Development of CAD Tool for a Design of Microwave Planar HTS Filters

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***Abstract*—Accurate design of microwave planar high-temperature superconducting (HTS) filters is obtained using a computer-aided design tool, which provides simulation of characteristics of the multicoupled transmission-line structures. The quasi-static spectral-domain method is applied. In order to achieve a good correspondence between measured and simulated filter performance, correct models of the microwave propagation characteristics of HTS planar transmission lines have to be used in the simulation procedure. The goal of this paper is to demonstrate that it is possible to design the planar HTS filters that do not need additional trimming after manufacturing. The trimmingless filters can be used in mass production for the next generation of mobile communication systems.**

***Index Terms*—Microwave filter, modeling, superconductors.**

I. Introduction

HIGH-QUALITY microwave filters play an extremely important role in designing modern and future communication systems, such as TV satellite broadcasting, cellular communications, and different navigation systems of a global range of operation. For some applications, the requirement of low insertion loss is of crucial importance. In this case, the use of high-temperature superconductors (HTS's) is an attractive alternative. The application of HTS technology also provides a higher selectivity of the filters and increased sensitivity of receiving systems. The increased sensitivity allows reducing the radiated power; the increased selectivity provides a decrease of out-of-band interference. There are many publications concerning design of high-performance HTS filters of [1]–[11]. Planar structures of HTS filters based on microstrip or coplanar transmission lines look very attractive because of small size and weight. The planar filters exhibit extremely low in-band insertion loss and very steep slope at the edges of the passband [1], [7]–[9], [11]. It should be mentioned that successful experiment with microwave HTS filters have been presented in framework of the High-Temperature Space Experiment-II (HTSSE-II) [12], [13] and in the Superconducting Communication System Project (SUCOMS) [11]. Thus, one may say that the HTS application in microwave engineering becomes a considerable part of the industrial production.

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Investigations of HTS thin-film physical properties and deriving correct phenomenological description of the film characteristics have become an important problem of modern electronics. The next step of the problem formulated above is development of computer-aided design (CAD)-oriented software for designing HTS filters, to provide the required characteristics without using any trimming procedure. In order to develop a CAD tool providing a correct simulation of the HTS filter characteristics, the following problems should be solved:

1) developing a reliable model of HTS microwave transmission lines, which is based on the correct description of the HTS film properties and the dielectric characteristics of the substrate;
2) developing a design procedure of microwave filters of higher order using a set of coupled resonators;
3) investigation of the sensitivity of the filter characteristics to the HTS film and the substrate parameter dispersion.

It should be noted that the dependence of film and substrate parameters on temperature has to be taken into account.

The goal of this paper is to show how the planar filter should be designed with a desired accuracy and to demonstrate an evidence of a real possibility of designing trimmingless planar HTS filters.

II. Design of Bandpass HTS Filters

There are three main constituents in the design of microwave HTS filters:

- accurate model of HTS film surface impedance;
- efficient CAD tool providing a correct simulation of planar HTS microwave structures;
- complementarity of the simulation and measurement data.

Many planar devices are designed using the same base of a simulation: resonators, filters, multiplexers etc. In this paper, we present the results of simulation and design of planar HTS filters based on microstrips and coplanar waveguides.

A. Model of HTS Surface Impedance

The model of HTS surface impedance is based on the two-fluid approach using the physical characteristics of superconducting materials: transition temperature T_c, normal conductivity σ_n, and the penetration depth of magnetic field, i.e., London penetration depth λ_L [14]–[16]. The last two parameters σ_n and λ_L are temperature dependent. The surface resistance of thin HTS films is described by the following phenomenological model [14]:

$$R_{\rm sur}(t) = \begin{cases} \dfrac{1}{\sigma_n(t) \cdot d}, & \text{for } t \geq 1 \\ \dfrac{(\omega\mu_0)^2 \sigma_n(t)}{1 + [\omega\mu_0\sigma_n(t)\lambda_L^2(t)]} \cdot \dfrac{\lambda_L^4(t)}{d}, & \text{for } t < 1 \end{cases} \tag{1}$$

where

$$\sigma_n(t) = \sigma_n(1) \begin{cases} t^{-1}, & \text{for } t \geq 1 \\ t^{\gamma-1} + \alpha \cdot (1 - t^\gamma), & \text{for } t < 1 \end{cases} \tag{2}$$

$$\left[\lambda_L(0)/\lambda_L(t)\right]^2 = 1 - t^\gamma \tag{3}$$

$$t = T/T_c. \tag{4}$$

Here, d is the thickness of the HTS film, $\omega = 2\pi f$, f is the frequency, μ_0 is the magnetic permeability of free space, $\sigma_n(1)$ is the conductivity of the normal charge carriers at the transition temperature T_c, $\lambda_L(0)$ is the London penetration depth at $T = 0$, α is the residual resistance parameter, and γ is the empirical exponent.

The surface reactance of the thin HTS film can be calculated as follows [14]:

$$X_{\rm sur}(t) = \begin{cases} 0, & \text{for } t \geq 1 \\ \omega\mu_0 \cdot \dfrac{\lambda_L^2(t)}{d}, & \text{for } t < 1. \end{cases} \tag{5}$$

Equations (1) and (5) are valid for thin-film samples with the thickness $d \leq 2\lambda_L(t)$. If this condition is not fulfilled, one has to use the model of bulk HTS material ($t < 1$)

$$R_{\rm sur}(t) = \frac{(\omega\mu_0)^2 \sigma_n(t)}{1 + \left[\omega\mu_0\sigma_n(t)\lambda_L^2(t)\right]^{3/2}} \cdot \frac{\lambda_L^3(t)}{2} \tag{6}$$

$$X_{\rm sur}(t) = \sigma\mu_0 \cdot \lambda_L(t). \tag{7}$$

Five fitting parameters are used: T_c, $\sigma_n(1)$ at $T = T_c$, the London penetration depth $\lambda_L(0)$ at $T = 0$, exponent γ, and the residual resistance parameter α. These parameters should be extracted from the experimental temperature dependence of the HTS film surface impedance.

From (5) and (7), one can find the kinetic inductance of the film in the superconducting state

$$L_k(t) = \mu_0 \cdot \frac{\lambda_L^2(t)}{d} \tag{8}$$

for a thin HTS film, and

$$L_k(t) = \mu_0 \cdot \lambda_L(t) \tag{9}$$

for a bulk material.

The temperature dependence of the kinetic inductance is conditioned by the temperature dependence of the London penetration depth (3) and is determined by the model parameter γ. The parameter $\gamma = 1.5$–2.5 depends on the HTS film quality: the higher the film quality, the higher is γ, and the less is the microwave surface resistance of the film [15], [16]. The exponent γ is responsible for the slope of the temperature dependence of the surface resistance in the vicinity of the transition temperature. An empirical equation for the London penetration depth at $T = 0$ was found as a function of the model parameter γ [15]

$$\lambda_L(0) = 0.13 \cdot \exp(1.27 - 0.5\gamma) \quad (\mu\text{m}). \tag{10}$$

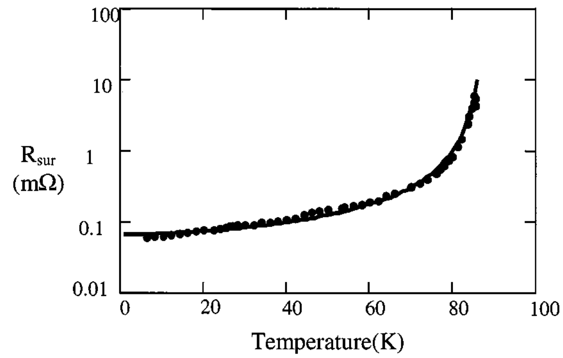


Fig. 1. Temperature dependence of YBCO film surface impedance: experiment [17] (filled circles) and simulated result (solid line)

In this case, the model of the surface impedance uses only four fitting parameters.

The validity of the model was verified using numerous experimental data. Fig. 1 shows an example of measured temperature dependence of the surface impedance of an unpatterned YBCO film [17]. The model description of the temperature dependence of the surface resistance of the film of thickness $d = 0.25$ μm at the frequency $f = 9.5$ GHz uses the following fitting parameters: $T_c = 88.2$ K, $\sigma_n(1) = 2.7 \cdot 10^6$ $(\Omega \cdot \text{m})^{-1}$, $\alpha = 3.1$, $\gamma = 1.95$. In accordance with (10), the London penetration depth at $T = 0$ is $\lambda_L(0) = 0.175$ μm. The modeled characteristic describes the experimental results with a high accuracy.

B. Modeling HTS Resonator Based on a Transmission-Line Section

In order to take into consideration the contribution of the edges of the patterned HTS film to the resistive characteristics of the film, the surface impedance should be measured on patterned HTS films. A half-wavelength resonator based on an HTS transmission-line section can be used for such measurements. It is necessary to describe scattering parameters of the HTS resonator, taking into account propagation characteristics of the HTS planar transmission lines: the phase velocity and attenuation coefficient. The phase velocity is influenced by the contribution of the kinetic inductance. The attenuation coefficient is determined mainly by the surface resistance of the superconducting film. All these characteristics are temperature dependent. Propagation characteristics of the HTS planar transmission lines have been presented in [14] and [16] and [18]–[20]. The nonhomogeneous current density distribution in the line cross section is taken into account.

The HTS film parameters are to be extracted from the experimental characteristics of the resonator (resonant frequency and

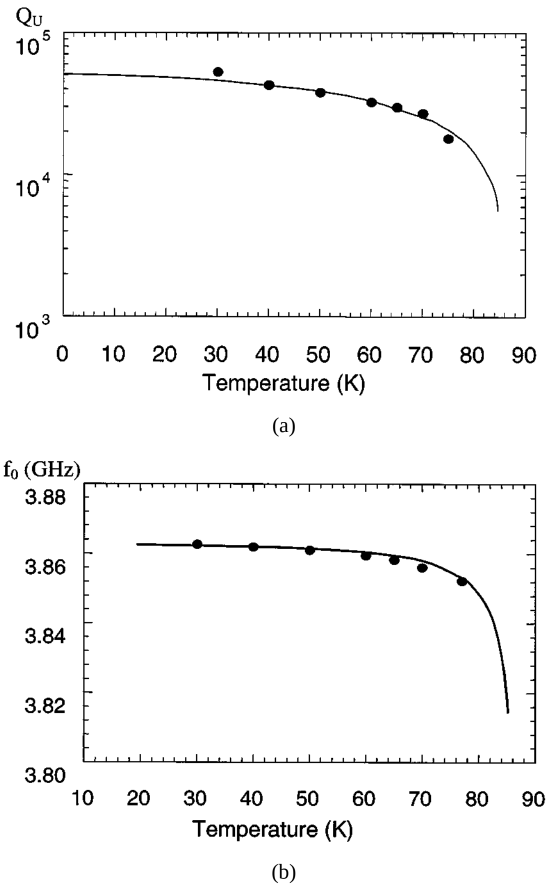


Fig. 2. (a) Temperature dependence of the unloaded Q-factor. (b) Resonant frequency of the half-wavelength YBCO microstrip resonator [experiment (filled circles) and the simulation results (solid lines)].

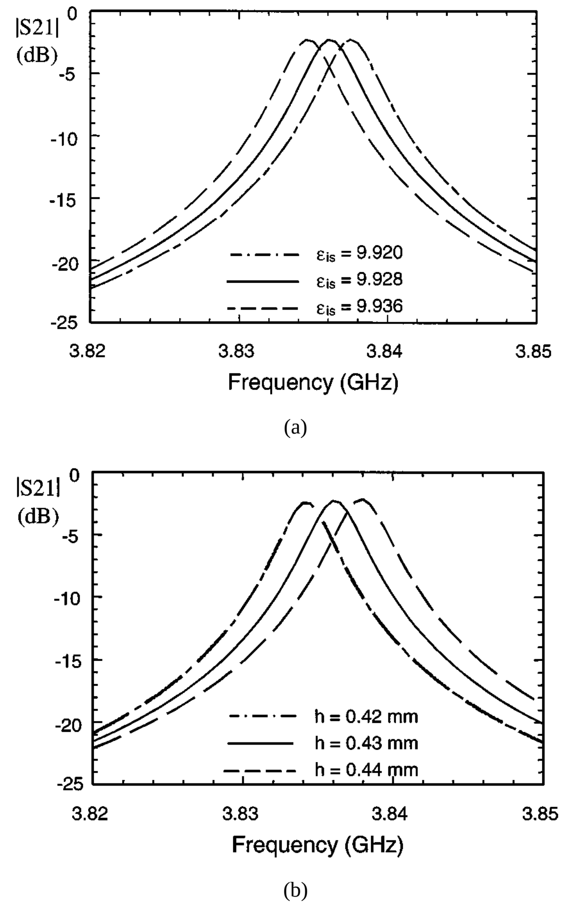


Fig. 3. Simulated characteristics of the YBCO microstrip resonator at $T = 65$ K for: (a) different dielectric constant of the substrate and (b) different substrate thickness.

unloaded Q-factor) obtained at different temperatures. Fig. 2 shows the experimental characteristics of the YBCO microstrip half-wavelength resonator on sapphire substrate. The resonator of 8-mm length and 0.4-mm width with two capacitive coupling gaps of 0.1-mm length was manufactured on 0.43-mm-thick sapphire substrate (r-cut). The angle between the wave propagation direction and the projection of the C-axis of sapphire on the substrate surface is 45°. The YBCO film thickness is $d = 0.33$ μm. The following fitting parameters of the surface impedance model were found: $T_c = 88$ K, $\sigma_n(1) = 2.7 \cdot 10^6$ $(\Omega{\cdot}\mathrm{m}^{-1})$, $\gamma = 1.9$, and $\alpha = 8$. The simulated characteristics are in a good agreement with the experimental data. The simulation procedure was based on using the equivalent isotropized dielectric constant of the anisotropic sapphire substrate [18]–[20]. In this case, $\varepsilon_{is} = 9.93$.

Having the precise model description of the HTS resonator, one can analyze how the resonance characteristic is influenced by a variation of parameters of the HTS film and dielectric substrate. Modeling the characteristic of the same microstrip resonator at $T = 65$ K revealed that the resonant frequency depends strongly on the dielectric constant of the substrate and its thickness (Fig. 3), and of the HTS film model parameter γ [see Fig. 4(a)]. The insertion loss at the resonant frequency is also remarkably dependent on the parameter γ. The HTS film thickness is also an essential parameter, which determines the resonant frequency and the insertion loss level [see Fig. 4(b)]. The residual resistance parameter $\alpha = 1$–20 and the normal conductivity $\sigma_n(1) = (1$–$3) \cdot 10^6$ $(\Omega \cdot \mathrm{m})^{-1}$ contribute in the insertion loss: the higher α and $\sigma_n(1)$, the greater the insertion loss at the resonant frequency. A reasonable variation of the transition temperature ($T_c = 87$–91 K) results in a negligibly small change of the resonant characteristic.

C. Modeling HTS Multicoupled Transmission Lines

As a rule, the bandpass filter structure is based on single or cascaded sections of multicoupled planar transmission lines (MCPL's). Thus, the problem of the filter structure analysis is reduced to an accurate estimation of the multiport impedance matrix of the coupled line system.

It is well known that the electrodynamic hybrid-mode analysis being applied to the task of such type is *a priori* more reliable compared with the quasi-TEM approach. On the other hand, the computational effectiveness of the quasi-TEM analysis is 2–3 orders higher than the electrodynamic simulation. Taking into consideration that the frequency used for wireless communication ($f = 1$-2 GHz) is not very high, one can suggest that the quasi-TEM approach does not introduce a significant error in the MCPL description. Thus, the choice of the quasi-TEM analysis for modeling MCPL is reasonable.

For a general planar geometry, the spectral-domain approach (SDA) is probably the most simple and widely used tool. In

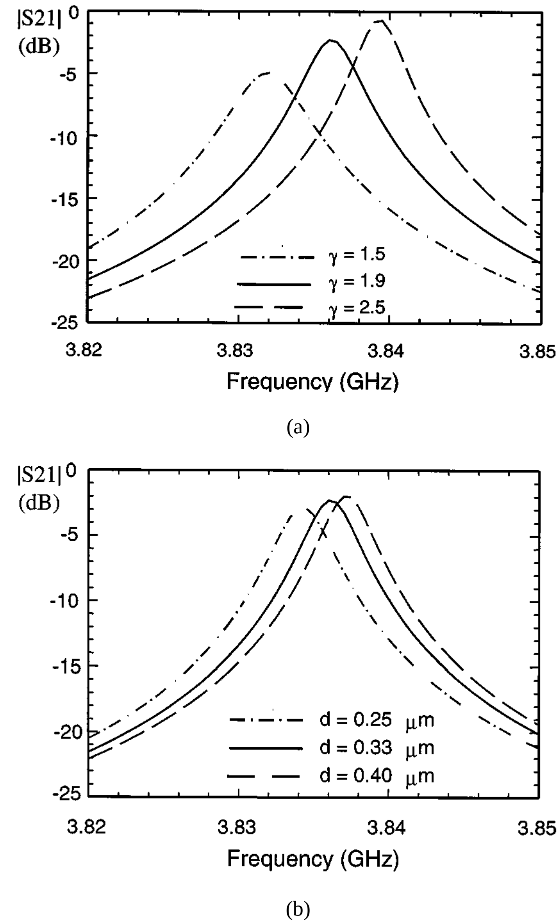


Fig. 4. Simulated characteristics of the YBCO microstrip resonator at $T = 65$ K for: (a) different model parameter γ and (b) different HTS film thickness.

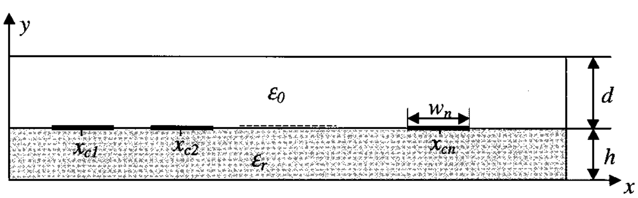


Fig. 5. Cross section of the boxed MCPL.

order to analyze the boxed MCPL in a general case (Fig. 5), we use very effective computational algorithm based on the SDA combined with the Galerkin's method proposed in [21].

The Green's function is estimated in accordance with [22] and is written as

$$G(\alpha_n) = \frac{1}{\alpha_n} \cdot \frac{1}{\coth(\alpha_n d) + \varepsilon_r \coth(\alpha_n h)} \tag{11}$$

where α_n is the discrete Fourier variable.

The following basis functions are used for modeling the surface charge distribution on the conductor surface:

$$\sigma_p^i(x) = \frac{2}{\pi w_i} \frac{T_p\left(\dfrac{x - x_{ci}}{w_i/2}\right)}{\sqrt{1 - \left(\dfrac{x - x_{ci}}{w_i/2}\right)^2}} \tag{12}$$

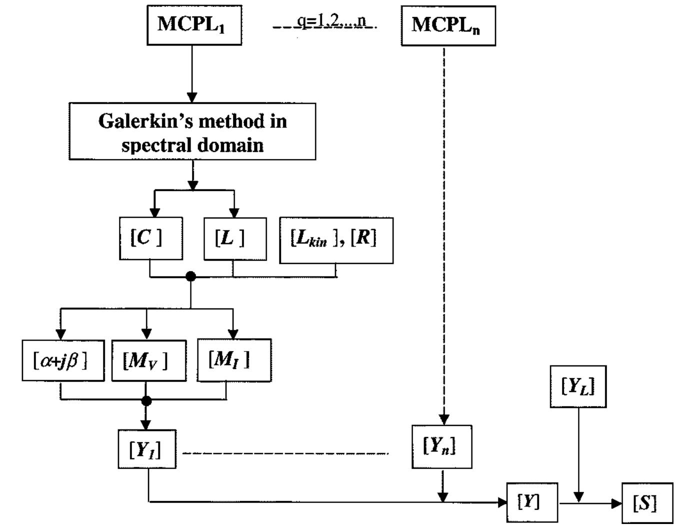


Fig. 6. Block diagram of the computation algorithm.

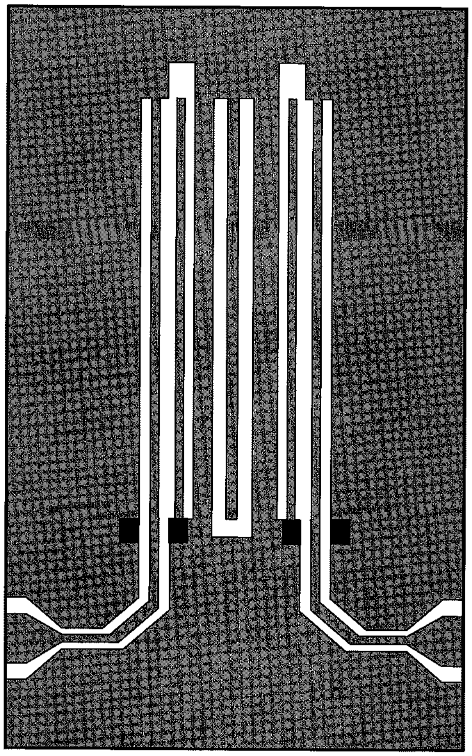
where w_i is the strip width, x_{ci} is the position of the strip center (Fig. 5), and T_p is the pth-order Chebyshev polynomial of the first kind. The surface charge distribution is used for determining the capacitance matrix of the structure.

The lossless MCPL structure in the quasi-TEM approach is fully described by the matrix of capacitance per unit length $[C]$ and the matrix of inductance per unit length $[L]$. The SDA-based algorithm is used for a calculation of the capacitance matrix, comprising the self-capacitances of single lines and all coupling capacitances of the structure. In order to calculate the $[L]$ matrix, it is necessary to evaluate the capacitance matrix $[C_{\mathrm{air}}]$ for the same structure with removed substrate (the dielectric constant of the substrate $\varepsilon_r = 1$). Using the relation between capacitance $[C_{\mathrm{air}}]$ and inductance $[L]$ matrices in nonmagnetic media, one can find

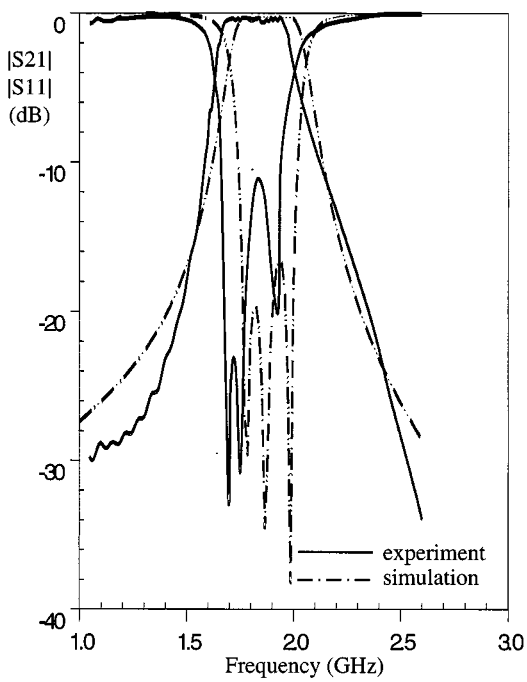
$$[L] = \frac{1}{c^2}[C_{\mathrm{air}}]^{-1}. \tag{13}$$

Here, c is the light velocity in free space.

It is now necessary to discuss some approximations used in the computation. The problem is how to take into account the contribution of the superconducting film parameters. The matrix of wavenumbers $[\beta]$ of the lossless MCPL is computed from the capacitance $[C]$ and inductance $[L]$ matrices. Since the contribution of the HTS film in the propagation characteristics can be considered as a small perturbation, one can find the propagation characteristics of the HTS MCPL by small corrections, which depend on the surface impedance of the HTS film. By adding the kinetic inductance per unit length $L_{kin}^i/w_{\mathrm{eff}}^i$ to the elements $L_{i,i}$ of the matrix $[L]$, one obtains the corrected wavenumbers of eigenmodes of the MCPL. The effective width of the ith strip w_{eff}^i is calculated taking into account the nonhomogeneous current density distribution in the strip cross section. Keeping in mind that the narrow-band filter is the subject of the dominant interest and a weak coupling is provided between adjacent lines, one may assume that the current distribution over the



(a)



(b)

Fig. 7. (a) Layout and (b) performance of the three-pole CPW filter.

strips in the MCPL will be only slightly perturbed as compared to the single line case. For the noncritical substrate thickness (large as compared with the HTS strip thickness), the surface current distribution can be approximated as in [14] without significant error.

The attenuation coefficient α of propagating waves is determined by the dielectric loss of the substrate and by the losses in superconducting strips and ground plane. The dielectric loss of the substrate can be omitted as far as $\tan\delta < 10^{-5}$. Thus, the surface resistance of the HTS film is responsible for the attenuation coefficient in the MCPL structure. The resistance matrix $[R]$ in diagonal form, with the elements defined as $R_{i,\,i} = R_{\text{sur}}/w_{\text{eff}}^{i}$, is introduced for calculation of the attenuation coefficient.

The diagram of the computation procedure is presented in Fig. 6. A set of multicoupled lines of the length l_q forms the block of MCPL_q ($q = 1, \cdots, n$). The procedure allows calculating the eigenvector matrices of voltage $[M_V]$ and current $[M_I]$ for each block, as well as the matrix of complex propagation characteristics $[\alpha + i\beta]$.

Now, the admittance matrix $[Y_q]$ of each block and the admittance matrix $[Y]$ of the whole filter structure can be evaluated. Taking into account the $[Y_L]$ matrix of the external terminations, one can find the $[S]$ matrix of the filter.

D. Design of Bandpass HTS Filters

The derived procedure of analysis of planar HTS filters was applied to a simulation of the performance of different filter structures. All computing processes were performed on a Pentium-II/330 MHz/64 MB RAM using Pascal computer codes. The typical computation time is about 12–15 s for 150 frequency points. The results presented below were obtained with no more than four basis functions (12). This number is enough to ensure a suitable accuracy of the computed characteristics.

Let us consider the simulated and measured characteristics of planar bandpass HTS filters. Fig. 7(a) shows the layout of the three-pole coplanar waveguide (CPW) filter. The filter was designed and manufactured as a multicoupled structure of the YBCO CPW lines on an $LaAlO_3$ substrate. The simulated and measured filter characteristics at $T = 77$ K [10] are depicted in Fig. 7(b). There is a discrepancy between the simulated and measured data. The simulation of the filter characteristic was performed using averaged model parameters of the YBCO film since the actual model parameters of the film have been unknown.

The layout of the ten-pole microstrip pseudointerdigital filter is shown in Fig. 8(a). As in the previous case, the YBCO film on the $LaAlO_3$ substrate was used. The filter was manufactured on the substrate of 0.5-mm thickness. The YBCO film (Prima Tek) of 0.4-0.5-μm thickness provided $R_{\text{sur}} = 0.6$ mΩ at $T = 77$ K ($f = 10$ GHz). The dielectric constant of the substrate $\varepsilon_r = 23.7$ was measured at $T = 300$ K. The temperature dependence of the dielectric constant of $LaAlO_3$ [23], [24] was used for finding ε_r at $T = 77$ K. The operation temperature of the filter was chosen $T = 77$ K. Since the surface resistance at $T = 77$ K is known, one could calculate the attenuation coefficient. We then had to find the correction to the propagating characteristic of the wave, which was caused by the contribution of the kinetic inductance. After two iterations, the parameter γ responsible for the kinetic inductance was found: $\gamma = 1.85$. The simulated and measured characteristics of the filter are presented in Fig. 8(b). Evidently, there is a good coincidence between experimental and simulation results.

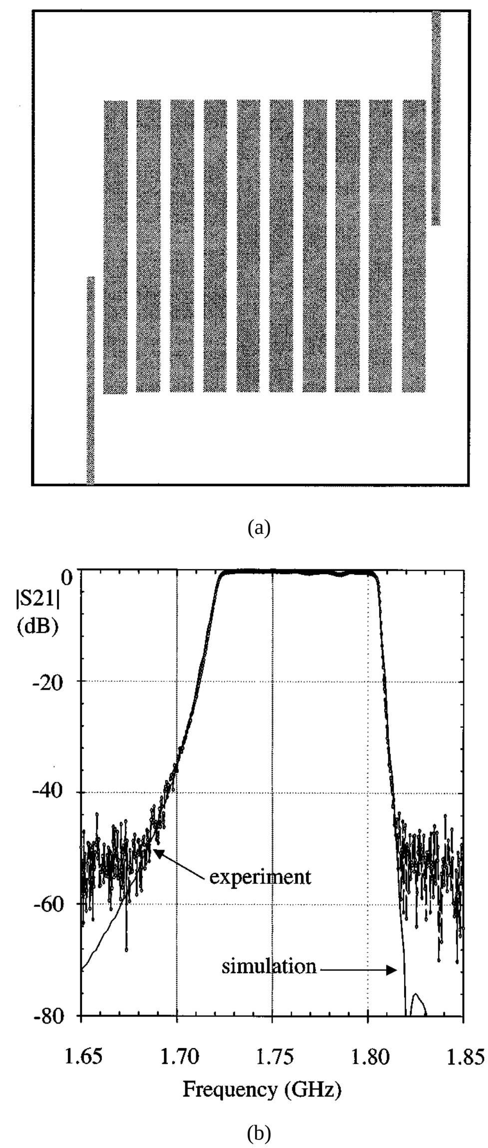


Fig. 8. (a) Layout and (b) performance of the ten-pole pseudointerdigital microstrip filter.

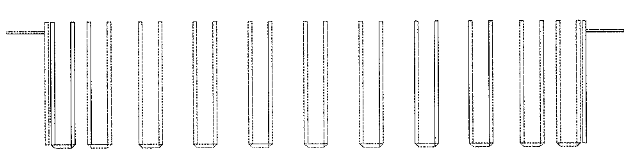


Fig. 9. Layout of the 11-pole hairpin resonator.

The next example demonstrates how the suggested design procedure has been completely realized. The layout of the 11-pole hairpin comb filter [9] is shown in Fig. 9. The single resonator of the filter structure was manufactured and measured at different temperatures [25]. The YBCO film of 0.45-μm thickness on a 0.45-mm-thick $LaAlO_3$ substrate was used for manufacturing the resonator and filter. The experimental characteristics of the resonator allowed extracting the model parameters of the YBCO film surface impedance and the dielectric permittivity of the substrate as well. The experimental resonant characteristics of the hairpin resonator in combination with the model description are shown in Fig. 10 for two different temperatures. The following fitting parameters were found: $T_c = 87.6\,\mathrm{K}$, $\sigma_n(1) = 2.5 \cdot 10^6\,(\Omega \cdot \mathrm{m})^{-1}$, $\gamma = 2.45$, and $\alpha = 13$; the dielectric permittivity of the substrate $\varepsilon_r = 23.6$ was taken at $T = 300$ K. We now use these parameters in the simulation procedure of the filter in accordance with Section II-C. The simulated and measured filter performance is presented in Fig. 11 for two different temperatures. The characteristics are in a very good agreement.

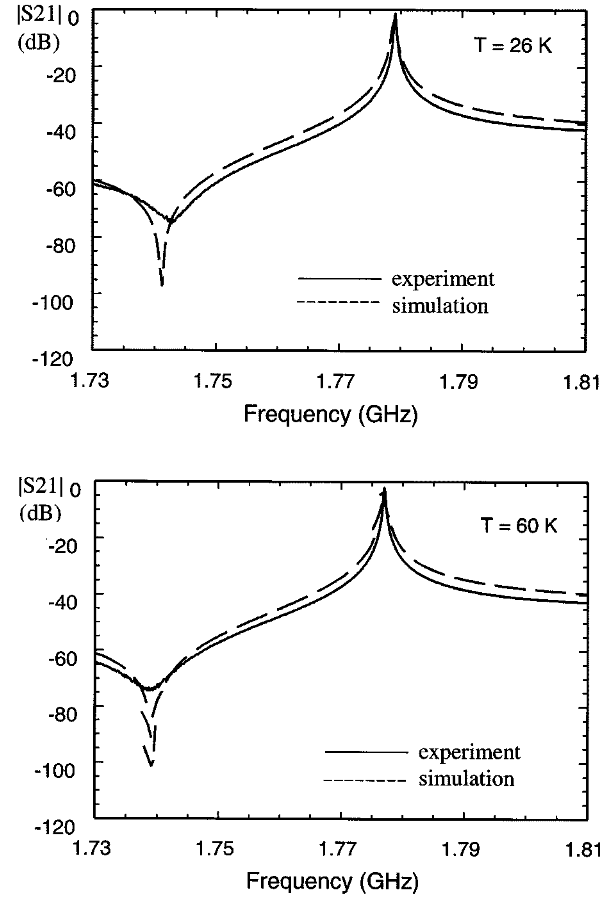


Fig. 10. Simulated and measured performance of the hairpin resonator.

III. Discussion

An accurate procedure of the planar HTS filter design is proposed. Taking into consideration the contribution of the HTS film, surface impedance is the special feature of the computation of the filter performance. The fitting parameters of the surface impedance model should be found from the experimental characteristics of the single HTS resonators. It was shown that the model parameter γ is of crucial importance. This parameter determines the temperature dependence of the London penetration depth and, as a consequence, the

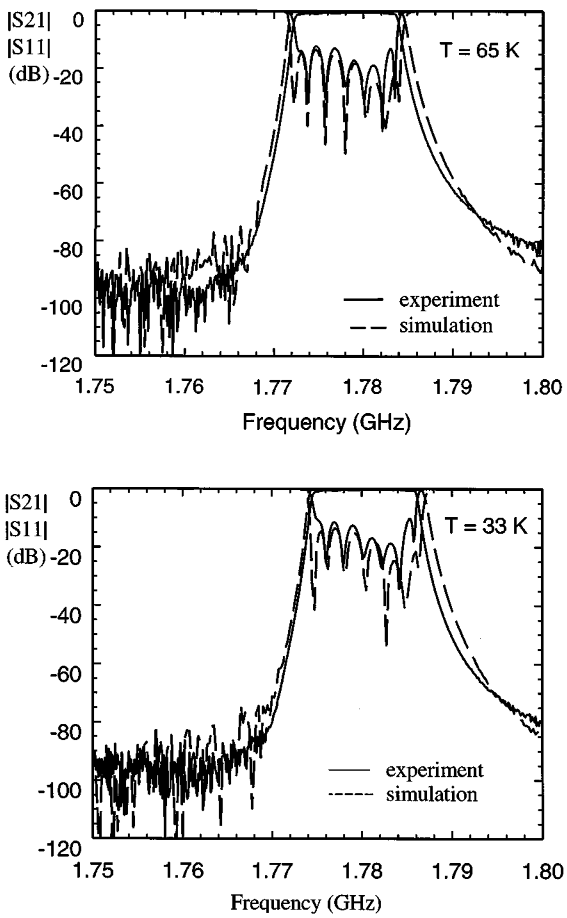


Fig. 11. Simulated and measured performance of the 11-pole hairpin comb filter.

contribution of the kinetic inductance of the HTS film in the phase velocity of the propagating electromagnetic wave. The influence of the parameter γ is much more significant at the temperature close to the transition temperature. The lower the temperature, the less is the contribution from the kinetic inductance and the less is the deviation of the central frequency of the passband of the filter. The central frequency is also strongly dependent on the substrate dielectric permittivity and thickness of the substrate. It is necessary to keep these two parameters invariable from substrate to substrate. The influence of the HTS film thickness can be remarkable in the case of thin HTS film [$d \leq 2\lambda_L(t)$]. In order to avoid the film thickness influence, thick HTS films should be recommended for use in filter structures. It was found that the film thickness $d \geq 0.6$ μm was enough to provide a low sensitivity of the filter performance to a small deviation of the film thickness (± 0.1 μm). From the point-of-view of the filter performance stability with respect to the substrate thickness, the CPW filter structure is preferable.

The following procedure of the HTS planar filter design is recommended.

1) A planar resonator of the same configuration as used in the planned filter structure should be manufactured and measured at different temperatures ($T < T_c$).
2) Model parameters of the microwave surface impedance of the HTS film and the dielectric permittivity of the substrate material are to be extracted from the experimental characteristics of the resonator.
3) The developed computation algorithm can be used for the filter performance simulation; in this case, the filter structure can be easily changed in order to find an optimal configuration.
4) The suitable filter structure found as a result of simulation may be used for manufacturing and measuring the filter (first iteration).
5) The experimental characteristics of the filter are to be used for a correction of the model parameters and the filter geometry.
6) It is expected that the second iteration is the final one.

The evaluation of tolerances of characteristics of the HTS film and dielectric substrate should be considered as a special problem. The solution to this problem is in progress.

IV. Conclusion

An effective computation procedure based on a more correct model of HTS transmission lines is proposed and verified by comparison of simulated and measured characteristics of narrow-band HTS filters of different configurations. The method provides designing filters with a high accuracy and avoids any trimming after the filter manufacturing. There are reasons to believe that the method proposed can provide manufacturing the HTS filters in mass production.

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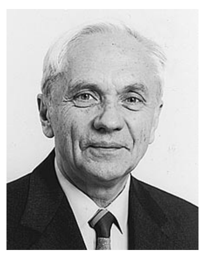

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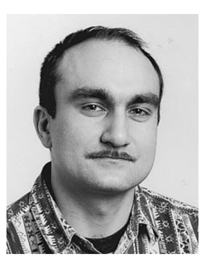

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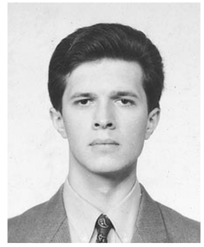

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Mr. Kholodniak was the recipient of a grant presented by the President of Russia.