International Journal of Electrical and Computer Engineering (IJECE)
Vol. 14, No. 4, August 2024, pp. 4183~4194
ISSN: 2088-8708, DOI: 10.11591/ijece.v14i4.pp4183-4194

Method for design and implementation of telecommunication devices for aircraft

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Article Info

Article history:

Received Feb 2, 2024
Revised Mar 6, 2024
Accepted Mar 9, 2024

Keywords:

Electrical filters
Frequency-selective circuits
Microwave technology
Telecommunication devices
Transmission lines

ABSTRACT

This article highlights the importance of electrical filters in radio engineering devices, emphasizing their role in transmitting signals in the transparency band and suppressing signals in the stop band. We examined methods for designing frequency-selective circuits with lumped parameters, which, in general, are a complete field of radio engineering and allow the synthesis of devices of varying complexity. The focus of the article is the frequency region, where the distributed properties of the synthesized structures appear. The article also provides an overview of various methods for synthesizing ultra-high frequency (UHF) filters. It is emphasized that for low-pass filters a transition from a low-frequency prototype to a high-frequency representation is applied, which, despite the crudeness of the approach, provides satisfactory results that can be improved at the production stage. The article also discusses various methods for implementing bandpass filters on distributed elements, including the use of short-circuited and open-circuited stubs, as well as weak-coupled lines. In conclusion, the paper highlights the need to improve these methods to improve process accuracy and make filter designers more efficient in radio engineering.

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1. INTRODUCTION

This article discusses an important direction in the synthesis of frequency-selective circuits of ultra-high frequencies (microwaves) - step transformers [1]–[3]. The main difference between these filters is the introduction of an approximation stage, which brings additional capabilities to the design process [4]–[6]. Stepped transformers [7]–[9], despite their widespread use for matching wave impedances [10], [11], often pose a challenge in achieving high selectivity rates of up to 150%. An important aspect that is worth highlighting is the use of computer-aided design (CAD) systems [12], [13] to create structures of telecommunication devices [14]–[16] on sections of transmission lines. CAD systems based on parametric synthesis methods provide an opportunity for developers to obtain the required structures, even without experience in this field. However, the main disadvantage of such systems is the limited control of the developer over the synthesis process.

The purpose of this article is to present a method for synthesizing filters on elements with distributed parameters, including an approximation stage. This method combines the advantages of the classical approach, providing the designer with greater control and achieving optimal results in the design and implementation of telecommunications devices for aircraft. Kokkoniemi *et al.* [17] examines the problem of providing reliable wireless communications in aircraft located far from satellites and base stations. To overcome the capacity limitations of traditional frequencies, the terahertz range (0.3-10 THz) is used [18]–[20]. A detailed channel model for airborne terahertz communications is proposed, taking into account the features of this range. The study shows that the use of terahertz links between aircraft and satellite can provide throughputs of 50 to 150 Gbps, which significantly exceeds the current capabilities of in-flight wireless networks and allows high data rates for passengers and crew throughout the flight. Shamim *et al.* [21] discusses the use of graphene in wireless communications, especially in terahertz applications, due to its amazing electrical, mechanical and optical properties. It presents a graphene-based antenna element for a microstrip printed antenna with a resonant frequency of 0.72 THz for wireless communication. The proposed antenna exhibits 37.50% impedance bandwidth from 0.53 to 0.84 THz with a center frequency of 0.72 THz. The results are presented in terms of reflection coefficient ($s11<-10$ dB), voltage standing wave ratio (VSWR), input impedance, and E-plane and H-plane radiation patterns. Design and simulation were carried out using computer simulation technology (CST) microwave studio electromagnetic simulator based on the finite-state method. differences in the time domain. The simulation results show a minimum reflectance of −59.97 dB, a VSWR of 1.007, and a good radiation pattern at a resonant frequency of 0.72 THz, making this antenna an excellent candidate for future wireless communications as well as medical imaging, defense systems, explosives detection, and materials characterization.

Xiao *et al.* [22] present a novel 1-bit Ku-band reconfigurable radiation array antenna (RTA) based on the Yagi-Vivaldi unit cell. The cell includes tightly coupled Yagi and Vivaldi microstrip antennas for reception and transmission, respectively. To achieve reconfigurable performance, a new phase shift from stub line to microstrip line was introduced using p-i-n diodes, providing a 180° phase shift based on a current variation mechanism. To improve performance at the incident angle, a metasystem based on sub-wavelength metal crosses in front of the antenna is introduced, which leads to an improvement in H-plane scanning loss by 2 dB. A prototype RTA with 16×16 elements was designed, manufactured and tested. The measured 1 dB bandwidth is 14% with a maximum gain of 22.3 dB at 13.6 GHz corresponding to an aperture efficiency of 25.6%. Achieved good beam scanning performance with scan loss less than 4 dB in both E-plane and H-plane at 60° angle. Wu *et al.* [23] provides a general overview of substrate-integrated transmission lines from the perspective of the historical context and progress of waveguide structures, as well as their impact on the development of microwave circuits and integration solutions. The technology path is covered, including five categorized generations of microwave circuits. In particular, substrate integration technologies are reviewed and discussed, with an emphasis on technical considerations, key design considerations, component development, structural evolution, and systems integration. Examples are provided demonstrating milestones in research and development in substrate-integrated transmission line technologies, with a particular focus on substrate-integrated waveguide (SIW) techniques. Practical applications and industrial interests are presented with key references and technical results highlighting product developments in end-use industries.

Ahmed *et al.* [24] reviews the further development of a highly efficient and durable unmanned aerial vehicle (UAV) for monitoring power lines. The project involves developing communication between the UAV, the ground control station and the mobile computing station via a 4G LTE network to improve the high-resilience system. The UAV will be equipped with the ability to wirelessly transmit video using a camera mounted on a special gimbal. The system will implement computer vision for video processing, image filtering and object avoidance. Monitoring of power lines will be carried out using testing and application of edge filters and edge detection. Xie and Wan [25] discusses the application of digital twin technology in smart manufacturing using new information technologies such as big data and artificial intelligence. Introducing the digital twin into production lines addresses the challenges associated with visualization and complex cyber-physical integration in discrete manufacturing. However, the application of this technology still faces problems of low simulation accuracy, response delay, and insufficient production line control precision. This paper proposes a four-dimensional digital twin modeling method to solve these problems by incorporating the information dimension into traditional digital twin modeling methods. The proposed method integrates geometric, physical, behavioral and control aspects to provide support for virtual-real intelligent mutual control. The effectiveness of the method has been confirmed by experiments on the production line, which opens up prospects for the digital transformation of discrete manufacturing enterprises.

2. METHOD

Distributed transmission lines, including short-circuited and open-circuited transmission lines, have received increased attention in the design of filters and other electronic devices. These lines have unique characteristics that can be compared to lumped reactance's, giving them specific electrical properties. In our study, we paid special attention to the input impedance of a short-circuited line under low-loss conditions. During a short circuit, this characteristic is determined in a specific way (1), which becomes important when designing filters and other electronic systems. The study and application of distributed transmission line parameters is becoming an integral part of the methodology required to achieve the required system performance in modern electronic devices.

$$Z_{in\_kz} = jZ_v tg(\beta l) \tag{1}$$

For a section of transmission line open at the end, the input resistance has the form (2):

$$Z_{in\_hh} = -jZ_v / tg(\beta l) \tag{2}$$

where Z_v is characteristic impedance of short-circuited section. βl is electrical length of the segment. From expressions (1), (2) it follows that the input resistance of the short-circuited and open-circuited segments at the end depends on the length *l* and is inductive or capacitive in nature, respectively. Half-wave and quarter-wave sections of transmission lines have a number of properties that are useful in the design of microwave devices. Consider the expression for the input resistance of a lossless transmission line segment (3):

$$Z_{in} = \frac{Z_n + Z_v tg(\beta l)}{1 + (Z_n/Z_v) tg(\beta l)} \tag{3}$$

At $\beta l = n\pi$, where *n*: integer, tangent value is zero. Substituting this value into (3) we have $Z_{in} = Z_n$. The value $\beta l = n\pi$ corresponds to the length of the transmission line segment, which is a multiple of half the wavelength. Therefore, the input impedance of a half-wave segment is equal to the value of the resistance connected to its output. Substituting into (3) the value $\beta l = n(\pi/2)$, n - is an odd integer, the tangent function tends to infinity (3.1), from where

$$Z_v = \sqrt{Z_{in} Z_n} \tag{3.1}$$

This property allows the use of transmission line segments with an odd number of quarter-wave elements with characteristic impedance Z_v to match active resistances $Z_{in},\ Z_n$. As noted, a distinctive feature of the classical theory of constructing matching-filtering devices is the presence of an approximation stage. Its essence lies in the fact that the transfer function that the device being developed must describe is specified initially. The Butterworth, Chebyshev, and Legendre functions are most often used. In wave theory, this approach is implemented only for the design of step transformers and filters on series-connected segments of the same length with different wave impedances. It is based on replacing the frequency variable [4] of the form

$$\omega^{2n} = (-1)^n \left(\frac{\alpha^2 (i\ \lambda)^2}{1 - (i\ \lambda)^2}\right)^n \tag{4}$$

where *n*: filter order, α^2: coefficient determining the filter bandwidth (4.1):

$$\lambda = tg(\beta l);$$
$$\beta = 2\pi / (c_{ef}/f); \tag{4.1}$$

where c_{ef}: wave speed in the transmission line, i: imaginary unit. The second-order Butterworth transfer function after such a transformation takes the form (5):

$$K(\lambda^2) = \frac{1}{1+\left(\frac{\alpha^2 (i\,\lambda)^2}{1-(i\,\lambda)^2}\right)^2} \quad (5)$$

The electrical circuit described by expression (5) contains two sections of a transmission line of the same length with different wave impedances; the frequency response is shown in Figure 1. Despite the fact that the filter is loaded on both sides with a standard resistance, with this approach it is difficult to obtain high selectivity indicators. This is due to the fact that all zeros of the transfer of the classical Butterworth function located at infinity are replaced by zeros contained in the function $1 - \lambda^2$. These zeros are located outside the real frequency axis, the complex frequency, which leads to a significant loss of selectivity, which can be increased by reducing the filter bandwidth as shown in Figure 1. Transmission zeros here refer to frequencies at which energy transfer from source to load is impossible.

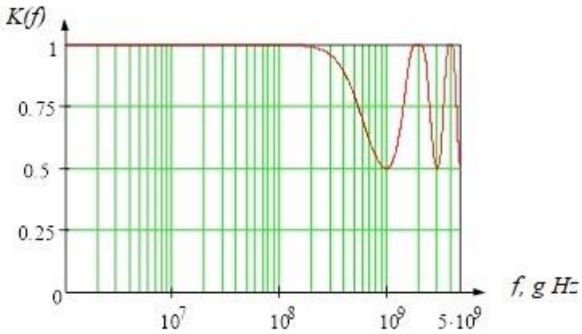


Figure 1. Frequency response of a second order butter worth filter

To increase selectivity and increase the transparency bandwidth, it is proposed to modify function 5 by introducing transmission zeros at the origin or at infinity (6) as follows:

$$\mathrm{K}(\lambda^2) = \frac{1}{1+\lambda^{2m}\left((-1)^n \frac{\alpha^2 (i\,\lambda)^2}{1-(i\,\lambda)^2}\right)^n} \quad (6)$$

where -$|m - n| = 1,\ m \geq 2$. The modification does not change the main principle of the synthesis of microwave circuits; the structure of the filter is known in advance. Implementation of λ^{2m} is possible with short-circuited (2) or open-circuited (3) sections of transmission lines. Due to the fact that in practice it is difficult to implement these segments in a serial branch, we will limit ourselves to their location in a parallel one. Another important limitation also follows from (6). Since this expression is a function of one variable, the length of the synthesized segments must have the same length *l*. The form of function (6) involves sequentially extracting from it the resistance of a short-circuited (open loop) in the parallel branch and then a segment of the transmission line in the serial branch. This determines the need to fulfill the condition. However, this condition may not be met. If $m - n = 1 + k$, then as a result of synthesis k times the short-circuited and open-circuited loops will be selected in a row. To increase the variability of the transfer functions of microwave filters, a further modification of expression (7) is proposed; for this, we will consider in detail function (8), as a result of which two segments of the same length with different wave impedances are formed. Let's get rid of the square in, which will lead to a reduction in one segment and double the length of the remaining one. Taking $\lambda = tg(\beta l)$ into account, we get:

$$\left(\frac{(-1)^2 \alpha^2 [i\ tg(\beta l)]^2}{1-[i\ tg(\beta l)]^2}\right)^2 \rightarrow \frac{(-1)\ \alpha^2 [i\ tg(2\ \beta l)]^2}{1-[i\ tg(2\ \beta l)]^2} \quad (7)$$

Using the tangent double argument property (8)

$$\left(\frac{(-1)^2\ \alpha^2 [i\ tg(\beta l)]^2}{1-[i\ tg(\beta l)]^2}\right)^2 \rightarrow \frac{4\ \alpha^2 [tg(\beta l)]^2}{1+2\ [tg(\beta l)]^2+[tg(\beta l)]^4} \quad (8)$$

Substituting (9) into (6) and making the reverse substitution $tg(\beta l) \rightarrow \lambda$:

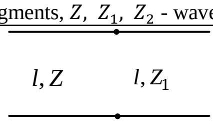
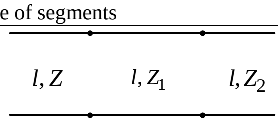
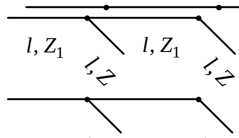
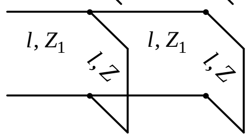
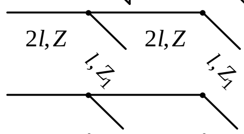
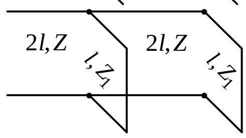
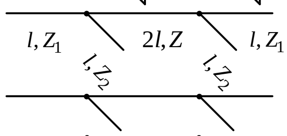
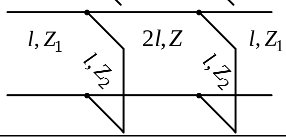
$$K(\lambda^2) = \frac{1}{1+\lambda^{2m}\left(\frac{4\ \alpha^2\lambda^2}{1+2\ \lambda^2+\lambda^4}\right)^n} \tag{9}$$

Segments of different lengths still remain a function of one variable. This is explained by the fact that a segment of double length is formed by sequentially extracting two segments having the same characteristic impedance. Maximum quality indicators of synthesized matching and filtering devices in the microwave range can be achieved using, as an approximating function, a function of the form (10).

$$K(\lambda^2) = \frac{1}{1+\lambda^{2m}\left(\frac{4\ \alpha^2\lambda^2}{1+2\ \lambda^2+\lambda^4}\right)^n\left((-1)^n\frac{\alpha^2(i\ \lambda)^2}{1-(i\ \lambda)^2}\right)^m} \tag{10}$$

Filters described by function (9) consist of loops in a parallel branch and segments of transmission lines in a serial branch with varying lengths. Table 1 outlines the basic structures of filters implemented through functions (5-10). The filter configurations encompass parallel loops and serial transmission line segments of different lengths, providing a comprehensive overview of their elementary structures as shown in Table 1.

Table 1. Transfer functions and the corresponding elementary structures of microwave filters

l – length of segments, 2*l* – double length of segments, $Z,\ Z_1,\ Z_2$ - wave resistance of segments					
1	$K(\lambda^2) = \frac{1}{1+\left(\frac{\alpha^2(i\ \lambda)^2}{1-(i\ \lambda)^2}\right)^n}$, borrowed from [4] for comparison	*n*=2		*n*=3	
2	$K(\lambda^2) = \frac{1}{1+\lambda^{2m}\left((-1)^n\frac{\alpha^2(i\ \lambda)^2}{1-(i\ \lambda)^2}\right)^n}$	*m*=2, *n*=2			
3	$K(\lambda^2) = \frac{1}{1+\frac{1}{\lambda^{2m}}\left((-1)^n\frac{\alpha^2(i\ \lambda)^2}{1-(i\ \lambda)^2}\right)^n}$	*m*=2, *n*=2			
4	$K(\lambda^2) = \frac{1}{1+\lambda^{2m}\left(\frac{4\ \alpha^2\lambda^2}{1+2\ \lambda^2+\lambda^4}\right)^n}$	*m*=2, *n*=2			
5	$K(\lambda^2) = \frac{1}{1+\frac{1}{\lambda^{2m}}\left(\frac{4\ \alpha^2\lambda^2}{1+2\ \lambda^2+\lambda^4}\right)^n}$	*m*=2, *n*=2			
6	$K(\lambda^2) = \frac{1}{1+\lambda^{2m}\left(\frac{4\ \alpha^2\lambda^2}{1+2\ \lambda^2+\lambda^4}\right)^n\left((-1)^n\frac{\alpha^2(i\ \lambda)^2}{1-(i\ \lambda)^2}\right)^m}$	*m*=2, *n*=1			
7	$K(\lambda^2) = \frac{1}{1+\frac{1}{\lambda^{2m}}\left(\frac{4\ \alpha^2\lambda^2}{1+2\ \lambda^2+\lambda^4}\right)^n\left((-1)^n\frac{\alpha^2(i\ \lambda)^2}{1-(i\ \lambda)^2}\right)^m}$	*m*=2, *n*=1			

Thus, the presented approximating functions allow you to initially set the frequency response and determine the structure of the filter. All presented functions have increased variable capabilities, which allows you to synthesize a wide variety of previously unused structures. Precise methods for synthesizing electrical circuits in classical theory involve extracting from $K(\lambda^2)$ functions $\rho(\lambda)$ which are related by the relation $K(\lambda^2) = 1 - \rho(-\lambda)\ \rho(\lambda)$. Taking into account (11):

$$\rho(-\lambda)\ \rho(\lambda) = \frac{q(\lambda)q(-\lambda)}{g(\lambda)g(-\lambda)} = \frac{\lambda^{2m}\left((-1)^n\frac{\alpha^2(i\ \lambda)^2}{1-(i\ \lambda)^2}\right)^n}{1+\lambda^{2m}\left((-1)^n\frac{\alpha^2(i\ \lambda)^2}{1-(i\ \lambda)^2}\right)^n} \tag{11}$$

From (11.1) where we determine $\rho(\lambda)$

$$\rho(\lambda) = \frac{q(\lambda)}{g(\lambda)} \tag{11.1}$$

The polynomial is equal to (λ) :

$$q(\lambda) = \sqrt{q_i}\prod_{i=1}^{n}[\lambda_i - (-\alpha_i \pm i\beta_i)] = \sqrt{q_i}(q_i\ \lambda^n + q_{i+1}\ \lambda^{n-1} + \ldots + q_n\ \lambda + q_{n+1})$$

and a polynomial $g(\lambda)$ defined as

$$g(\lambda) = \sqrt{g_i}\prod_{i=1}^{n}[\lambda_i - (-\alpha_i \pm i\beta_i)] = \sqrt{g_i}(g_i\ \lambda^n + g_{i+1}\ \lambda^{n-1} + \ldots + g_n\ \lambda + g_{n+1})$$

For functions (9), (10), factorization is performed in a similar way. After the reflection coefficient is selected, the resistance function is determined:

$$Z(\lambda) = \frac{Z_0 - \rho(\lambda)}{Z_0 + \rho(\lambda)}$$

where Z_0 - standard (normalized) resistance (1) equal to 50 Ohms. The final stage of the synthesis is to determine the values of the wave impedances of the transmission line segments. It is most convenient to use for this purpose the method of indefinite coefficients, described in detail in (2) and further developed in (1).

3. RESULTS AND DISCUSSION

As an illustrative example, we will calculate the filter using function (8) using the second-order Chebyshev correction polynomial (instead of λ^{2m} at m=2 polynomial applied $(2\lambda^2 - 1)^2$), $n = 1$, $m = 2$, $\alpha = 1$. With this in mind:

$$K(\lambda^2) = \frac{(\lambda^2 + 1)^4}{16\ \lambda^{10} - 15\ \lambda^8 + 8\ \lambda^6 + 6\ \lambda^4 + 4\ \lambda^2 + 1}$$

The result of factorization of the reflection coefficient is the expression:

$$\rho(\lambda) = \frac{\lambda^5 + 0.5\ \lambda}{\lambda^5 + 2.42\ \lambda^4 + 3.4\ \lambda^3 + 2.81\ \lambda^2 + 1.29\ \lambda + 0.25}$$

and the corresponding resistance function has the form:

$$Z(\lambda) = \frac{9.68\ \lambda^4 + 11.6\ \lambda^3 + 11.24\ \lambda^2 + 5.15\ \lambda + 1}{8\ \lambda^5 + 9.68\ \lambda^4 + 15.6\ \lambda^3 + 11.24\ \lambda^2 + 5.15\ \lambda + 1}$$

Determining wave impedances using the method of indefinite coefficients leads to the result $Z_1 = Z_5 = 1.315, Z_1 = Z_5 = 1.315, Z_3 = 1.261$. The corresponding filter structure is illustrated in Figure 2, while Figure 3 outlines its specific characteristics. The determined wave impedances provide essential insights into the filter's behavior and performance.

Modeling was carried out in CAD Advanced Design System 2016 for a substrate thickness of $h = 0.1$ mm and dielectric constant $\varepsilon = 1$. To obtain a low-frequency response with a cutoff frequency of $f_{gr} = 1$ GHz, the length of the segments l is equal to.

$$l = \frac{\left(\frac{\lambda}{\sqrt{\varepsilon_{ef}}}\right)}{8} = 37.5\ mm$$

where λ is free space wavelength. ε_{ef} is effective dielectric constant of the substrate material. The free space wavelength is a fundamental concept in electromagnetics, representing the distance a wave travels in a vacuum during one complete cycle. It is inversely proportional to the frequency of the wave. The effective dielectric constant of the substrate material plays a crucial role in determining the propagation speed of electromagnetic waves within a medium, influencing the wavelength and overall behavior of signals in that material. Understanding these parameters is essential in the design and analysis of various electronic and communication systems.

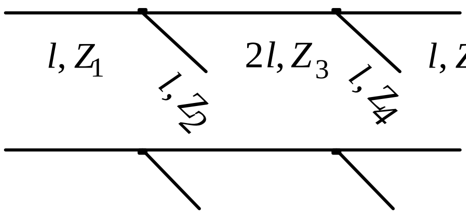


Figure 2. Filter structure

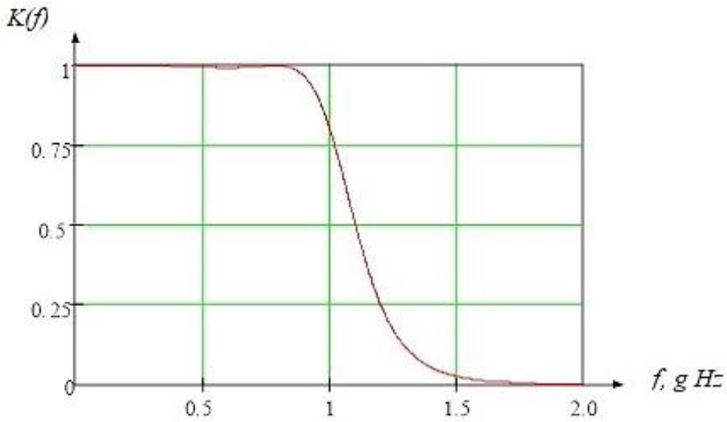


Figure 3. Frequency response of a fifth-order microwave filter

The resulting filter exhibits outstanding frequency response linearity in the transparency region, highlighting its ability to transmit different frequencies without distortion. This quality makes it attractive for use in communication and data transmission systems where signal transmission accuracy is important. High linearity ensures the filter's stability and reliability when working with a variety of frequency signals. To numerically evaluate the advantages of the developed mathematical apparatus, it seems appropriate to compare the results of the direct synthesis method and the method of synthesis of matching-filtering devices based on equivalent transformation. When designing devices in practice, the most widely used principle is based on the calculation of equivalent circuits using lumped elements and their subsequent replacement by elements with distributed parameters. With this approach, synthesis consists of several stages. At the first stage, a prototype is synthesized with a given transfer function, which is optimal for the required conditions. At the second stage, the elements are denormalized and the transition to the required type of the device being developed is carried out. At the third stage, the transition to elements with distributed parameters (11) is carried out.

$$K(-s^2) = \frac{1}{1+30^2 s^{16}(2s^2-1)^2} \tag{11.2}$$

The result of the implementation of the third stage of synthesis (taking into account the parameters of the Rogers dielectric material of the RO3003 series (PTFE) with thickness $H = 0.75$ mm, electrical constant ε = 3 and dielectric loss tangent tg = 0.0003) are alternating sections of transmission lines with high $Z_v = 130$ Ohm and low $Z_n = 13$ Ohm wave impedances. Taking this into account, as well as taking into account the values of the elements of the prototype with SP, the lengths of the transmission line segments were determined, which turned out to be equal: $l_2 = 2.26$ mm, $l_3 = 6.016$ mm, $l_4 = 3.846$ mm, $l_5 = 7.333$ mm, $l_6 = 3.825$ mm, $l_7 = 7.416$ mm, $l_8 = 3.372$ mm, $l_9 = 3.747$ mm, $l_{10} = 0.785$ mm, $W_0 = 1.901$ mm, $W_{\text{н}} = 10.772$ mm, $W_{\text{в}} = 0.216$ mm. Figure 4 shows a cross-sectional view of a microstrip line.

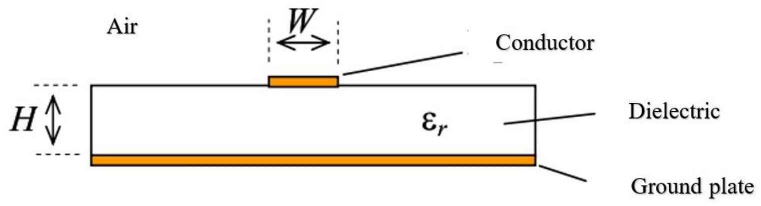


Figure 4. Cross-sectional view of a microstrip line

The use of exact synthesis methods to determine the values of the elements of matching circuits with distributed parameters implies the presence of the original transfer function in the form of a reflection coefficient $S_{11}(s)$. Due to the fact that the elements of the generalized scattering matrix are functions of the reflection and transmission coefficients of the matching-filtering devices, it seems appropriate to use the exact synthesis method to implement the function $S_{11}(s)$ using the expression (12):

$$K(-s^2) = \frac{1}{1+30^2 s^{16}(2s^2-1)^2} = 1 - S_{11}(s)S_{11}(-s) \tag{12}$$

By selecting poles and zeros in the left half-plane, the reflectance function taking into account the transformation (13):

$$s^2 = (-1)\ \alpha^2 \frac{\lambda^2}{1-\lambda^2} \tag{13}$$

$S_{11}(s)$ equal to (14):

$$S_{11}(\lambda) = \frac{\lambda^8 30 (\lambda^2+1)}{30\lambda^{10}+133\lambda^9+325\lambda^8+540\lambda^7+651\lambda^6+580\lambda^5+380\lambda^4+179\lambda^3+57\lambda^2+11\lambda+1}. \tag{14}$$

In Figure 5, the normalized characteristics of power conversion for circuits are depicted, calculated through both the method based on equivalent conversion and the exact mathematical method of direct synthesis utilizing an element of the generalized scattering matrix, denoted as $S_{11}(\lambda)$. The comparison between these two methods is visually presented, offering insights into the efficiency and accuracy of the power conversion calculations. This figure provides a comprehensive view of the performance of the circuits under the different mathematical approaches employed.

To determine a numerical assessment of the advantages of the developed mathematical apparatus, a comparison was made of the level of transmitted power by the devices shown in Figure 5 with the ideal frequency characteristics in the filtering band as a percentage. To do this, in the range from 0 to 1 relative units. (corresponds to half power level) criterion (15)-(17) applied:

$$A = \int_0^1 |\xi(f) - K_1(f)|\, df = 0.042 \tag{15}$$

$$B = \int_0^1 |\xi(f) - K_1(f)|\, df = 0.028 \tag{16}$$

$$(A - B) \cdot 100\% = 1.4\% \tag{17}$$

where A,B – magnitude of the integral error in the transparency band, A devices in Figure 5. The application of (10)–(12) made it possible to determine the numerical gain of the exact synthesis method based on the element of the generalized scattering matrix, which amounted to 1.4 percent. Particular attention should be paid to the fact that the use of an exact synthesis method based on an element of the generalized scattering matrix allows one to initially set the frequency response and determine the structure of the filter.

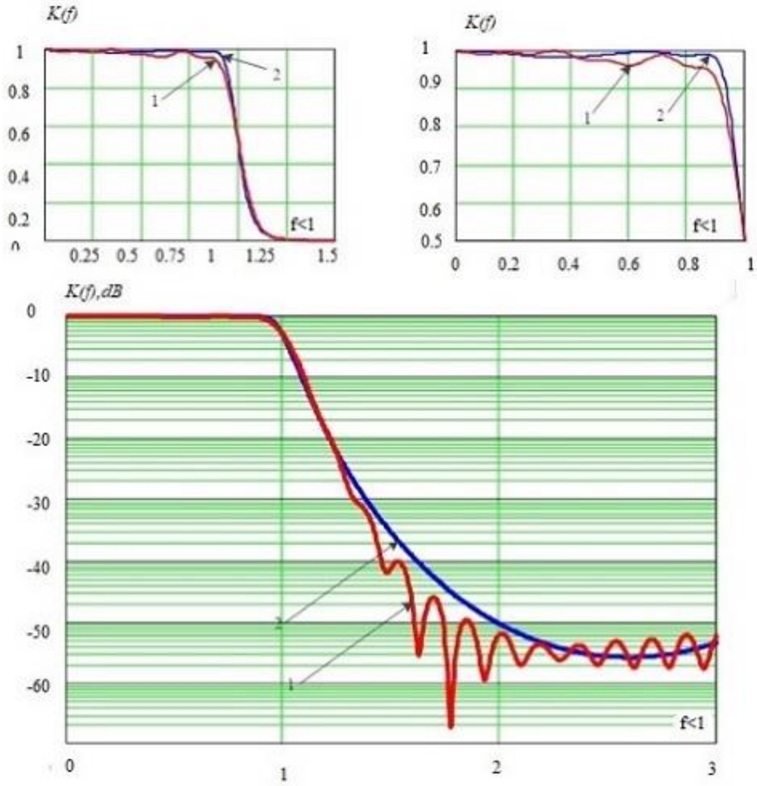


Figure 5. Frequency characteristics of matching-filtering devices: 1– equivalent transformation method 2– direct synthesis method based on an element of the generalized scattering matrix

4. CONCLUSION

In conclusion of the article, we can emphasize the importance of the developed method for the synthesis of microwave electrical filters, which introduces the approximation stage and thereby provides engineers and developers with significant advantages when designing radio devices. The new method not only provides an exact match to the required frequency response, but also provides the ability to determine the filter structure, which is a significant advantage over existing methods. An important advantage of the proposed method is the increased variability of the functions presented in expressions (6) and (8). This opens up wide opportunities for the synthesis of diverse and previously unused filter structures, which in turn helps to improve the quality characteristics of the synthesized devices. Table 1 shows only a limited set of potential variations, and developers are given the opportunity to further customize by changing the orders of various parts of the denominator polynomials, as well as the orders of the correction polynomials.

Particular attention should be paid to the applicability of the proposed functions (6) and (8) in broadband matching problems. Their unique structure, containing segments of various lengths and transmission zeros of various classes, provides additional opportunities for solving matching problems over a wide frequency range. This opens up new horizons for application in various fields of radio engineering, where efficient and accurate matching of microwave filters is required. Thus, the developed method not only expands the set of tools for designing microwave electrical filters, but also provides new perspectives for improving the performance and efficiency of radio engineering systems.

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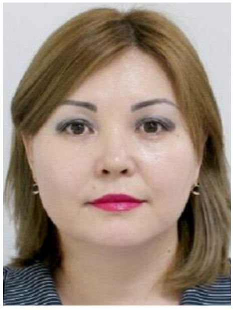

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