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Bandpass filter based on microstrip resonators with additional galvanic coupling

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An expression is obtained for the coupling coefficient of microstrip resonators with inductive, capacitive, and galvanic interaction caused by an additional inductive element. It is shown that the capacitive coupling coefficient has a sign opposite to the galvanic coupling coefficient, and the inductive coupling coefficient can have both a positive and negative sign. Theoretically and experimentally, using two-order microstrip filters with the same conductor topology, the possibility of significantly increasing the bandwidth due to the introduction of galvanic coupling, as well as mutual compensation of three types of couplings, leading to the disappearance of the first passband, has been demonstrated. A sixth-order bandpass filter based on quarter-wave microstrip resonators with additional galvanic coupling has high frequency-selective properties due to the appearance of two transmission zeros on either side of the passband. The developed filter is made on a substrate with a permittivity $\varepsilon = 80$, thickness 2 mm and dimensions 25×50 mm. The central frequency of the filter passband is $f_0 = 0.5$ GHz, and its fractional bandwidth $\Delta f/f_0 = 15\%$.

Keywords: dielectric substrate, microstrip resonator, coupling coefficients, oscillation mode.

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Bandpass microwave filters of order N are normally constructed on the basis of N series-connected resonators. Additional couplings between non-adjacent resonators, which induce the formation of transmission zeros on either side of the passband, are usually used to improve the frequency-selective characteristics of such filters without changing the number of resonators [1–3]. This approach helps improve the characteristics of devices with microstrip [2], coaxial [4,5], and waveguide [3,6] resonators and, naturally, of filters based on quasi-lumped inductive (L) and capacitive (C) elements [7]. Symmetric positioning of transmission zeros to the left and to the right of the passband, which provides the steepest slopes of the frequency response (FR), is achieved by introducing additional capacitive coupling either between the input and the output in a fourth-order microstrip filter [1] or between the second and the fifth resonators in a sixth-order filter [1,2]. In a twelfth-order waveguide filter [3], four additional couplings are established: two of them are capacitive (between pairs of non-adjacent resonators 4–9 and 3–10), and the other two are inductive (between resonators 5–8 and 2–11).

It is evident that either capacitive or inductive additional couplings between non-adjacent resonators are used to improve the frequency-selective characteristics of filters of all the designs examined above. However, microstrip bandpass filters, where adjacent resonators are coupled electromagnetically to each other, allow one to establish additional galvanic coupling with an inductive element even between adjacent resonators. This inductive element connects certain points on the conductors of adjacent resonators. The present study is focused on the examination of additional galvanic coupling in quarter-wave microstrip resonators and a sixth-order filter based on them.

The coefficients of capacitive (k_C), inductive (k_L), and total coupling (K) are commonly used to characterize quantitatively the interaction of a pair of microstrip resonators (MSRs). These coefficients for identical MSRs with the maximum coupling of conductors near resonant frequencies are determined from the per-unit-length characteristics of interacting microstrip lines that form resonators [8]:

$$k_L = \frac{L_{12}}{L}, \qquad k_C = \frac{C_{12}}{C + C_{12}}, \qquad K = \frac{k_L - k_C}{1 - k_L k_C}, \quad (1)$$

where L and C are the inductance and capacitance per unit length of an individual microstrip line, and L_{12} and C_{12} are the per-unit-length mutual inductance and capacitance of coupled regular microstrip lines. It is important to note that coefficients k_C and k_L are equal in this case at relative substrate permittivity $\varepsilon = 1$, and inequality $k_L > k_C$ is always fulfilled at $\varepsilon > 1$ [8].

It is common knowledge that a pair of electromagnetically coupled microstrip resonators at the frequencies of the first passband is characterized by an equivalent network of two coupled oscillatory circuits. It is also known that the fractional bandwidth of two-section filters constructed based on any electrodynamic resonators (as well as on circuits) is proportional to the total coupling coefficient, which may be expressed in terms of the natural frequencies of even (f_e) and odd (f_o) modes of coupled oscillations [8–10]:

$$K = \frac{f_e^2 - f_o^2}{f_e^2 + f_o^2}. \quad (2)$$

The known expressions for frequencies f_e, f_o and the coupling coefficients of oscillatory circuits with inductive (*1*), capacitive (*2*), and inductive-capacitive (*3*) interaction are presented in the table for comparison. The obtained expressions for circuits with additional galvanic coupling, which is established by means of inductance G_{12} (*4*), and galvanic-capacitive (*5*), galvanic-inductive (*6*), and mixed (*7*) couplings are also listed there. Note that the coefficient of total coupling of circuits with both inductive and capacitive interaction assumes different values in the cases of co-directional and counter-directional connection of inductive elements and is also affected by the introduction of additional galvanic coupling in the circuit. It is always possible to achieve complete compensation of all interactions at the resonant frequency of the circuits in such cases ($K = 0$). A transmission zero is then observed at the position of the filter passband in such circuits. In microstrip two-section stages, this effect is achieved if regular strip conductors are made irregular [11] to reduce inductive coupling coefficient k_L to the level of capacitive coefficient k_C (1). Note that compensation of the total coupling of microstrip resonators at the frequency of the first oscillation mode is used in devices for protection against high-power radio pulses based on a high-temperature superconductor film [12] and in various types of sensor devices [13,14].

When additional galvanic interaction between circuits is introduced, the coefficient of total coupling may be equal to zero in the cases when only galvanic-capacitive, only galvanic-inductive, and, naturally, mixed coupling is present. This follows from the approximate formula corresponding to small magnitudes of inductive, capacitive, and galvanic interaction of the circuits when quantities of the second order of smallness may be neglected in the formula for the total coupling coefficient:

$$K \approx k_G \pm k_L - k_C. \quad (3)$$

The „plus" and „minus" signs correspond to co-directional and counter-directional connection of inductances in the circuits, respectively. It is evident from formula (3) that galvanic k_G and capacitive k_C coupling coefficients have opposite signs, while coefficient of inductive coupling k_L changes its sign depending on the type of inductance connection.

Four kinds of two-section microstrip structures based on quarter-wave resonators, which differed in the type of MSR coupling, were fabricated to verify the discovered pattern. For objective comparison, all microstrip structures were manufactured on substrates made of high-frequency TBNS ceramics with a thickness of 2 mm and relative permittivity $\varepsilon = 80$. The resonators had the same length $l_R = 18$ mm and width $w_R = 6$ mm of strip conductors and the same size of gaps between conductors $s_R = 1$ mm. The first two traditional designs based on co-directional and counter-directional resonators were tuned as bandpass filters by selecting the points of conductive connection of transmission lines with a wave impedance of 50 Ω to strip conductors so that reflection level $S_{11}(f)$ in the passband did not exceed −20 dB. The fractional bandwidth of filters measured at 3 dB from minimum losses was $\Delta f/f_0 = 19\%$ ($K \approx k_L - k_C = 0.134$) and $\Delta f/f_0 = 24\%$ ($K \approx -k_L - k_C = -0.170$). The corresponding points of connection of SMA connectors to the strip conductors were located at a distance of 7.65 mm (the first design) and 5.65 mm (the second design) from their free ends.

The third design based on co-directional resonators (Fig. 1, *a*) was also tuned as a bandpass filter by moving the conductor of additional galvanic coupling by a distance of 3.35 mm from the ends of the strip conductors connected to the shield, but with the transmission lines connected to the free ends of MSR conductors. In the fourth design (Fig. 1, *b*), all three coupling coefficients (k_G, k_L, and k_C) compensate each other, and a transmission zero is thus observed instead of a passband in the frequency response. However, an enhancement or suppression of the additional galvanic coupling of resonators leads in this case to the emergence of a passband with a transmission zero to the left or right of it, which makes corresponding slope of the frequency response significantly steeper. Note that the central frequency of passbands of all filters falls within the range of $f_0 = 510-550$ MHz, and the width of conductors of additional galvanic coupling is $s_G = 0.3$ mm.

Figure 1 shows not only the frequency response of the devices, but also the topologies of conductors of the studied structures with galvanic coupling and photographic images of samples fabricated according to the design parameters obtained by parametric synthesis of 3D models in the CST Studio Suite electrodynamic analysis package. Solid curves (Fig. 1) denote the frequency response of lumped-element equivalent networks *6* and *7* (see the table) connected to transmission lines with a wave impedance of 50 Ω. The values of circuit elements were determined using one-dimensional models of coupled microstrip lines with their per-unit-length parameters calculated in a quasi-static approximation. The values of elements of equivalent networks obtained this way ($C = 23.3$ pF, $L = 4.06$ nH, $C_{12} = 0.427$ pF, $L_{12} = 0.617$ nH, and $G_{12} = 22.7$ nH for co-directional and 23.4 nH for counter-directional resonators) provide a qualitative fit between the calculated and measured FRs (Fig. 1). Since the FRs calculated in the CST Studio Suite package for electrodynamic analysis of 3D models match the measurement results well, they are not shown so as to avoid overloading the figure. It is important to note that, in contrast with L_{12}, an increase in inductance of additional galvanic coupling leads to the suppression of interaction of the circuits, and G_{12} is thus more than an order of magnitude higher than mutual inductance L_{12}.

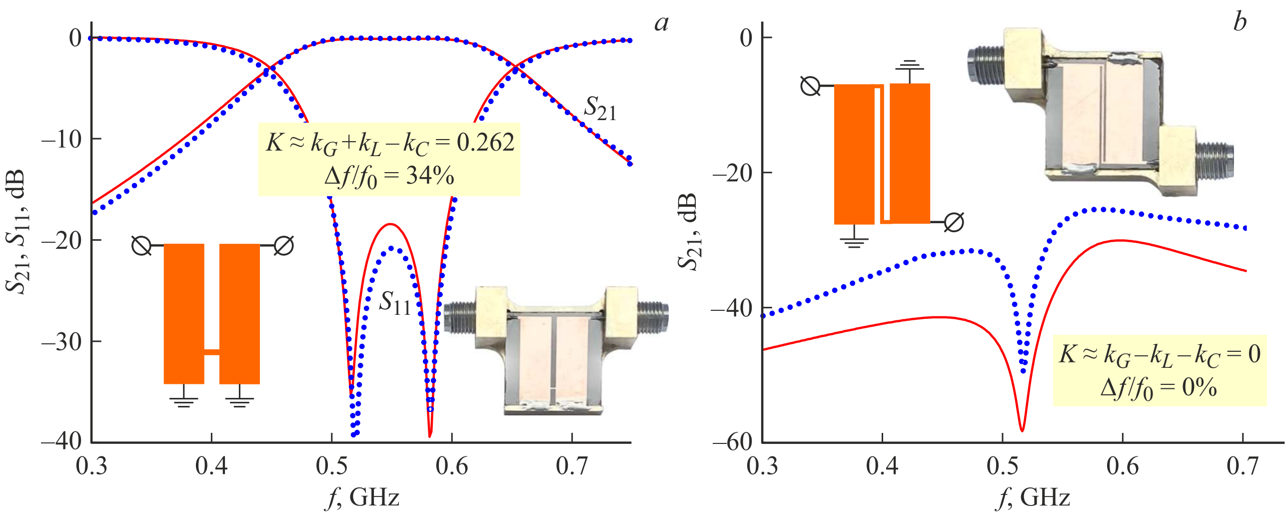
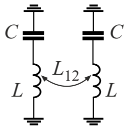
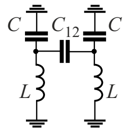
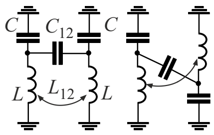
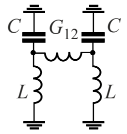
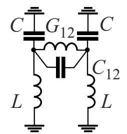
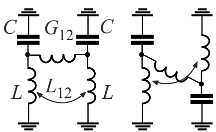
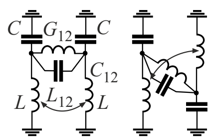


Figure 1. Frequency response of the studied microstrip two-resonator structures. Solid and doted curves represent the results of calculations for equivalent circuits *6* and *7* (see the table) and the measurement data, respectively. The conductor topologies and photographic images of fabricated samples are shown in the insets.

Natural frequencies and coupling coefficients of two circuits with different types of interaction

Coupling type		Frequencies of oscillation modes and coupling coefficients
1.		$f_e = \frac{1}{2\pi}\frac{1}{\sqrt{(L+L_{12})C}}$, $f_o = \frac{1}{2\pi}\frac{1}{\sqrt{(L-L_{12})C}}$, $K = k_L = \frac{L_{12}}{L}$
2.		$f_e = \frac{1}{2\pi}\frac{1}{\sqrt{LC}}$, $f_o = \frac{1}{2\pi}\frac{1}{\sqrt{L(C+2C_{12})}}$, $K = k_C = \frac{C_{12}}{C+C_{12}}$
3.		$f_e = \frac{1}{2\pi}\frac{1}{\sqrt{C(L+L_{12})}}$, $f_o = \frac{1}{2\pi}\sqrt{\frac{L_{12}+2(L-L_{12})}{L_{12}(L-L_{12})(C+2C_{12})}}$, $K = \frac{\pm k_L - k_C}{1 \mp k_L k_C}$
4.		$f_e = \frac{1}{2\pi}\frac{1}{\sqrt{LC}}$, $f_o = \frac{1}{2\pi}\sqrt{\frac{G_{12}+2L}{G_{12}CL}}$, $K = k_G = \frac{L}{L+G_{12}}$
5.		$f_e = \frac{1}{2\pi}\frac{1}{\sqrt{LC}}$, $f_o = \frac{1}{2\pi}\sqrt{\frac{G_{12}+2L}{G_{12}L(C+2C_{12})}}$, $K = \frac{k_G - k_C}{1 - k_G k_C}$
6.		$f_e = \frac{1}{2\pi}\frac{1}{\sqrt{C(L\pm L_{12})}}$, $f_o = \frac{1}{2\pi}\sqrt{\frac{G_{12}+2(L\mp L_{12})}{G_{12}(L\mp L_{12})C}}$, $K = \frac{k_G \pm k_L \mp k_G k_L - k_G k_L^2}{1 - k_G k_L^2}$
7.		$f_e = \frac{1}{2\pi}\frac{1}{\sqrt{C(L\pm L_{12})}}$, $f_o = \frac{1}{2\pi}\sqrt{\frac{G_{12}+2(L\mp L_{12})}{G_{12}(L\mp L_{12})(C+2C_{12})}}$, $K = \frac{k_G \pm k_L - k_C \mp k_G k_L - k_G k_L^2 + k_G k_L^2 k_C}{1 - k_G k_C \mp k_L k_C - k_G k_L^2 \pm k_G k_L k_C + k_G k_L^2 k_C}$

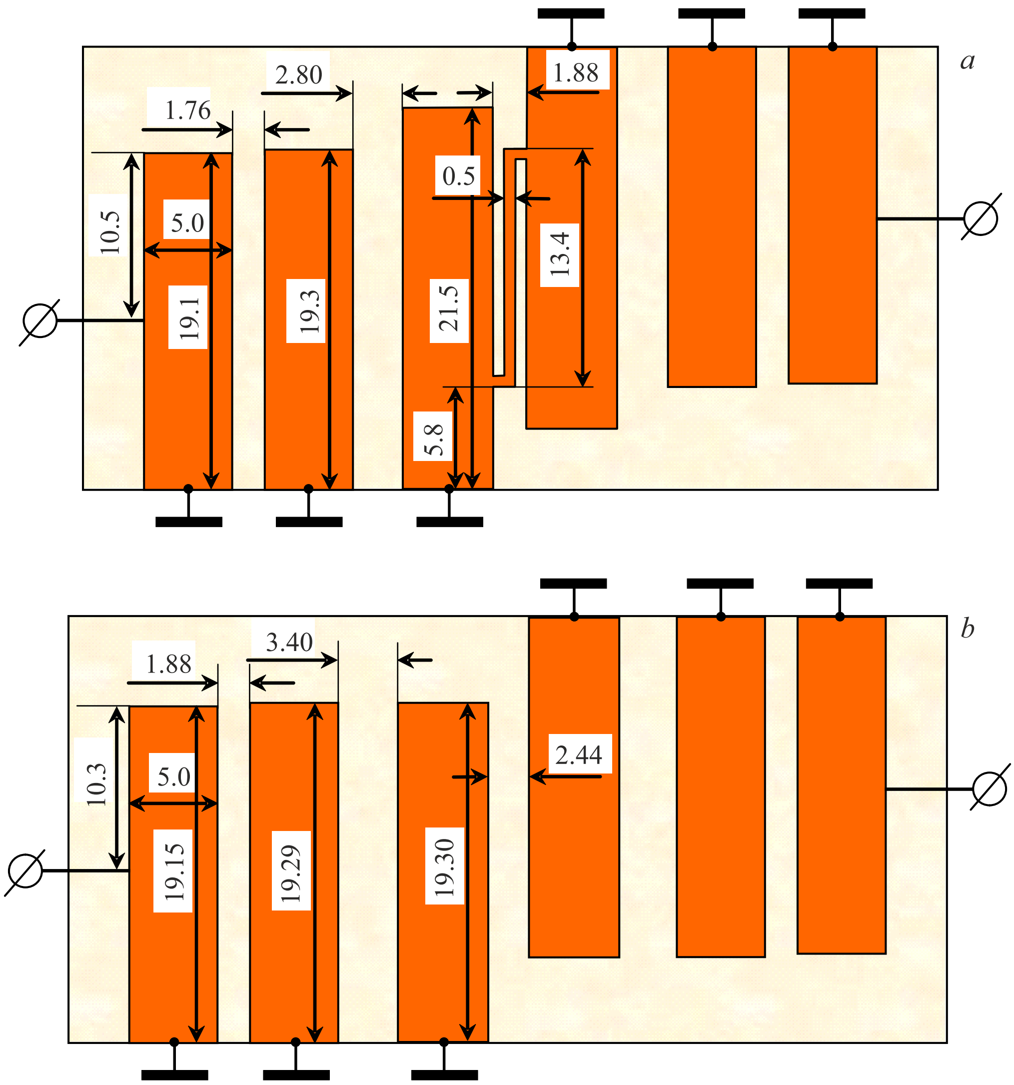


Figure 2. Topology of conductors of a sixth-order filter based on quarter-wave microstrip resonators with additional galvanic coupling between resonators 3 and 4 (*a*) and without additional coupling (*b*). All dimensions are given in millimeters.

A sixth-order filter was synthesized and fabricated in order to evaluate the selectivity of filters with additional galvanic coupling between resonators. Its conductor topology and dimensions are presented in Fig. 2, *a*. Additional galvanic coupling in it is established between resonators 3 and 4 only. The filter was tuned to central frequency $f_0 = 500$ MHz of the passband and its fractional bandwidth $\Delta f/f_0 = 15\%$. The minimum loss in the passband of the device is 1 dB. The filter design includes a top cover positioned at a height of 20 mm from the substrate surface. A similar filter without additional coupling was designed and fabricated for comparison. Its conductor topology and dimensions are shown in Fig. 2, *b*. The frequency responses of the examined filters are presented in Fig. 3. It is evident that the introduction of even a single additional galvanic coupling between the central resonators triggers the emergence of transmission zeros near the passband, which enhance significantly the frequency-selective characteristics of a filter.

Thus, the introduction of additional galvanic interaction between adjacent resonators of a microstrip structure allows one, first, to enhance significantly the frequency selectivity of multisection filters due to the formation of transmission zeros on either side of the passband. Second, such galvanic interaction in two-section stages provides an opportunity to balance out all coupling coefficients at the frequencies of the first passband even in a structure based on regular microstrip resonators. This enables the fabrication of various highly sensitive sensors based on „compensated" microstrip structures with ferroelectric or magnetic films and devices with high-temperature superconductor films used for protection against high-power radio pulses.

Third, the use of galvanic coupling in microstrip structures allows for a significant increase in the total coupling

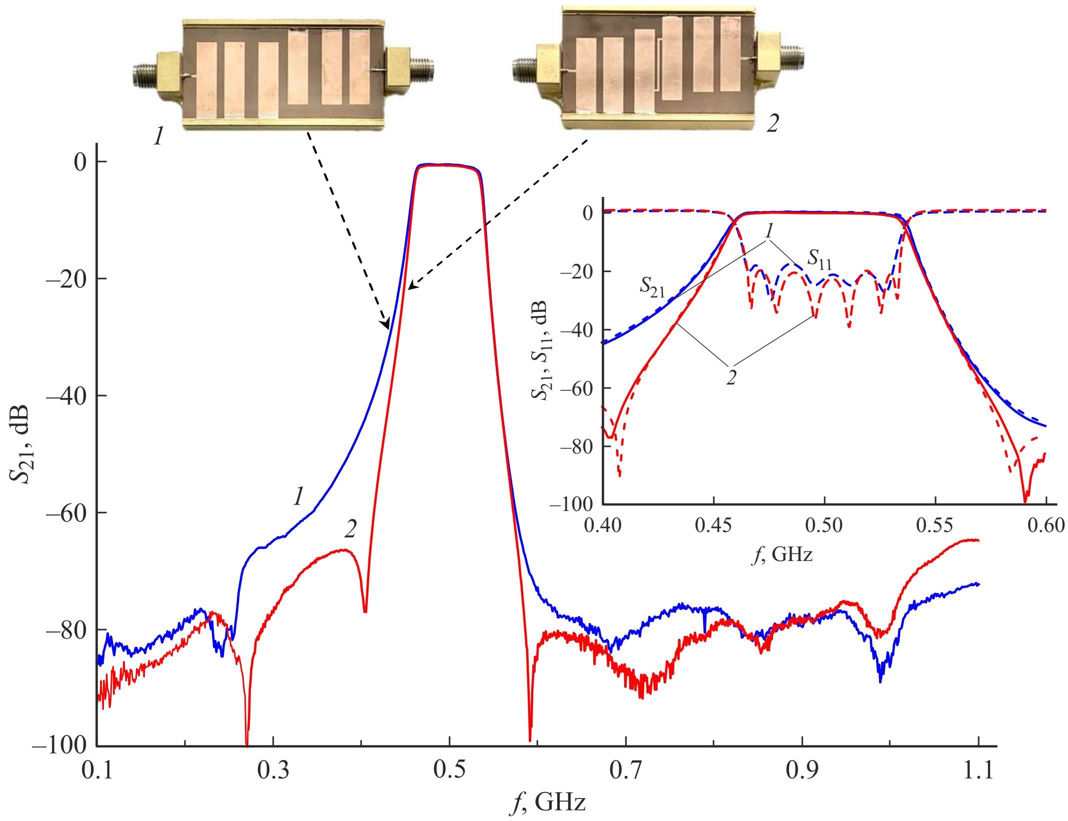


Figure 3. Characteristics of prototype filters in wide and narrow frequency ranges. Solid curves correspond to the measurement data, while dashed and dotted lines represent the results of calculations. The insets show photographic images of the devices. The numbers next to curves correspond to the numbers of devices in the insets.

coefficient of resonators and, consequently, an expansion of the passband of filters based on them. This provides an opportunity to construct small-sized wideband and ultra-wideband bandpass microstrip filters.

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Conflict of interest

The authors declare that they have no conflict of interest.

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