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FEATURES OF CENTIMETER-BAND FILTER-BANK DESIGN BASED ON GaAs pHEMT-TECHNOLOGY

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Abstract. This article presents the results of the filter-bank design in the centimeter range. The filters are made in the form of microwave monolithic integrated circuits based on domestic GaAs pHEMT technology. The filter-bank includes bandpass filters operating in four subbands of the total frequency band 5.8 ... 18.2 GHz. The developed filters have VSWR of no more than 1.5 in the passband. Stopband suppression at 30% offset or more from the passband center frequency is more than 45 dB. When constructing filters of different subranges, different implementation options were used: lumped filters and microstrip filters based on interdigital and hairpin structures. Using the example of the microstrip bandpass filters design, the article discusses the features of modeling microwave monolithic integrated circuits in the AWR Design Environment.

Keywords: microwave, bandpass filter, GaAs pHEMT, interdigital, hairpin, electromagnetic simulation

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ОСОБЕННОСТИ РАЗРАБОТКИ БАНКА ФИЛЬТРОВ САНТИМЕТРОВОГО ДИАПАЗОНА НА ОСНОВЕ GaAs pHEMT-ТЕХНОЛОГИИ

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Аннотация. В данной статье представлены результаты разработки банка фильтров сантиметрового диапазона, выполненных в виде СВЧ монолитных интегральных схем на основе отечественной GaAs pHEMT-технологии. Банк включает в себя полосовые фильтры, работающие в четырех поддиапазонах общей полосы частот 5,8 ... 18,2 ГГц. Разработанные фильтры банка имеют в полосе пропускания КСВН не более 1,5, при подавлении в полосе заграждения не хуже 45 дБ на отстройках от центральной частоты, превосходящих ±30%. Вносимые потери в полосе пропускания не превосходят 6 дБ в низкочастотной и 3 дБ в высокочастотной части исследуемого диапазона. При построении фильтров разных поддиапазонов были использованы различные варианты реализации: фильтры на сосредоточенных элементах и микрополосковые фильтры на встречно-штыревых и шпилечных структурах. На примере разработки микрополосковых полосовых фильтров в статье рассмотрены особенности моделирования СВЧ монолитных интегральных схем в среде AWR Design Environment.

Ключевые слова: СВЧ, полосовой фильтр, GaAs pHEMT, встречно-стержневой, шпилечный, электромагнитное моделирование

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Introduction

One of the main elements of a preselector in modern radio-receiving microwave devices is a filter-bank, which includes a set of bandpass filters (BPF). The characteristics of such partial BPFs – the main components of the filter-bank – determine several important parameters of the entire radio-receiving device. When developing bandpass microwave filters in the form of monolithic microwave integrated circuits (MMIC), they can be designed as lumped or distributed element circuits and implemented in various transmission line structures [1–3].

Bandpass filters based on microstrip structures are often used in the microwaves due to their compact size and high yield. Several examples of simplified microstrip BPF's circuits that are widely used in practice are shown in Fig. 1: *a*) interdigital, *b*) hairpin, *c*) open-loop [1, 4].

Despite the large number of works devoted to the microwave filters studies (e.g. [1–9]), the choice of the basic structure and the development of the final circuit in each case is a separate, rather labor-intensive task. An example of such a task is the development of a microwave bandpass filter-bank that covers a large frequency band (an octave or more). In this case, it is necessary, based on a single MMIC technology, to

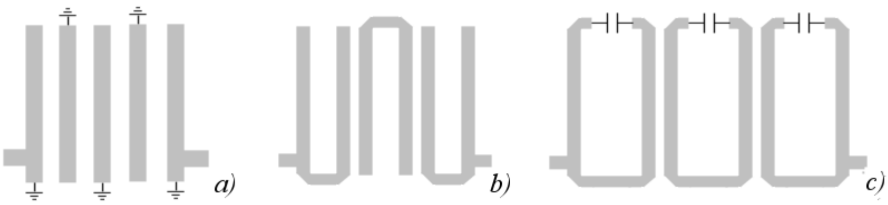

Fig. 1. Simplified schemes of microwave microstrip bandpass filters

build several circuits that provide the required characteristics in several frequency sub-ranges of preselectors of broadband radio receiving devices [10, 11].

Introduction of monolithic microwave integrated circuits in the creation of filters allows drastic reduction of the mass, size, labor intensity of manufacturing, while increasing reliability and improving the repeatability of characteristics. In conditions of mass serial production, it also reduces costs per unit of production [2]. At the same time, it is important to note that due to permanent restrictions on the supply of electronic devices and dual-use systems to Russia it is becoming increasingly difficult to obtain the required products from abroad [12]. In this regard, the expansion of the microelectronic microwave devices nomenclature, produced on the basis of domestic technology and optimization of the process of their modeling are of particular relevance [12, 13].

This paper describes the design of the bandpass filter-bank on the example of studies of various MMIC BPF's circuits covering the frequency range, including most of the C-, X-, and Ku-bands in four sub-ranges: No. 1 – 5.8 ... 8.2 GHz, No. 2 – 7.8 ... 11.2 GHz, No. 3 –10.8 ... 15.2 GHz, No. 4 – 14.8 ... 18.2 GHz.

The following frequency responses were used as required performance for the developing bandpass filters: insertion loss in sub-range No. 1 – 6 dB, No. 2 – 5 dB, No. 3 – 5 dB, No. 4 – 4 dB, VSWR in passband is not more than 1.6, stopband suppression is no less than 45 dB at 30% offset and more from the passband center frequency.

The filters under study are designed based on the domestic GaAs pHEMT technology using complex design tools for solid-state microwave devices for the 0.25 μm technological process (PDK_pHEMT025D) [14–16].

LC-bandpass filters on lumped elements

At the comparative study of various bandpass filters operating in the low-frequency part of the investigated frequency range (sub-ranges No. 1: 5.8 ... 8.2 GHz and No. 2: 7.8 ... 11.2 GHz), it was found that it is reasonable to use LC-filter circuits on lumped elements as basic schemes. The simulation showed that the use of microstrip structures (Fig. 1) as basic circuits of bandpass filters at these frequency leads to excessively large dimensions of the MMIC chip (more than 5000×5000 μm)

The conducted studies of the idealized filter prototype have shown that to obtain the required characteristics it is necessary to choose filters with frequency responses having transmission zeros in the stopband. In addition, the simulation of an idealized filter prototype showed that the filter order should be quite large. Therefore, in this paper a quasi-elliptic filter on lumped elements of the 9th order was chosen as a basis for the BPF construction of the 1st and 2nd sub-ranges. An equivalent circuit of such an idealized bandpass LC-filter with transmission zeros near the passband edges is shown in Fig. 2, *a* [1].

When selecting the scheme and order of the idealized prototype filter, the initial parameters should be used with a certain margin both in terms of insertion loss in the passband (a few dB) and rejection in the stopband (up to 50 dB).

Besides, the results of the research have shown that it is expedient to increase values of the central operating frequency and a bandwidth (in comparison with initial data). Further design of the filter generally leads to narrowing of the initial bandwidth and some shift to the low frequency area. For example,

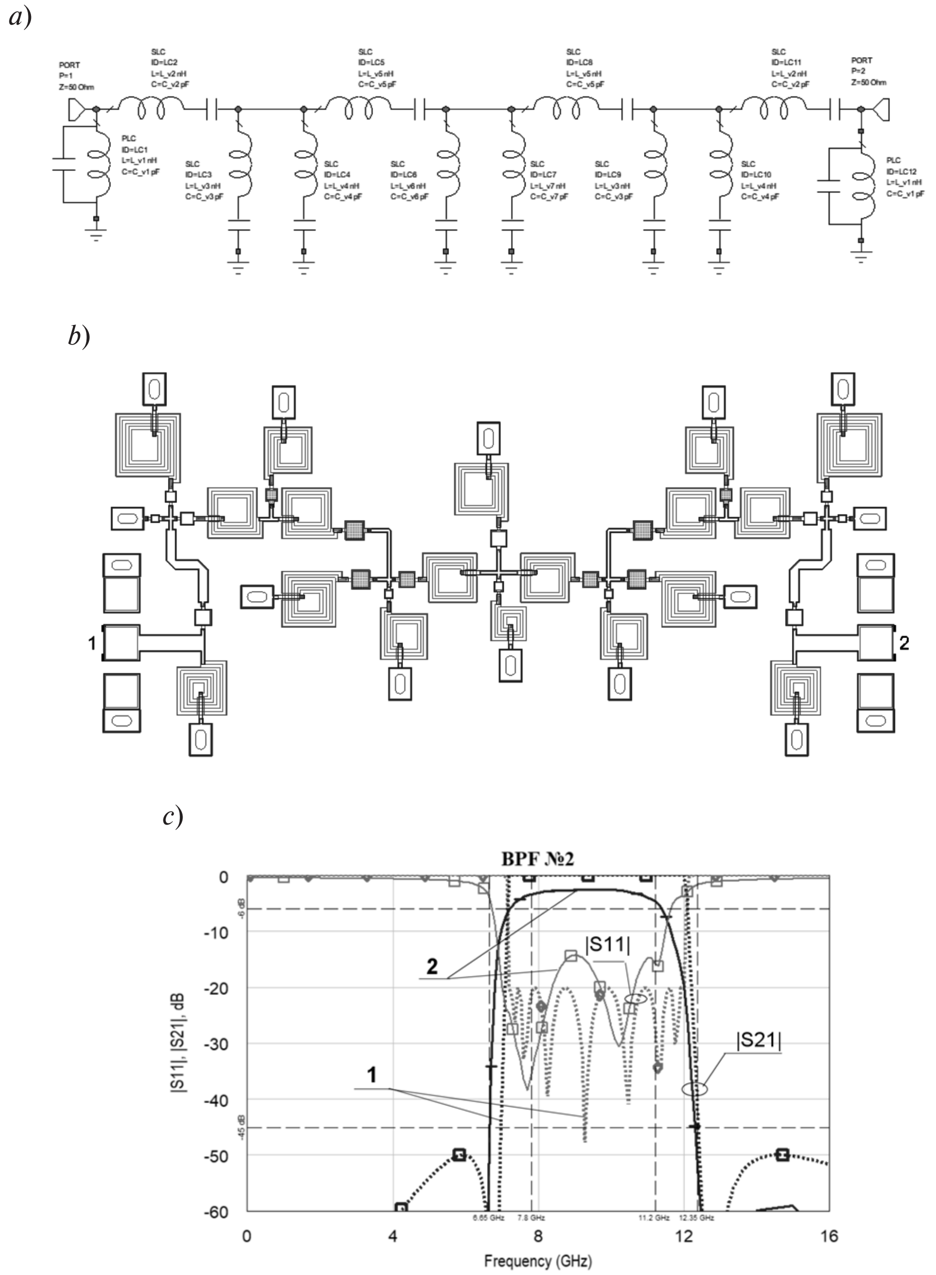

Fig. 2. Schematic of idealized BPF No. 2 (*a*), MMIC layout (*b*) and frequency responses (*c*): 1 – idealized BPF, 2 – MMIC

when designing 1st sub-range BPF (BPF No. 1) at the given values of the center frequency f_0 = 7 GHz and 6.8 ... 8.2 GHz bandwidth it is reasonable to set a significantly larger bandwidth of the initial prototype to 5.45 ... 8.65 GHz with the center frequency f_0 = 7.1 GHz.

When designing the BPF of the 2nd sub-range (BPF No. 2) these corrections, as simulation has shown, make the following values: at given values of the center frequency f_0 = 9.5 GHz and 7.8 ... 11.2 GHz bandwidth it is necessary to set the bandwidth of the initial prototype to 7.2 ... 12 GHz with f_0 = 9.6 GHz.

As an example, Fig. 2 shows: the circuit of the initial idealized BPF No. 2 (Fig. 2, *a*), MMIC topology designed on its basis (Fig. 2, *b*) and BPF frequency responses |S11|, |S21| – insertion loss and input return loss. In the Fig. 2, *c*, the dashed lines show the responses of the idealized circuit (curves 1), and the solid

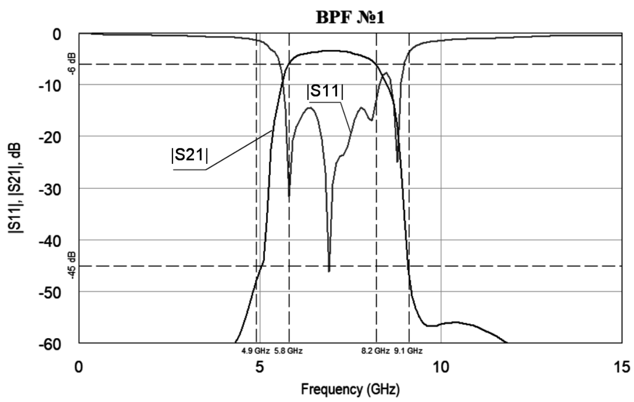

Fig. 3. S-parameters of BPF No. 1

lines show the simulation result of the final MMIC (curves 2). It should be noted that MMIC design included two stages: basic circuit design using process design kit (PDK) and MMIC topology design using electromagnetic (EM) simulation.

The frequency responses of the MMIC BPF No. 1 is shown in Fig. 3.

The designed BPFs No. 1 and No. 2 have the following characteristics: insertion loss is no more than 6 dB for sub-range No. 1 and 4.6 dB for sub-range No. 2 (3.5 dB and 2.5 dB at center frequency f0 for No. 1 and No. 2, respectively), return loss is not worse than 14.5 dB (input and output VSWR for both filters no more than 1.5). Stopband suppression at 30% offset or more from the passband center frequency is not worse than 45.5 dB for sub-range No. 1 and 49 dB for sub-range No. 2. MMIC chip of BPF No. 1, 2 have dimensions less than 2000×2600 μm.

Bandpass filters on microstrip structures

According to the results of the studies about filters operating on higher frequency ranges (BPF No. 3 – 10.8 ... 15.2 GHz and BPF No. 4 – 14.8 ... 18.2 GHz), filters on microstrip structures are preferred when selecting initial schemes (Fig. 1). It is easier to provide the required level of insertion loss using microstrip structures in comparison to LC-filters on lumped elements. Comparative modeling of microstrip bandpass filters showed that the best choice in this frequency range would be the schemes of the hairpin (Fig. 1, *a*) and interdigital (Fig. 1, *b*) filters. In this case, four microstrip structures were considered. Along with the two mentioned above, open-loop filters (Fig. 1, *c*) and filters on stepped microstrip resonators were also studied. Filters based on open-loop resonators and stepped microstrip resonators showed generally worse performance. In particular, the dimensions of open-loop microstrip structures are significantly larger in comparison with hairpin and interdigital structures (in this case up to 9000 μm or more).

To determine filter orders needed to achieve the required parameters, studies of corresponding idealized prototype filter circuits were conducted. Studies showed, that when designing interdigital and hairpin filters, it is advisable to use prototypes of the 15th and 7th orders, respectively.

The standard topology of microstrip elements in the PDK_pHEMT025D library consists of two metallization layers. The first layer with a thickness of 1 μm is formed by sputtering, while the second layer is formed by chemical deposition, with a thickness about 5 μm. As a result of deposition, the microstrip does not have a strictly rectangular profile, and its thickness can vary in length. To increase the yields of the MMIC, it was decided to use only the first layer of metallization during design. Fig. 4 shows the resulting designed MMIC topologies: the interdigital filter – BPF No. 3 (Fig. 4, *a*) and the hairpin filter – BPF No. 4 (Fig. 4, *b*).

The simulation showed that the filter circuits require input and output matching circuits to achieve the required frequency responses. Some of the matching circuits include both microstrip line segments and lumped LC-elements, for example, a MIM-capacitor for the hairpin filter circuit (Fig. 4, *b*).

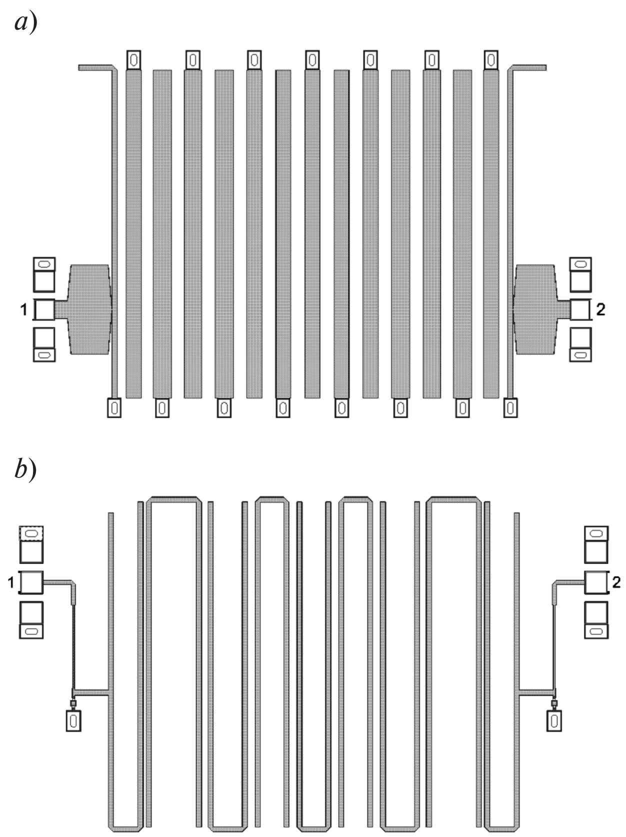

Fig. 4. MMIC topologies of bandpass filters No. 3 (*a*) and No. 4 (*b*)

When selecting the basic circuits, simulation of the two above mentioned filters was carried out in both sub-ranges No. 3 and No. 4. In the considered sub-ranges interdigitated and hairpin filters have similar performance and satisfy the required characteristics.

Detailed comparison of these two filters performances (insertion loss, return loss, selectivity and chip size) allows us to make a choice in sub-range No. 3 in favor of an interdigital BPF (Fig. 4, *a*), and in sub-range No. 4 in favor of a hairpin BPF (Fig. 4 *b*).

Fig. 5 shows the characteristics of the interdigital (sub-range No. 3) and hairpin filter (sub-range No. 4) in the form of frequency responses |S11| and |S21|. In Fig. 5 the dashed lines show the responses of the BPF No. 3 (curves 3), and the solid lines – BPF No. 4 (curves 4).

The designed MMIC of the BPF No. 3 has the following frequency responses: the minimum value |S21| = –4.2 dB at the frequency f = 15.2 GHz in the passband. Stopband suppression at 30% offset or more from the passband center frequency is more than 49 dB. Return loss is not worse 14 dB (input and output VSWR is not more than 1.5).

The designed MMIC of the BPF No. 4 has following performance: maximum value of insertion loss in the passband is 3.1 dB at the frequency f = 18.2 GHz and 2.4 dB at the center frequency f_0 = 16.5 GHz of the passband. Stopband suppression at 30% offset or more from the passband center frequency is more than 48.5 dB. Return loss is not worse than 25 dB in passband (input and output VSWR is not more than 1.1). MMIC sizes of both BPF No. 3 and No. 4 do not exceed 2100×3100 μm.

The MMIC filter simulation features

A distinctive feature of microwave circuits topology development, as noted above, is the need for electromagnetic (EM) modeling at the final stage of design. The results of EM simulation often differ significantly from the results of circuit simulation based on library elements [17–21]. This difference turns out

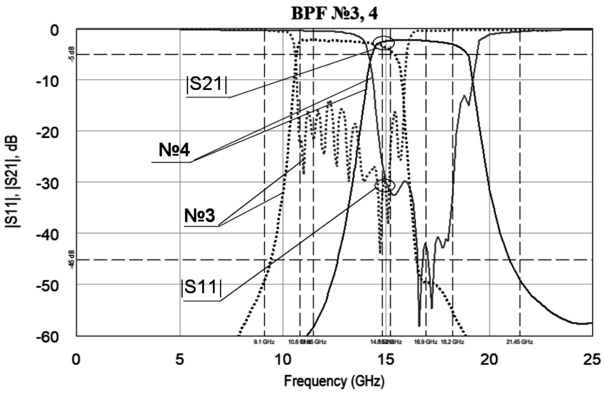

Fig. 5. S-parameters of BPF No. 3 and No. 4.

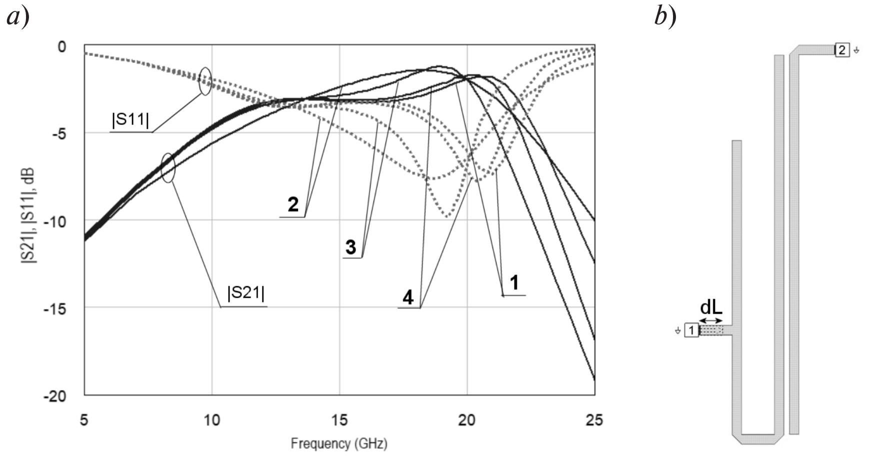

Fig. 6. Frequency responses of the test circuit (*a*) and a part of the test structure topology (*b*)

to be more essential on the higher frequency range as well as with more complex chip design. At the same time, to achieve the required parameters and define the final chip topology, it is often necessary to conduct many EM simulation cycles. Therefore, methods to optimize the chip design process were considered during filter development. A study of the effect of different MMIC design ways on the resulting characteristics was carried out. For this purpose, a fragment of the hairpin BPF circuit was considered as a test sample (Fig. 4, *b*). This test circuit was divided into two subcircuits and on their basis the following variants of forming topology and simulating characteristics were implemented:

1. EM simulation of a complete test structure. This simulation variant was used as a reference.
2. Circuit simulation of the test structure based on library elements.
3. EM simulation of two subcircuits of the test structure separately. Simulation of a two-stage circuit, where EM models act as composite stages.

Fig. 6, *a* shows the frequency response simulation of the considered test circuit obtained by three different methods. Numbers 1, 2 and 3 in the graph indicate three different simulation ways described above.

As follows from the graphs in Fig. 6, *a*, the results of circuit simulation (Fig. 6, *a*, curves 2) and EM simulation (Fig. 6, *a*, curves 1) are significantly different. EM simulation of the divided the test structure (Fig. 6, curves 3), allows to slightly approximate the responses to the reference ones. However, the discrepancy is still quite significant. This may be a consequence of the unaccounted electromagnetic interaction between the individual stages of the whole circuit. This situation occurs in the modeling for option 3.

The studies have shown that the use of de-embedding procedure with reference plane shifting can help to reduce the differences between the full EM model of the whole circuit (Fig. 6, *a*, curves 1) and the composite one (Fig. 6, *a*, curves 3) consisting of EM models of subcircuits [18, 21]. Fig. 6 *b* shows the topology of one of the test structures subcircuits, indicating the value dL of the input port reference plane shift. The result of modeling a two-stage circuit, where EM models of individual subcircuits with a shift value of the reference plane dL = 30 μm act as composite stages, is marked with number 4 in Fig. 6, *a*.

As shown in Fig. 6, the use of reference plane shift makes it possible to significantly reduce the difference between the frequency responses of a complete EM model of a single circuit (Fig. 6, a, curves 1) and the frequency responses of a two-stage circuit, where EM models act as composite stages (Fig. 6, *a*, curves 4). Design and simulation of such composite circuit are much faster. The results of modeling have shown that it is reasonable to choose the value of such shift in the range of 20 ... 50 μm for the considered structures.

Comparison with State-of-the-Art Filters

The characteristics of the filters studied in this work and their foreign analogues, based on monolithic GaAs MMICs, are provided in Table 1.

Table 1

Characteristics of filters based on monolithic GaAs MMICs

Ref.	f_0, GHz	$f_1 - f_2$, GHz (FBW %)	$IL(f_0)$, dB	Ripple, dB	$RLmin$, dB	Reject., dB	Size, mm^2
BWBF-8/12 [22]	10	8-12 (40%)	1.7	2	15	>25	1.42
BWBF-12/16-7C3 [22]	14	12-16 (28%)	2.5	1	12	<20	1.28
PDBF-15R7/17R7-D2 [22]	16.7	15.7-17.7 (12%)	2	1*	11	40	5.12
XBF-163-D+ [23]	16	15.5-16.5 (6.3%)	4	N/A	17	>50	3.44
MFBP-0002CH [24]	6.65	5.9-7.4 (22%)	1.5	1*	18	45	5.76
MFBA-0003CH [24]	12	10.1-14.1 (33%)	2.1	1*	15	44	9.6
MFBA-00001CH [24]	16	14.1-17.9 (22%)	2.4	1.2*	17	45	9.6
This work BPF#1	7	5.8-8.2 (34%)	3.5	2.5	14.5	46	4.04
This work BPF#2	9.5	7.8-11.2 (36%)	2.5	2.2	14.2	48	3.46
This work BPF#3	13	10.8-15.2 (34%)	2.3	1.9	14	49	6.41
This work BPF#4	16.5	14.8-18.2 (21%)	2.4	0.8	25	48.5	5.36

* – approximate

The table shows the following parameters: f_0 – central frequency of the passband; $f_1 - f_2$ (FBW) – minimum and maximum frequencies of the passband (relative bandwidth); IL – insertion loss at the central frequency; Ripple – ripple in the passband; $RLmin$ – minimal return loss in the passband; Reject. – stopband suppression (at offset from the center frequency $f_0 \pm 0.3f_0$ and more), Size – MMIC size. A comparison of the given data shows that filters BPF No. 1–3 are somewhat inferior to some samples in terms of the chip size [22], as well as the ripple in the passband [22, 24]. At the same time, the developed BPFs are highly selective. According to this characteristic, the filters under study are superior to most of the presented samples. In addition, for example, BPF No. 4 turns out to be better than filters of a similar range in other parameters [22, 24]. As shown, using the domestic GaAs pHEMT technology, the overall performance of the designed BPFs is very competitive.

Conclusion

The paper presented a comparative study of various implementations of bandpass filters on monolithic integrated circuits, covering the most part of C-, X- and Ku-bands. The results show that in the lower part of the considered frequency range it is expedient to build filters on the basis of lumped elements, while in the high-frequency part bandpass filters on microstrip structures have the advantage over LC-filters, among which hairpin and interdigital filters have demonstrated the best characteristics.

To reduce the number of required EM simulation cycles and decrease the total design time of a MMIC, it is advisable to divide the designed circuit into several subcircuits with step-by-step EM simulation of individual stages.

As a result of this research, a filter-bank was designed, including four MMIC bandpass filters overlapping the total operating bandwidth of 5.8 ... 18.2 GHz. The partial BPFs of filter-bank are designed on the basis of domestic GaAs pHEMT-technology using comprehensive tools for designing solid-state microwave devices (PDK_pHEMT025D) and have a bandwidth of 20–35%. All four designed filters have VSWR of no more than 1.5 in the passband. Stopband suppression at 30% offset or more from the passband center frequency is more than 45 dB.

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