistor to 50Ω . It also provides good in-band frequency selectivity and simultaneously controls the impedances at $2f_0$ and $3f_0$ for open- and short-circuited conditions, respectively. To validate the advantages of the proposed method, three class- F^{-1} MAs with different OMNs such as conventional class- F^{-1} MA (CMA), CMA cascaded with CBPF, and the proposed co-design of class- F^{-1} MA with BPF OMN are designed and fabricated for performance comparison. Their block structures can be seen in Figure 3. By using the co-design method, the issues occurred by the cascading of CMA to CBPF can be eliminated. Moreover, the conventional HCN, conventional OMN, and CBPF can be integrated by the proposed BPF OMN with HCF as can be seen in Figure 3c. According to these, the electrical performances and the circuit size of the proposed co-design of class- F^{-1} MA and BPF OMN with HCF are better and smaller than that of CMA cascaded with CBPF.

2 | Analysis of BPF MN With HCF

To perform the co-design of class- F^{-1} MA and BPF, the BPF should be able to operate as the MN and simultaneously control the impedances at $2f_0$ and $3f_0$ as described in the previous section. It means that the BPF can match the arbitrary real or complex impedances to 50Ω . The investigation into the proposed BPF MN with HCF has been done as follows.

2.1 | Characteristics of CBPF With Alternative J/K Inverters and $\lambda/4$ Resonator

Figure 4a shows the equivalent circuit of CBPF by using the J/K inverters cascaded alternatively via the $\lambda/4$ transmission line (TL) resonators, which can be seen in [16–20]. Typically, the CBPF with $\lambda/4$ TL resonator is designed with the TI of $50\text{--}50\Omega$. This CBPF is designed using the same method presented in [20] with f_0 of 1 GHz, order of resonator (n) of 4, fractional bandwidth (FBW) of 5%, and return loss (RL) of 20 dB.

As can be seen in Figure 4b, the frequency location at f_0 is denoted by the blue square symbol while that of even-order harmonics is denoted by a black triangle symbol. Similarly, the frequency

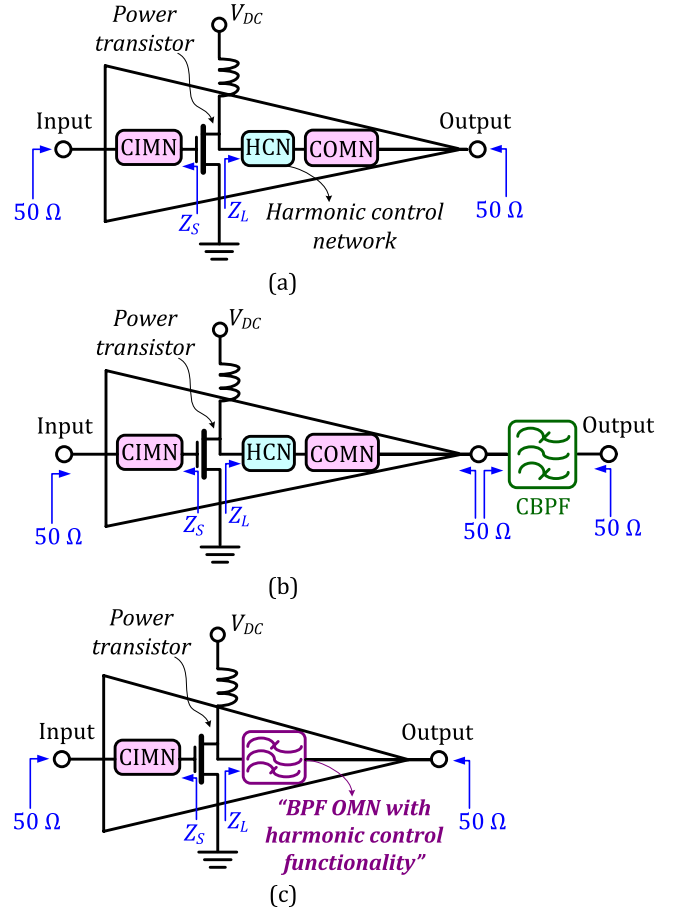


FIGURE 3 | Structures of the designed class- F^{-1} MAs for performance comparison: (a) CMA, (b) CMA cascaded with CBPF, and (c) proposed co-design of MA and BPF OMN with HCF.

location of odd-order harmonics is denoted by a green circle symbol. The frequency locations at f_0 and odd-order harmonics are duplicated to each other at the center point of the Smith chart. A more interesting point is that the locations of the even-order harmonic frequencies are located at the open-circuited point. Figure 4c presents the ideal simulated S -parameters of that CBPF. As can be seen, the passband has occurred not only at the f_0 but also at $3f_0$, $5f_0$, and higher odd-order harmonic frequencies. In contrast, the high attenuation level occurred at $2f_0$, $4f_0$, and higher even-order harmonic frequencies.

Typically, J/K inverters can be realized as $\lambda/4$ TLs, as illustrated in Figure 5. For J -inverter, the value of input admittance ($Y_{in,J}$)

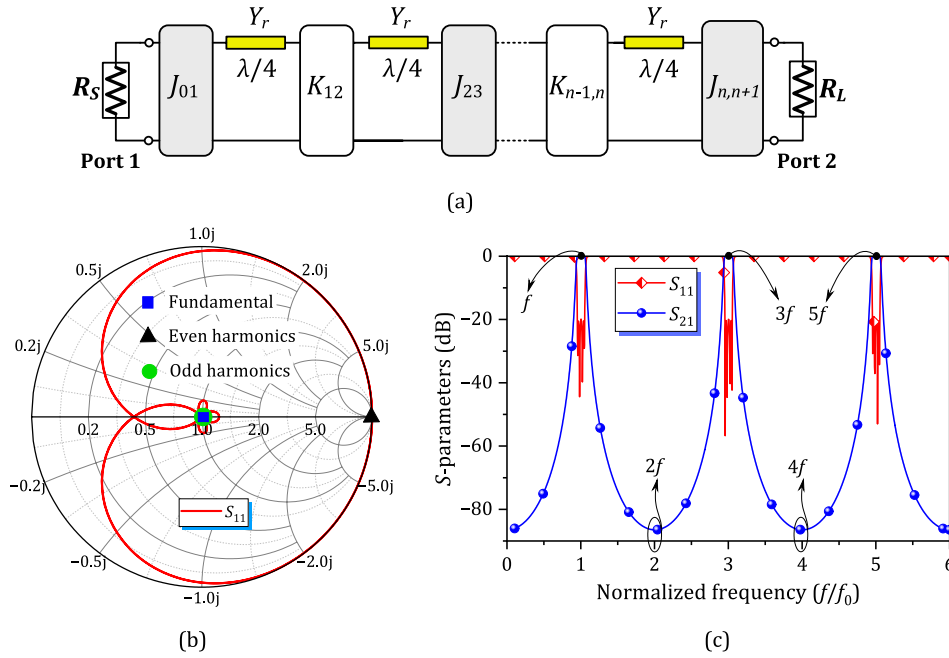


FIGURE 4 | CBPF with $\lambda/4$ resonator and alternative J/K inverters: (a) equivalent circuit, (b) frequency locations of fundamental and harmonic components on Smith chart, and (c) S -parameters.

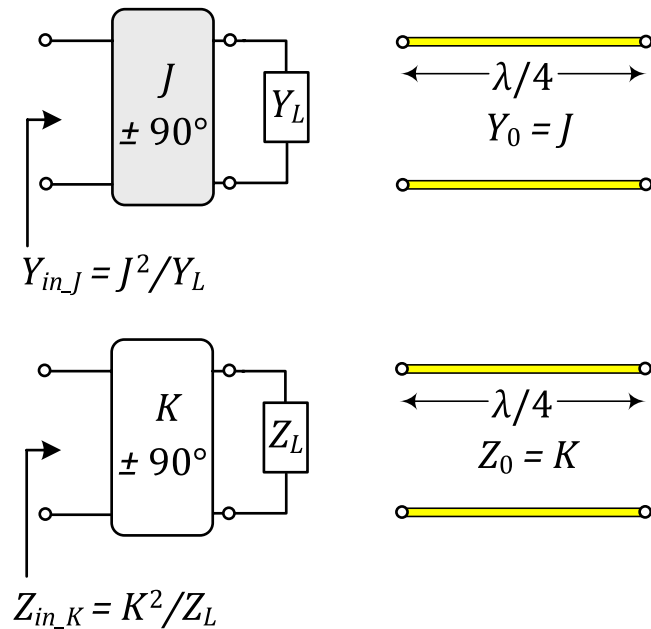


FIGURE 5 | Typical structure of J/K inverters and their equivalent circuits.

approaches zero when load admittance (Y_L) tends toward infinity, and vice versa. Similarly, the value of input impedance (Z_{in_K}) approaches zero when the load impedance (Z_L) tends toward infinity, and vice versa for K -inverter. It seems like there is a possibility for transforming open-circuited to short-circuited or short-circuited to open-circuited using J/K inverters.

2.2 | Proposed BPF MN With HCF

Based on the characteristics of J/K inverters, the BPF MN with HCF is developed, and its equivalent circuits are presented in

Figure 6 according to even/odd order. The TIs at Ports 1 and 2 are $Z_S = R_S \pm jX_S$ and $Z_L = R_L = 50\Omega$, respectively. The proposed BPF MN with HCF consists of J/K inverters cascaded alternatively via the $\lambda/4$ TL resonators. Generally, the $\lambda/4$ TL resonator is essentially a distributed version of an LC resonator, with frequency-dependent inductance and capacitance along its length. A short-circuited $\lambda/4$ TL resonator behaves like a parallel LC resonator at resonance. In contrast, an open-circuited $\lambda/4$ TL resonator behaves like a series LC resonator at resonance. Z_r and Y_r are the characteristic impedance and admittance of the $\lambda/4$ resonators, respectively.

As can be seen in Figure 4b, the frequency locations at f_0 and odd-order harmonics of CBPF realized using $\lambda/4$ TL resonators are duplicated to each other. To make the third harmonic component become short-circuited at Port 1 for the circuit in Figure 6, Point A should be short-circuited at $3f_0$. To fulfill this requirement, the shunt-series $L_{H3}C_{H3,1}$ resonator with the resonance frequency of $f_{H3,1}$ is connected at the output of the $\lambda/4$ TL resonator (Point A).

According to the impedance transforming characteristic of $\lambda/4$ TL resonator and the first J -inverter (J_{01}), open- and short-circuited characteristics at $2f_0$ and $3f_0$ are obtained at Port 1, respectively. Because the proposed BPF MN with HCF will be applied as the OMN of the proposed class- F^{-1} MA, the attenuation characteristics at $2f_0$ and $3f_0$ must be improved as much as possible. Generating transmission zeros (TZs) at those frequencies is a solution for this requirement. Therefore, the other shunt-series $L_{H2}C_{H2}$ and $L_{H3}C_{H3,2}$ resonators with the resonance frequencies of $2f_0$ and $3f_0$, respectively, are connected to the $\lambda/4$ TL resonators.

However, detuning the resonant frequency of the first resonator is required for matching the CTI, and it can be calculated using the following equation [13]:

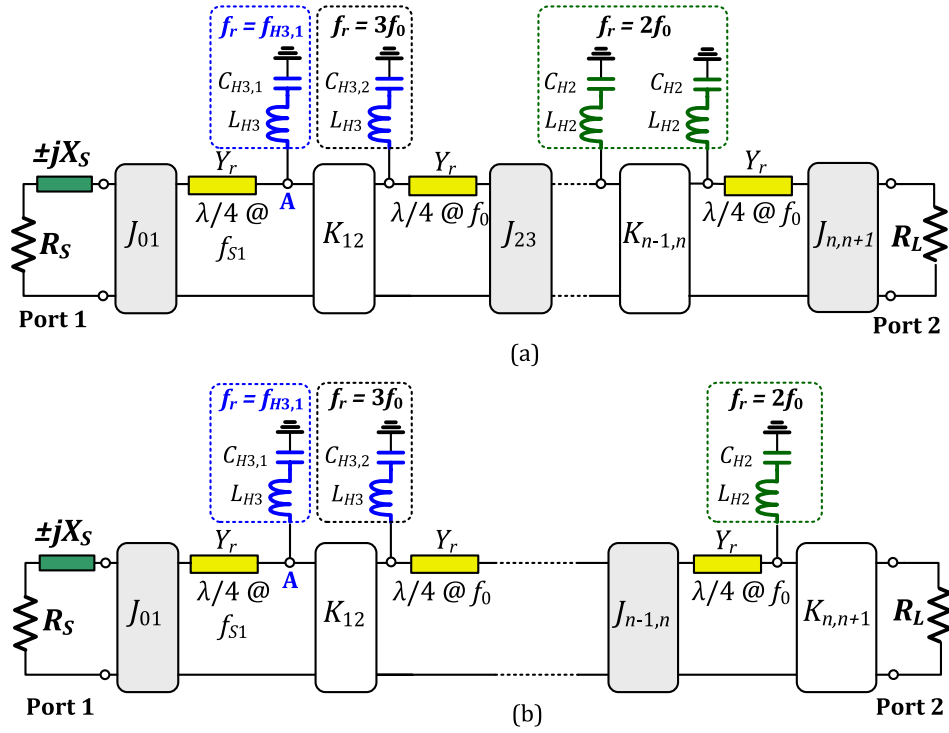


FIGURE 6 | Proposed circuits of BPF MN with HCF: (a) even order and (b) odd order.

$$f_{S1} = f_0 \left[\sqrt{1 + \left(\frac{X_S \text{FBW}}{2R_S g_0 g_1} \right)^2} + \frac{X_S \text{FBW}}{2R_S g_0 g_1} \right] \quad (3)$$

When the resonant frequency of the first $\lambda/4$ TL resonator is detuned from f_0 to f_{S1} , the resonant frequency of the $L_{H3}C_{H3,1}$ resonator is also detuned from $3f_0$ to $f_{H3,1}$, and it can be calculated using Equation (4). In contrast, the location of harmonic frequency at $2f_0$ is not affected by the f_{S1} . By arbitrarily defining the values of L_{H2} and L_{H3} , the values of C_{H2} , $C_{H3,1}$, and $C_{H3,2}$ can be calculated using Equation (5). For arbitrary real-to-real TIs, f_{S1} is equal to f_0 , and $C_{H3,1}$ is equal to $C_{H3,2}$.

$$f_{H3,1} = \begin{cases} 3f_0 - 6(f_{S1} - f_0) & \text{iff } f_{S1} > f_0 \\ 3f_0 + 6(f_0 - f_{S1}) & \text{iff } f_{S1} < f_0 \end{cases} \quad (4)$$

$$C_{H2} = \frac{1}{(2\omega_0)^2 L_{H2}} \quad (5a)$$

$$C_{H3,1} = \frac{1}{(\omega_{H3,1})^2 L_{H3}} \quad (5b)$$

$$C_{H3,2} = \frac{1}{(3\omega_0)^2 L_{H3}} \quad (5c)$$

Once the shunt-series $L_{H3,1}C_{H3,1}$, $L_{H2}C_{H2}$, and $L_{H3,2}C_{H3,2}$ resonators are connected into the $\lambda/4$ TL resonators, the equivalent

slope parameters of them can also be determined. The optimum matching characteristic must be obtained at f_0 , not at the harmonic frequencies. Therefore, the slope parameters are calculated as the function of f_0 . The equivalent susceptances and reactance slope parameters can be calculated as follows:

$$b_1 = \frac{\omega_{S1} Y_r L_{H3}}{Y_r + 8f_{S1} L_{H3}} \quad (6a)$$

$$b_2 = \frac{\omega_0 Y_r L_{H2}}{Y_r + 8f_0 L_{H2}} \quad (6b)$$

$$b_3 = \frac{\omega_0 Y_r L_{H3}}{Y_r + 8f_0 L_{H3}} \quad (6c)$$

$$x_1 = \frac{\pi(Z_r + 8f_{S1} C_{H3,1})}{4} \quad (6d)$$

$$x_2 = \frac{\pi(Z_r + 8f_0 C_{H2})}{4} \quad (6e)$$

$$x_3 = \frac{\pi(Z_r + 8f_0 C_{H3,2})}{4}, \quad (6f)$$

where b_1 is the equivalent susceptance slope parameter of the first $\lambda/4$ TLs and shunt-series $L_{H3}C_{H3,1}$ resonator, b_2 is the equivalent susceptance slope parameter of $\lambda/4$ TLs and shunt-series $L_{H2}C_{H2}$ resonators, and b_3 is the equivalent susceptance slope

parameter of $\lambda/4$ TLs and shunt-series $L_{H3}C_{H3,2}$ resonators. The same approach is applied to the equivalent reactance slope parameters x_1, x_2 , and x_3 .

Once the slope parameters have been determined, the J/K inverters can be defined using Equation (7) for even-order BPF MN with HCF, and the same approach can be applied for odd-order resonator.

$$J_{01} = \sqrt{\frac{\text{FBW}b_1}{R_S g_0 g_1}} \quad (7a)$$

$$K_{12} = \text{FBW} \sqrt{\frac{x_1 x_3}{g_1 g_2}} \quad (7b)$$

$$J_{23} = \text{FBW} \sqrt{\frac{b_2 b_3}{g_2 g_3}} \quad (7c)$$

$$K_{n-1,n} = \text{FBW} x_2 \sqrt{\frac{1}{g_{n-1} g_n}} \quad (7d)$$

$$J_{n,n+1} = \sqrt{\frac{\text{FBW}b_2}{R_L g_n g_{n+1}}} \quad (7e)$$

TABLE 1 | Calculated values for J/K inverters of 20–50 Ω BPF MN with HCF.

J_{01}	K_{12}	J_{23}	K_{34}	J_{45}
0.009073	3.535018	0.001089	3.535130	0.005740

TABLE 2 | Calculated values for J/K inverters of $(31+j23)$ to 50 Ω BPF MN with HCF.

J_{01}	K_{12}	J_{23}	K_{34}	J_{45}
0.005798	3.535018	0.001089	3.535130	0.005740

To prove the validity of the proposed BPF MN with HCF, the investigation on simulation results is performed. The BPF MNs are designed with the TIs of 20–50 Ω and $(31+j23)$ to 50 Ω . The f_0 , FBW, RL, and n are defined as 6 GHz, 10%, 20 dB, and 4, respectively. For the BPF MN with the TI of 20–50 Ω , $f_{S1}=f_0=6$ GHz is calculated by using Equation (3). $L_{H2}=0.53$ nH and $L_{H3}=0.29$ nH are selected. Using Equations (4) and (5), $C_{H2}=0.331895$ pF and $C_{H3,1}=C_{H3,2}=0.269585$ pF are calculated. By defining $Z_r=50$ Ω and $Y_r=1/Z_r$, all slope parameters can be calculated using Equation (6) while J/K inverters can be determined using Equation (7). The values of calculated J/K inverters are listed in Table 1.

Similarly, $f_{S1}=6.239518$ GHz is calculated by using Equation (3) for the BPF MN with the TI of $(31+j23)$ to 50 Ω . All of the design parameters can be calculated using the same approaches mentioned previously. The values of calculated J/K inverters are listed in Table 2. For optimum short-circuited at $3f_0$, a slight modification or tuning on the capacitance of $C_{H3,1}$ is required.

As can be seen in Figure 7a, the BPF MNs provide a passband at f_0 and are simultaneously matched the target TIs. The frequency locations of $2f_0$ and $3f_0$ are located at open- and short-impedance points on the Smith chart, respectively. The simulated S -parameters of the BPF MNs are shown in Figure 7b. The TZs occurred at $2f_0$ and $3f_0$. As a result, the attenuation level at those frequencies can be enhanced.

Figure 8 presents the circuit implementations of J/K inverters and shunt-series $L_{HX}C_{HX}$ resonators. Figure 8a shows the J -inverter and its implementation circuit. J -inverter with TL resonator can be realized with coupled lines, which are widely used in the design of coupled BPFs [16–20]. The even- and odd-mode impedances of coupled lines with arbitrary electrical lengths (θ) can be determined as follows:

$$Z_{0e}|_{i,i+1} = Z_r \frac{1 + J_{i,i+1} Z_r \csc \theta + J_{i,i+1}^2 Z_r^2}{1 - J_{i,i+1}^2 Z_r^2 \cot^2 \theta} \quad (8a)$$

$$Z_{0o}|_{i,i+1} = Z_r \frac{1 - J_{i,i+1} Z_r \csc \theta + J_{i,i+1}^2 Z_r^2}{1 - J_{i,i+1}^2 Z_r^2 \cot^2 \theta} \quad (8b)$$

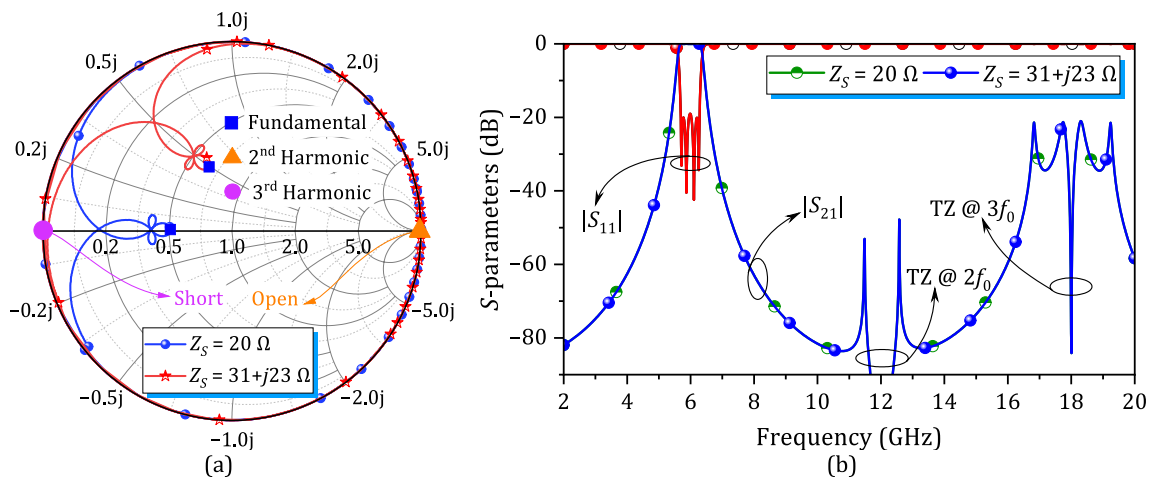


FIGURE 7 | Ideal simulation results of BPF MN with HCF: (a) impedance point on Smith chart and (b) S -parameter.

where i is the even number starting from 0.

Similarly, Figure 8b presents the K -inverter and its T-type TL implementation circuit, which consists of two series TLs with electrical parameters of Z_r and ϕ_{jj+1} and a shunt short-circuited TL stub with electrical parameters of Z_{Kr} and θ_{jj+1} . The design equations can be obtained by equating the $ABCD$ -parameters of the K -inverter and T-type network. The electrical lengths of ϕ_{jj+1} and θ_{jj+1} can be calculated using Equation (9).

$$\phi_{jj+1} = -\tan^{-1}\left(\frac{K_{jj+1}}{Z_r}\right) \quad (9a)$$

$$\theta_{jj+1} = \tan^{-1}\left(\frac{Y_{Kr}Z_r\cos\phi_{jj+1}\sin\phi_{jj+1}}{\sin^2\phi_{jj+1} - \cos^2\phi_{jj+1}}\right), \quad (9b)$$

where j is the odd number starting from 1.

Figure 8c shows the circuit implementations of a shunt-series $L_{HX}C_{HX}$ resonator. This resonator can also be realized using an open-circuited stub TL.

3 | Design of Class-F⁻¹ MAs

3.1 | Design of Conventional Class-F⁻¹ MA (CMA)

Currently, the research interest in the design of RF circuits for the upcoming sixth-generation (6G) communication systems has attracted much attention. However, the specific operating frequency band of 6G has not been officially announced yet. Various research associations and companies focused on communication services have predicted that a few of the most compelling bands for 6G may be located at around 6~7 GHz [21] and 7~15 GHz [22, 23]. In addition, the effective implementation of 6G will also require tapping into the existing spectrum below 7 GHz, and it has been mentioned in [24].

In this work, the f_0 of the designed class-F⁻¹ MAs is selected at 6 GHz. For experimental demonstration and performance comparison, three different class-F⁻¹ MAs are designed with different OMNs. To obtain the optimum source (Z_{ST}) and load (Z_{LT}) impedances of transistor, source- and load-pull simulations are performed using the manufacturer transistor model with a predefined bias condition. In this design, the CGHV1F006S GaN HEMT from Cree is used with bias condition of $V_{GS} = -2.85$ V and $V_{DS} = 40$ V. The load-pull and source-pull simulations are performed on advance design

system (ADS) software. During the load-pull process, the impedances at $2f_0$ and $3f_0$ are set to infinity and zero, respectively. The contours of delivered power to the load (Pdel) and PAE are presented in Figure 9. With the aforementioned conditions, the CGHV1F006S can be provided with the maximum Pdel and PAE of 38.65 dBm and 61.54%, respectively. As can be seen, the optimum points of Pdel and PAE are separated from each other. With the trade-off of impedance points between optimum Pdel and PAE, $Z_{ST} = 18.5 - j13 \Omega$ and $Z_{LT} = 17.68 + j18.73 \Omega$ are extracted with Pdel and PAE of 38.39 dBm and 58.31%, respectively.

Figure 10a shows the conventional HCN for designing CMA. The conventional HCN consisted of TL₁, TL₂, TL₃, and TL₄ with the electrical lengths of $\lambda/8$, $\lambda/4$, $\lambda/24$, and $\lambda/12$ at f_0 , respectively. The characteristic impedances of these TLs can be chosen arbitrarily. A radial stub is connected to TL₂, and this TL₂ is operated as a short-circuited stub for $2f_0$ at junction P₁. The realization of a radial stub based on its resonant frequency can be done by following the same approach presented in [25]. The short impedance at P₁ is transformed to an open circuit through TL₁. As a result, Z_{in} equals to infinity or open-circuited at $2f_0$ can be obtained at the input of the HCN. Similarly, TL₄ provides a short circuit at junction P₂ for $3f_0$. According to the electrical length of TL₁ and TL₃, Z_{in} equals to zero or short-circuited at $3f_0$ can be obtained at the input of HCN. In CMA design, the input MN (IMN) and OMN are realized with TLs based on general matching methods.

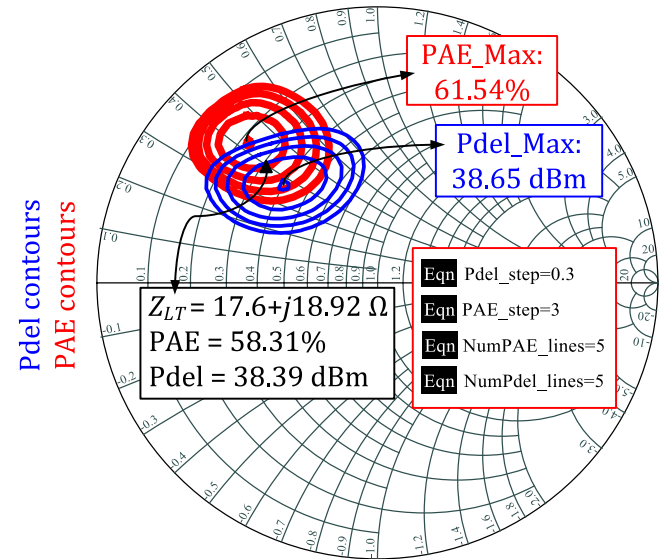


FIGURE 9 | Pdel and PAE contours obtained from load-pull simulation.

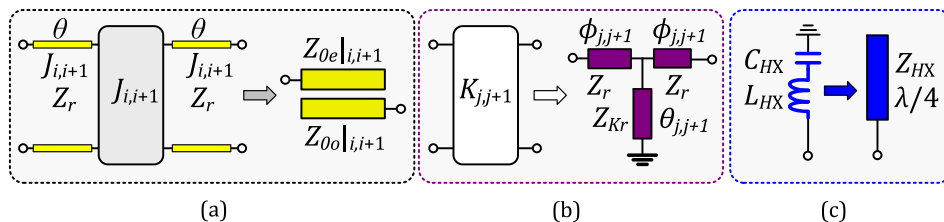


FIGURE 8 | Circuit implementations of (a) J -inverter, (b) K -inverter, and (c) shunt-series $L_{HX}C_{HX}$ resonator.

All circuits presented in this paper are implemented using Taconic-TLY PCB with $\epsilon_r = 2.2$ and $h = 0.508$ mm. In real circuit implementation, the IMN, HCN, and OMN contain some IL. Figure 10b shows the photograph of the fabricated HCN and OMN of CMA. The comparison of impedance points obtained from electromagnetic (EM) simulation and measurement is presented in Figure 10c. The circuit is matched well to the target Z_{LT} at f_0 . However, the locations of the $2f_0$ and $3f_0$ are located slightly inside the Smith chart because of the limited quality factor (Q) of the used PCB. In addition, the comparison of S -parameters is presented in Figure 10d. The IL of this network at f_0 is around 0.35 dB. The TZs occurred at around $2f_0$ and $3f_0$ because of the open- and short-circuited termination, respectively.

3.2 | Design of CMA Cascaded With CBPF

For performance and circuit size comparison, the CMA described in Subsection 3.1 is cascaded with a CBPF. The CBPF is designed with the TI of 50–50 Ω based on the structure presented in Figure 3a. The design specifications are $f_0 = 6$ GHz, FBW = 10%, $n = 4$, and RL = 20 dB. According to these, the values of J/K inverters can be calculated using the same approach presented in [20]. The determined values are listed in Table 3.

The equivalent circuit of CBPF can be seen in Figure 11a. In circuit realization, the J/K inverters are utilized using coupled lines and T-type networks, respectively. The characteristic impedances of the T-type network are selected as $Z_r = Z_{Kr_{12}} = Z_{Kr_{34}} = 50 \Omega$.

By using Equations (8) and (9), all design parameters are determined as seen in Table 4.

Because the values of ϕ_{12} and ϕ_{34} are negative, the length of the coupled lines might be slightly shorter than $\lambda/4$, and unwanted capacitive cross-coupling between them may occur. The photograph of fabricated CBPF is presented in Figure 11b while the comparison of S -parameters between EM simulation and measurement is shown in Figure 11c. The TZs that occurred beside the passband are related to the aforementioned cross-coupling. The maximum IL of 1.49 dB and minimum RL better than 19.5 dB are obtained within the operation band measured from 5.75 to 6.4 GHz. The attenuation levels better than 40 dB are measured from 1 to 5.3 GHz and 8 to 16.35 GHz at the lower and higher stopbands, respectively.

3.3 | Proposed Co-Design of Class-F⁻¹ MA and BPF OMN With HCF

The proposed class-F⁻¹ MA was realized using the conventional IMN and BPF OMN with HCF. The IMN is identical to that of CMA. The proposed BPF OMN can control the location of

TABLE 3 | Calculated values for J/K inverters of CBPF.

J_{01}	K_{12}	J_{23}	K_{34}	J_{45}
0.005742	3.534103	0.001090	3.534004	0.005742

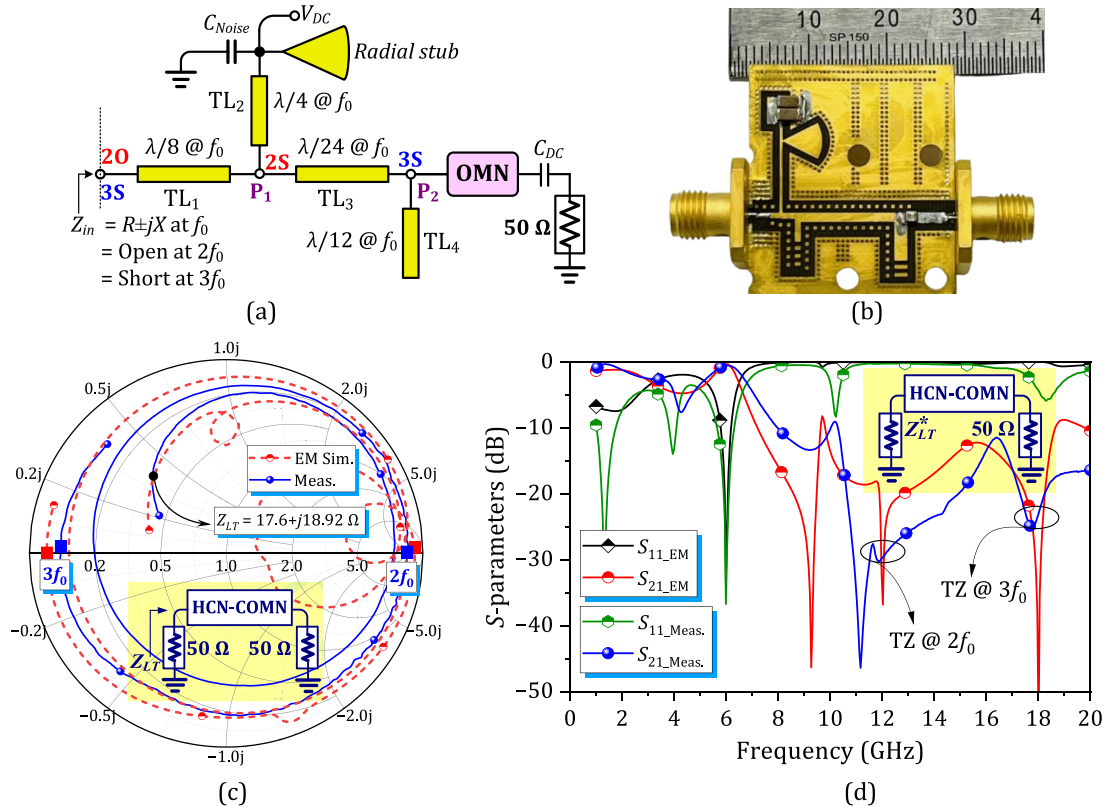


FIGURE 10 | Conventional HCN of CMA and OMN of CMA: (a) equivalent structure, (b) fabricated circuit, (c) impedance points on Smith chart, and (d) S -parameters.

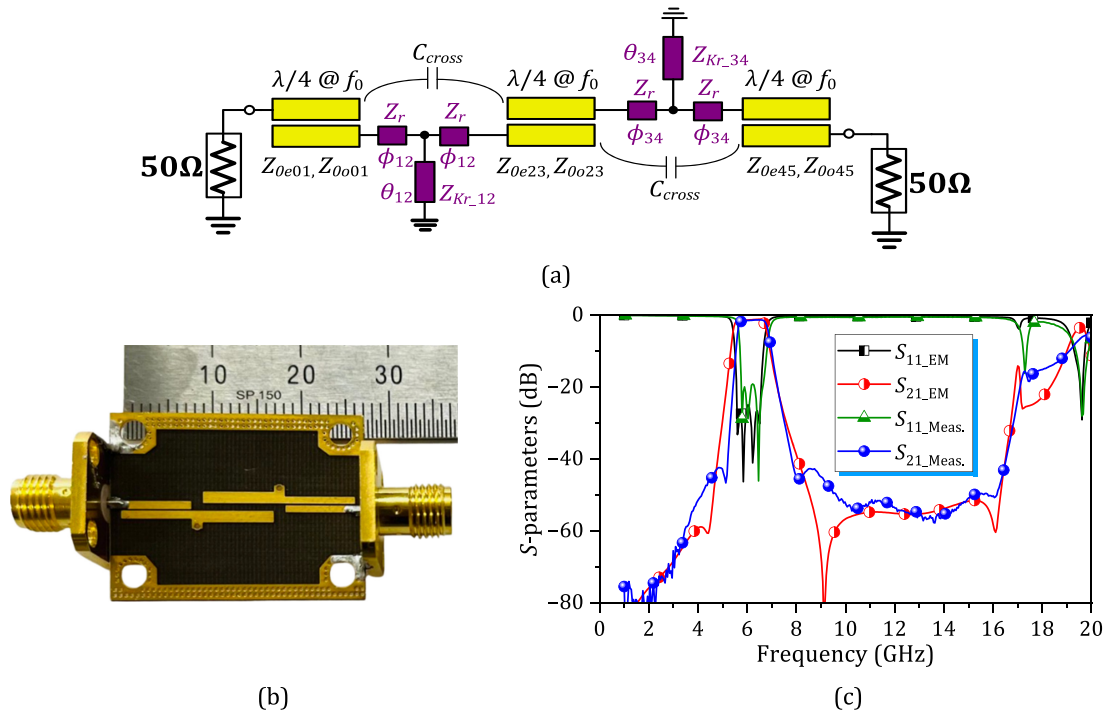


FIGURE 11 | 50-50Ω CBPF: (a) realization circuit, (b) fabricated circuit, and (c) S-parameters.

TABLE 4 | Design parameters of CBPF.

Z_{0e01}/Z_{0o01} (Ω)	Z_{0e23}/Z_{0o23} (Ω)	Z_{0e45}/Z_{0o45} (Ω)	$\phi_{12} = \phi_{34}$ ($^\circ$)	$\theta_{12} = \theta_{34}$ ($^\circ$)
68.8/39.6	52.9/47.4	68.5/39.8	-4.04	4.06

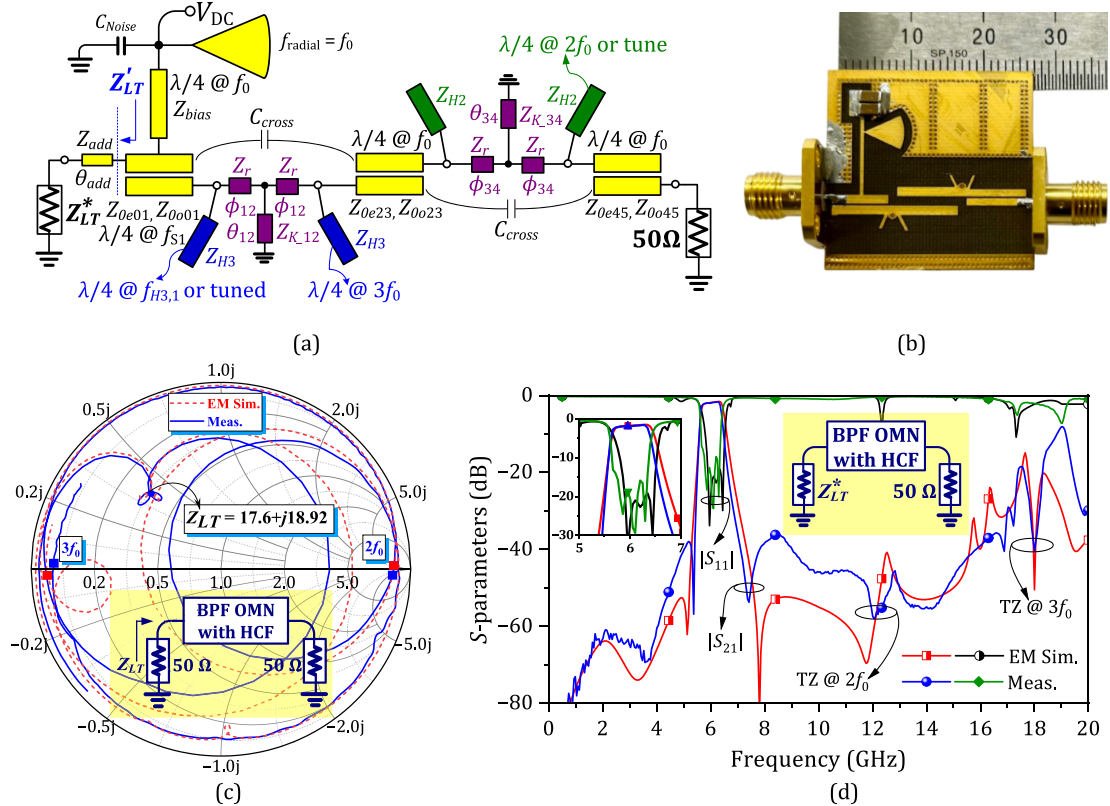


FIGURE 12 | Proposed BPF OMN with HCF: (a) realized structure, (b) fabricated circuit, (c) impedance points on Smith chart, and (d) S-parameters.

harmonics and simultaneously match the target load impedance Z_{LT} at f_0 . Therefore, the proposed BPF OMN does not require the conventional HCN.

The realized structure of the proposed BPF OMN with HCF is presented in Figure 12a. Because the proposed BPF OMN consists of coupled lines at both ends and is difficult to directly connect with the drain pin of the transistor, a very short TL with electrical parameters of $Z_{add} = 70 \Omega$ and $\theta_{add} = 15^\circ$ at f_0 is added to the network. The values of Z_{add} and θ_{add} can be selected according to the width and length of the drain pin. With the additional TL, $Z'_{LT} = 16.43 - j1 \Omega$ is extracted and will be used in the calculation for the design parameters of the proposed BPF OMN. Moreover, the bias circuit utilizing a $\lambda/4$ TL and radial stub is combined with the proposed BPF OMN. The bias circuit must be connected at the center point of the first coupled lines to maintain the desired impedances at $2f_0$ and $3f_0$.

The same design specifications used for CBPF are applied for the proposed BPF OMN. With the complex conjugate of Z'_{LT} , detuning frequency of the first resonator, $f_{s1} = 6.0191$ GHz, is calculated using Equation (3). Because $f_0 = 6$ GHz, the resonant frequencies of the shunt-series $L_{H2}C_{H2}$, $L_{H3}C_{H3,1}$, and $L_{H3}C_{H3,2}$ resonators in Figure 5 are determined as 12, 17.8854, and 18 GHz, respectively. By choosing $L_{H2} = 0.53$ nH and $L_{H3} = 0.29$ nH, other elements of $C_{H2} = 0.3318$ pF, $C_{H3,1} = 0.2731$, and $C_{H3,2} = 0.2695$ pF are calculated using Equation (5). Using Equation (6), the equivalent susceptance and reactance slope parameters can be determined. According to Equation (7), the values of J/K inverters are

TABLE 5 | Calculated values for J/K inverters of the proposed BPF OMN with HCF.

J_{01}	K_{12}	J_{23}	K_{34}	J_{45}
0.006818	3.535018	0.001089	3.535130	0.005740

TABLE 6 | Design parameters of the proposed BPF OMN with HCF.

$Z_{0e01}/Z_{0o01} (\Omega)$	$Z_{0e23}/Z_{0o23} (\Omega)$	$Z_{0e45}/Z_{0o45} (\Omega)$	$\phi_{12} = \phi_{34} (^\circ)$	$\theta_{12} = \theta_{34} (^\circ)$
72.9/38.8	52.9/47.4	68.5/39.8	-4.04	4.06

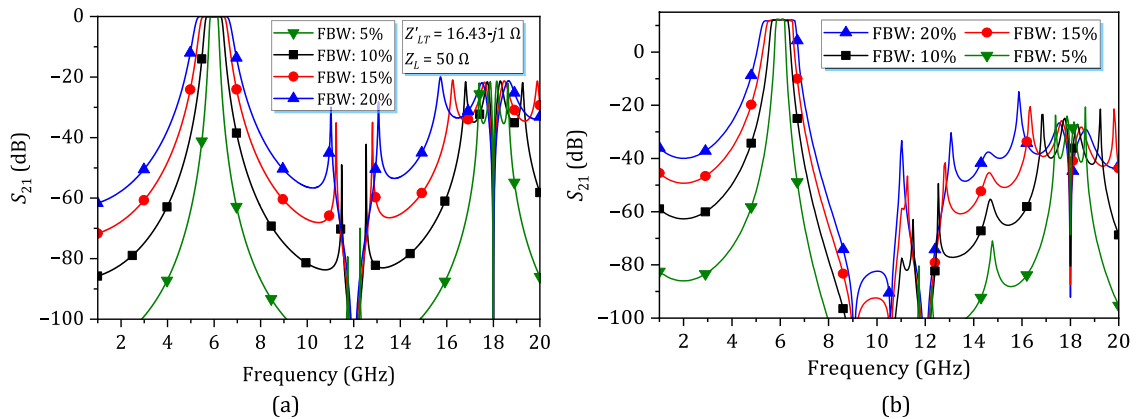


FIGURE 13 | Simulated $|S_{21}|$ according to the different FBWs: (a) BPF OMN with HCF and (b) co-design of class-F⁻¹ MA and BPF OMN with HCF.

determined as can be seen in Table 5. By using Equation (8), the even- and odd-mode impedances of coupled lines are calculated. Similarly, by selecting $Z_r = Z_{Kr,12} = Z_{Kr,34} = 50 \Omega$, the electrical length of TLs in a T-type network can be calculated using Equation (9). Moreover, all determined values are listed in Table 6.

By using coupled lines, the DC-block capacitor can be eliminated from the network. Because of the additional TL with θ_{add} , the electrical length of shunt open stub TL that is equivalent to the series $L_{H3}C_{H3,1}$ resonator must be slightly modified on the realization process for optimum harmonic impedances.

Figure 12b shows the photograph of the fabricated BPF OMN with HCF. According to the addition of TL for soldering and bias circuit, the frequency locations of $2f_0$ and $3f_0$ are shifted from the open and shorted points on Smith chart, respectively, especially for $2f_0$. Therefore, reduction on the length of the green open stub TLs in Figure 12a during EM simulation is required to achieve the desired frequency location of $2f_0$. Similarly, a slight modification on the length of the blue open stub TL is conducted to obtain the desired frequency location of $3f_0$. The comparison of impedance points between EM simulation and measurement is depicted in Figure 12c. The target impedance at f_0 is obtained. The impedances of harmonics are slightly located inside the Smith chart but closed to the open- and short-impedance points. The comparison of S-parameters can be seen in Figure 12d. A good agreement between simulation and measurement results can be achieved. The IL of 1.4 dB and RL of 23.7 dB are measured at f_0 . There are TZs occurred at $2f_0$ and $3f_0$ with the measured attenuation level of 55.4 and 42.1 dB, respectively. Moreover, the 3-dB bandwidth of around 580 MHz is measured from 5.7 to 6.28 GHz.

Typically, the bandwidth of high-efficiency MA can be extended by using feedback network and broadband MNs. For slight bandwidth enhancement, an increased FBW of the proposed

BPF OMN with HCF is enough. Figure 13a presents the simulated $|S_{21}|$ of the proposed BPF OMN with HCF for the different FBWs such as 5%, 10%, 15%, and 20%. It is designed to match Z'_{LT} to 50 Ω . Similarly, the simulated $|S_{21}|$ characteristics of the co-design of class- F^{-1} MA and BPF OMN with the different FBWs are presented in Figure 13b. As can be seen, the bandwidth of the proposed class- F^{-1} MA is increased as the bandwidth of the BPF OMN with HCF increases.

4 | Performance Comparison of Fabricated Class- F^{-1} MAs

As can be seen in Figure 14, three different class- F^{-1} MAs are fabricated and measured. During the experiment process, the CMA is cascaded with a CBPF through a high-Q chip capacitor for the evaluation of the electrical performances as can be seen in Figure 14c. Figure 15 shows the measured S-parameters of all fabricated class- F^{-1} MAs within the frequency range from 1 to 20 GHz. The small-signal gains of CMA, CMA cascaded with CBPF, and the proposed MA with BPF OMN are 11.92, 10.46, and 10.94 dB, respectively. The proposed MA with BPF OMN provided the minimum attenuation levels of 45.3 dB measured from 1 to 5.64 GHz at the lower stopband and 40.45 dB measured from 7.02 to 20 GHz at the higher stopband.

For the output power test, the continuous wave (CW) signal is used. Figure 16 shows the comparison of measured performances between all fabricated class- F^{-1} MAs according to the operating frequency from 5.8 to 6.3 GHz. At f_0 of 6 GHz, the measured gain at 3-dB compression points of CMA, CMA cascaded with CBPF, and the proposed MA with BPF OMN are 9.07, 7.62, and 8.23 dB, respectively, as can be seen in Figure 16a. The measured P_{out} and PAE are presented in Figure 16b,c, respectively. At f_0 , the measured P_{out} and PAE of

CMA are 37.82 dBm and 49.08%, respectively. The P_{out} and PAE of CMA cascaded with CBPF are decreased to 36.62 dBm and 35.27%, respectively, because of the additional IL of the CBPF and not perfect matching between them. However, the P_{out} and PAE of the proposed MA with BPF OMN are 37.04 dBm and 40.31%, respectively. It is necessary to note that all electrical performances of the proposed MA with BPF OMN are superior to those of CMA cascaded with CBPF. Moreover, the circuit size is reduced by about 35.68%. The comparison of the measured electrical performances for all fabricated class- F^{-1} MAs is summarized in Table 7. CMA shows better electrical performance than the proposed MA with BPF OMN because it is not considered frequency selectivity.

However, the electrical performances of CMA are degraded when it is cascaded with CBPF because of the mismatch between them and additional IL. The IL of OMN in the case of

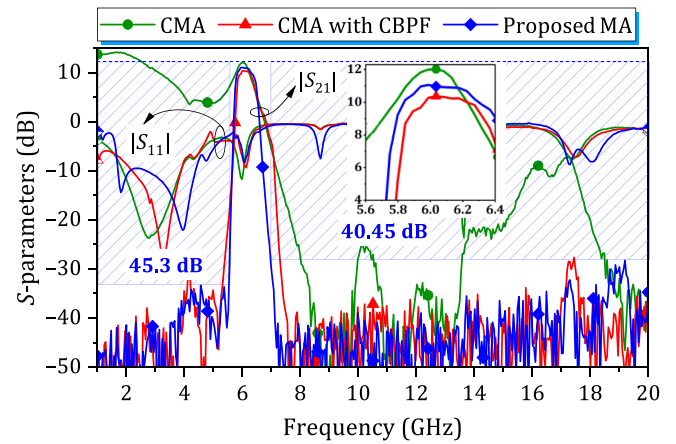


FIGURE 15 | S-parameters comparison of three different class- F^{-1} MAs.

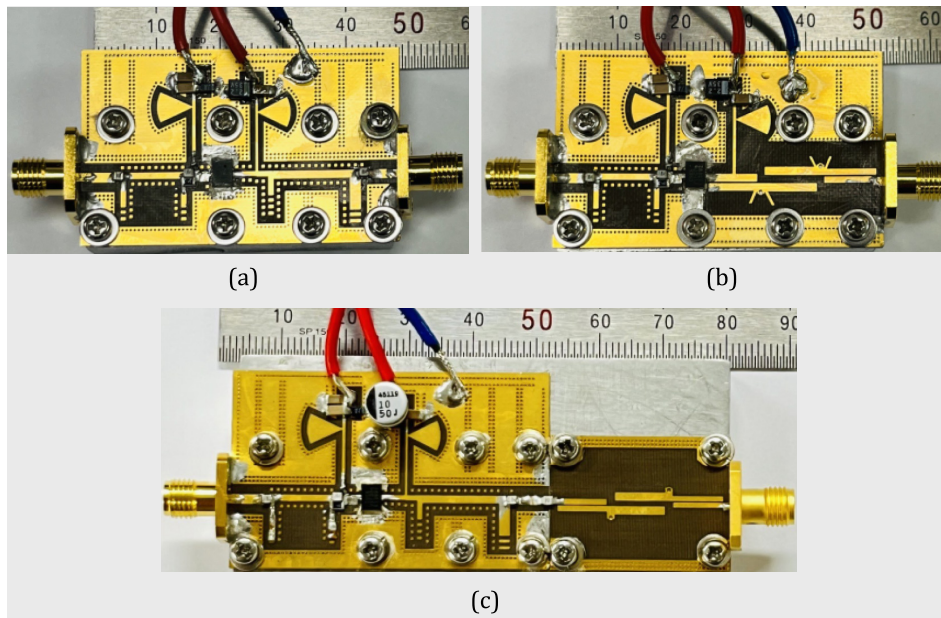


FIGURE 14 | Photographs of three different class- F^{-1} MAs: (a) CMA, (b) proposed co-design of MA and BPF OMN with HCF, and (c) CMA cascaded with CBPF.

CMA cascaded with CBPF is considered as the sum between the ILs of COMN and CBPF.

Figure 17a presents the IIP3 and OIP3 of the proposed class-F⁻¹ MA with BPF OMN. The measured IIP3 and OIP3 are approximately 40.6 and 50.6 dBm, respectively. Similarly, the measured IIP3 and OIP3 of CMA cascaded with CBPF are approximately 40 and 50 dBm, respectively, as can be seen in Figure 17b. The performances of the proposed MA with BPF OMN are compared with those of the state-of-the-art MAs in Figure 18. The proposed MA with BPF OMN is implemented using microstrip lines, which is a cheap design process. Because the frequency selective

is not considered in [26], the P_{out} and PAE of that MA will be degraded when cascaded with BPF. Furthermore, because the BPF OMNs in [11, 12] are realized using cavity resonators, the fabrication process is more difficult, and their circuit sizes are bulky. In [13], the BPF OMN are designed using SIW cavity. The IMN, bias circuit, transistor connection parts, and SIW BPF OMN are implemented on the PCB with different thickness. Therefore, it is difficult to properly connect these circuits together to form a MA. Moreover, The BPF OMN with HCF is not considered in [11–13]. In this work, the PAE of the proposed co-design of MA and BPF OMN with HCF is improved by about 5.1% over that of CMA cascaded with CBPF, while the P_{out} is improved by about 0.42 dB.

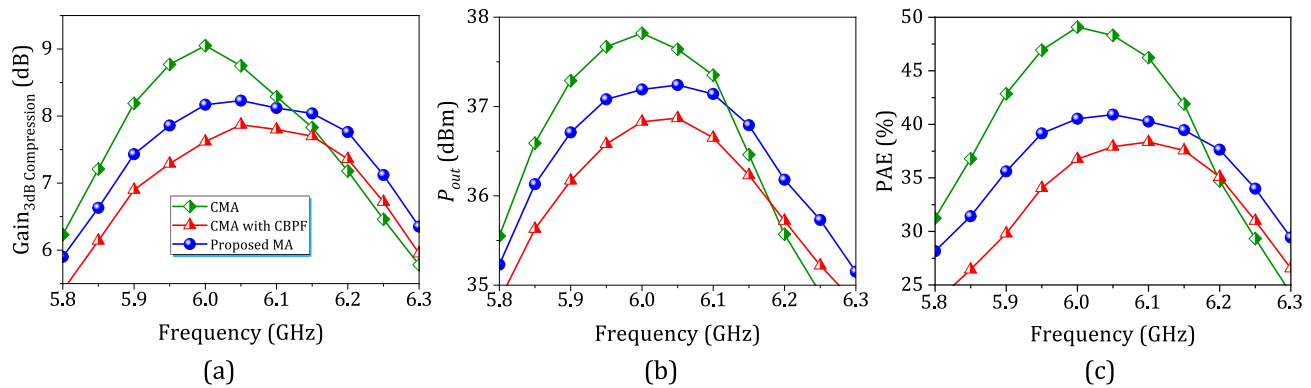


FIGURE 16 | Measured electrical performances of three different class-F⁻¹ MAs: (a) gain at 3-dB compression point, (b) P_{out} , and (c) PAE.

TABLE 7 | Performance comparison of the three fabricated different class-F⁻¹ MAs at $f_0 = 6$ GHz.

MAs	$ S_{21} $ (dB)	P_{out} (dBm)	PAE (%)	DE (%)	Size (mm ²)	IL of OMN (dB)
CMA	11.92	37.82	49.08	53.47	48.1 × 30	0.35
CMA + CBPF	10.46	36.62	35.27	39.02	76.5 × 30	1.84
Proposed MA	10.94	37.04	40.31	44.12	49.2 × 30	1.4

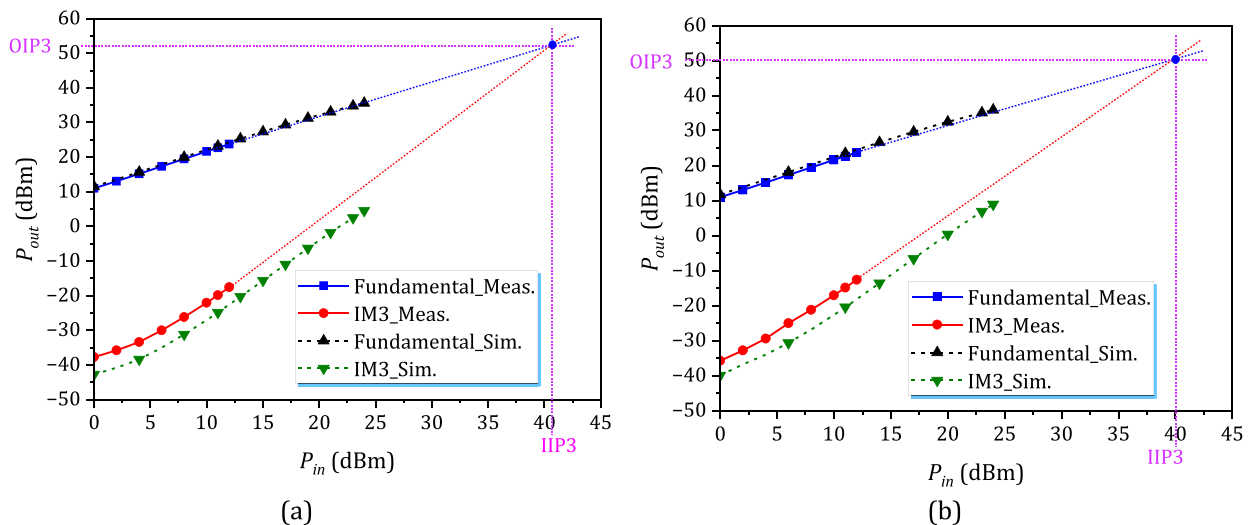


FIGURE 17 | IIP3 and OIP3 of the (a) proposed class-F⁻¹ MA with BPF OMN and (b) CMA cascaded with CBPF.

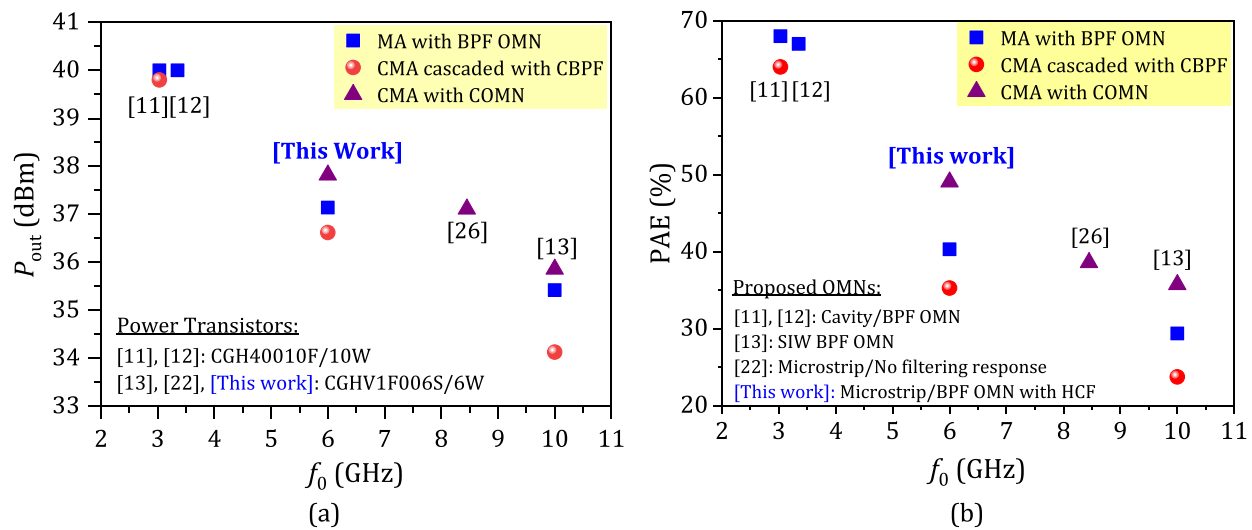


FIGURE 18 | Performance comparison with state-of-the-art MAs: (a) P_{out} and (b) PAE.

5 | Conclusion

A co-design approach of a class-F⁻¹ MA and BPF OMN with HCF is proposed and investigated in this paper. To show the validity of the proposed class-F⁻¹ MA, three kinds of class-F⁻¹ MAs are fabricated and measured. The proposed MA with BPF OMN can remove CBPF cascaded with CMA. Therefore, the electrical performance, overall circuit size, and the reduction of complexity of the transmitter in the RF front end are improved simultaneously. The proposed design method can also be applied to the harmonic control/tuning MA design to improve the efficiency and suppress out-of-band signal to secure the high quality of the desired signal. Altogether, the electrical results indicate that the proposed BPF OMN with HCF design method can be applied to microwave circuit and system designs.

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Data Availability Statement

The data that support the findings of this study are available from the corresponding author upon reasonable request.

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