Electrodynamic Theory of Multiport Structures Using Magnetostatic Waves in Ferrite Films and Its Applications

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***Abstract*—The paper presents an electrodynamic analysis of tunable multiport ferrite-dielectric structures with parallel transmission lines of an arbitrary type, coupled through propagating magnetostatic modes of magnetized multilayered ferrite films. The structures are supposed to be excited at one port by an incident electromagnetic wave, and amplitudes and phases of electromagnetic waves at other ports are obtained by an analytical procedure. The model holds for an arbitrary direction of a magnetizing field and describes the interaction of magnetostatic modes in ferrite films of a finite width. The solution is obtained in a self-consistent approach, i.e., a reaction of magnetostatic waves (MSWs) on transducers, which excite them, is taken into account. Derived closed-form expressions for response functions of multiports provide the base for the modeling of a wide class of MSW devices: multichannel adjustable filters and delay lines, directional couplers, frequency-selective power dividers, tunable oscillators and active filters, and multiport resonators. The theory is also valid for the analysis of multi-element, interdigital, and meander MSW transducers. Applications of a general theory are demonstrated for numerical calculations of frequency responses of surface and forward volume MSW filters, delay lines with new types of strip-line transducers (two-port and T-type), and for the analysis of a phenomenon of mutual coupling of transducers in conventional devices.**

***Index Terms*—Ferrimagnetic films, ferrite devices, magnetostatic waves (MSWs), microwave integrated circuits.**

I. Introduction

MAGNETOSTATIC wave (MSW) technology has been advancing during recent years and is proposing effective solutions for microwave and UHF signal processing systems [1]. Multiport circuits using ferrite films can significantly extend applications of MSW devices providing the base for their novel designs. Various MSW devices such as adjustable filters and delay lines [2], [3], tunable oscillators and active filters [4], [5], directional couplers [6], frequency-selective power dividers and multiport resonators [2], frequency and phase selective switchers [7], and modulators are realized by using pin-diodes, varactors, transistors, and Gunn diodes as terminations of transducers in multiport structures. Microstrip multiport structures are also used for the realization of MSW transducers themselves in interdigital, meander, and unidirectional configurations [1], [8], which can provide narrow-band frequency responses or to reduce the insertion loss in devices due to the bidirectional radiation of waves. The interaction of transducers in all these multiport structures, especially strong in the case of volume MSW, and an arbitrary characteristic of loadings significantly complicate the electrodynamic analysis of structures and such analysis has not been performed until the present time [9]–[11].

As recently shown in [12], a self-consistent electrodynamic problem of the MSW excitation and reception by a transducer of an arbitrary type can be analytically solved in a general case, if a complete set of eigenwaves propagating in a ferrite film and their dispersion characteristics are known. The approximation for the obtained solution [12] depends on the approximation for propagating waves—electrodynamic, magnetostatic, or dipole-exchange. The solution is constructed by using an electromagnetic field distribution for a dominant mode in a transmission line forming a transducer (in the absence of a ferrite film). This approach can be extended to multiport structures and arbitrary magnetized multilayered infinite or finite-width ferrite films (MSW waveguides). Although, in fact, an explicit form of eigenmodes in ferrite films and their dispersion characteristics for such complicated structures are known only for some special cases (e.g., see [13]–[15]), the possibility of constructing an analytical solution in a general case is very attractive. Moreover, in calculations according to this approach, there can be used numerical solutions for MSW modes, applicable for arbitrary ferrite-dielectric waveguiding structures [16], [17], along with numerical solutions for an electromagnetic field in dielectric-filled transmission lines forming transducers (e.g., see [18], [19]).

In this paper, such a unified technique, to provide solutions for frequency responses of general multiports using MSWs in multilayered thin ferrite films, is presented as an extension of the approach [12]. The developed theory is based on the analytical solution of the system of singular integral equations, each pair of equations formulating a self-consistent electrodynamic problem for an individual transducer. Electromagnetic coupling of transducers besides magnetostatic modes of ferrite films is neglected. By using a general theory, numerical calculations are performed for frequency responses of surface and forward volume MSW filters and delay lines with new types of strip-line transducers (two-port and T-type) and for the analysis of the phenomenon of mutual coupling of transducers in conventional devices.

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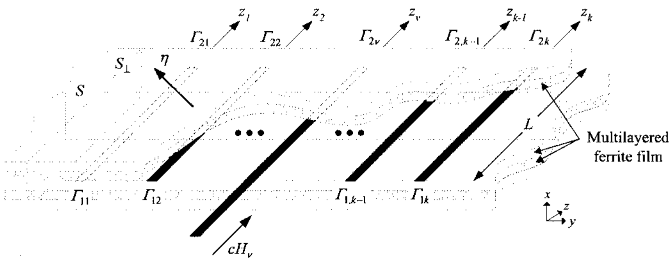

Fig. 1. MSW multiport structure using multilayered ferrite film (z_p are longitudinal axes of transducers; $S_\perp$ and S are the cross and longitudinal sections of the film).

II. General Theory

Consider a multiport MSW structure with k parallel transducers of an arbitrary type coupled to an m-layered ferrite film infinite along the y axis (Fig. 1). Ferrite layers can be separated by dielectric layers, forming a multilayered ferrite-dielectric waveguiding structure for MSWs. In the case of the metallization of layers from above or from below, the structure represents a system of parallel slot-lines.

The structure is excited at the port 1ν at the frequency ω, and the magnetic field of the incident electromagnetic wave of the dominant ν-mode in the feeding transmission line at $z_\nu = 0$ is $c\mathbf{H}_\nu$. Terminations at ports give reflection coefficients for transverse magnetic fields of incident electromagnetic waves Γ_{1p} $(p = 1, 2, \ldots, k)$ at $z_p = 0$ $(p \neq \nu)$ and Γ_{2p} at $z_p = L$ (which in a special case of meander-type transducers may be reflection coefficients of neighboring transducers, transformed to a corresponding port). Eigenwaves of the magnetization $\mathbf{m}_{nj} = \mathbf{m}^0_{nj}(x, z)\exp(-ik_n y)$ of the jth layer $(j = 1, 2, \ldots, m)$ and dispersion characteristics $\omega_n(k_n)$ of waves in a multilayered film, magnetized to the saturation in an arbitrary direction η, are supposed to be known (x is the axis orthogonal to the ferrite film; k_n is the longitudinal wavenumber of the nth mode). Letting in transmission lines forming transducers there can propagate (in the absence of a ferrite film) only fundamental electromagnetic modes of the types p with the electromagnetic fields

$$\begin{aligned}\mathbf{E}_{\pm p} &= \mathbf{E}_{\pm p0}(x, y)\, e^{\pm i\gamma_p z_p}\\ \mathbf{H}_{\pm p} &= \mathbf{H}_{\pm p0}(x, y)\, e^{\pm i\gamma_p z_p}\end{aligned} \tag{1}$$

where γ_p are the propagation constants.

Using the principle of a superposition, the high-frequency magnetization of the m-layered ferrite film interacting with the pth transducer can be represented as

$$\mathbf{M}_p = \sum_{j=1}^{m} \mathbf{M}_{pj} \tag{2}$$

where $\mathbf{M}_{pj}$ is the high-frequency magnetization of the jth layer. According to [12], the magnetic field of this transducer, external for the film, can be derived in the one-mode approximation from the Lorentz lemma for a transmission line, excited by the given equivalent magnetic current $i\omega\mu_0\mathbf{M}_p$ (μ_0 is the permeability of free space). Thus, for $(p \neq \nu)$, we have [12]

$$\mathbf{h}_p = c_p(z)\,\mathbf{H}^r_p + c_{-p}(z)\,\mathbf{H}^r_{-p} \tag{3}$$

where

$$\begin{aligned}c_p &= -\frac{i\omega\mu_0}{N^r_p}\int_0^{Z_p} dz_p \int_S \left(\mathbf{M}_p \cdot \mathbf{H}^r_{-p}\right) dS\\ c_{-p} &= -\frac{i\omega\mu_0}{N^r_p}\int_{Z_p}^{L} dz_p \int_S \left(\mathbf{M}_p \cdot \mathbf{H}^r_{p}\right) dS\end{aligned} \tag{4}$$

$$\begin{aligned}\mathbf{H}^r_p &= \mathbf{H}_p + \Gamma_{2p}\mathbf{H}_{-p}\\ \mathbf{H}^r_{-p} &= \mathbf{H}_{-p} + \Gamma_{1p}\mathbf{H}_{p}\end{aligned} \tag{5}$$

$$N^r_p = (1 - \Gamma_{1p}\Gamma_{2p})\, N_p \tag{6}$$

$$N_p = \int_S \left\{[\mathbf{E}_{-p} \times \mathbf{H}_p] - [\mathbf{E}_p \times \mathbf{H}_{-p}]\right\}_{Z_p} dS \tag{7}$$

and for $p = \nu$

$$\begin{aligned}\mathbf{h}_\nu &= [c + c_\nu(z_\nu)]\,\mathbf{H}_\nu,\\ &\quad + \left\{c_{-\nu}(z_\nu) + [c + c_\nu(L)]\,\Gamma_{2\nu}e^{-2i\gamma_\nu L}\right\}\mathbf{H}_{-\nu}\end{aligned} \tag{8}$$

$$\begin{aligned}c_\nu &= -\frac{i\omega\mu_0}{N_\nu}\int_0^{Z_\nu} dz_\nu \int_S (\mathbf{M}_\nu \cdot \mathbf{H}_{-\nu})\, dS\\ c_{-\nu} &= -\frac{i\omega\mu_0}{N_\nu}\int_{Z_\nu}^{L} dz_\nu \int_S (\mathbf{M}_\nu \cdot \mathbf{H}_{\nu})\, dS.\end{aligned} \tag{9}$$

The normalization coefficient N_ν in (9) is calculated according to (7). The one-mode approximation (3) and (8) is valid, when ferrite fills a small part of a transmission-line cross section (that takes place in most cases, when ferrite films are used).

The high-frequency magnetization of a ferrite layer, excited by the given magnetic field of an external source, can be represented as an integral mode expansion into longitudinal wavenumbers of eigenwaves, with expansion coefficients being obtained from the linearized Landau–Lifshitz equation [12]. Residues of the integral give propagating MSWs, while its principle value comprises the local magnetization, which induces the magnetic field, giving, in the sum with the external field, the near field of a source. If, in the structure under consideration, the near field of the pth transducer vanishes in the vicinity of neighboring transducers, the magnetization of the jth layer in the neighborhood of the pth transducer can be represented following [12], as the sum of the magnetization excited by the pth transducer and the magnetization of propagating MSW coming from other transducers, with magnetic fields of transducers $\mathbf{h}_p$ $(p = 1, 2, \ldots, k)$ being considered as external fields of an assumed distribution

$$\begin{aligned}\mathbf{M}_{pj} &= \sum_n \int_{-\infty}^{+\infty} c_{np}\mathbf{m}_{nj}\, dk_n\\ &\quad - 2\pi i \sum_n \sum_{q-1}^{p-1} \operatorname{res}\left(c_{nq}\mathbf{m}_{nj}e^{-ik_n l_{qp} - k'_n l_{qp}}\right)_{\omega+}\\ &\quad + 2\pi i \sum_n \sum_{q=p+1}^{k} \operatorname{res}\left(c_{nq}\mathbf{m}_{nj}e^{ik_n l_{qp} - k'_n l_{qp}}\right)_{\omega-}.\end{aligned} \tag{10}$$

The expansion coefficients are obtained as [12]

$$c_{np} = \varphi_{nj}\int_{V_j} \left(\mathbf{h}_p \cdot \mathbf{m}^*_{nj}\right) dV_j \tag{11}$$

where

$$\varphi_{nj} = \frac{i\omega_{Mj}}{\Phi_{nj}} \cdot \frac{1}{\omega_n - \omega} \tag{12}$$

$$\Phi_{nj} = 2\pi \int_{S_\perp} \left[\mathbf{m}_{nj}^0 \times \mathbf{m}_{nj}^{0*}\right]_\eta dS_\perp \tag{13}$$

where $\omega_{Mj} = \mu_0 \gamma M_{sj}$, γ is the gyromagnetic ratio, M_{sj} is the saturation magnetization of the jth ferrite layer, V_j is the volume of the jth layer, and l_{qp} is the distance between the qth and the pth transducers. Indexes $\omega_\pm$ in (10) indicate the direction of the propagation of resonantly excited MSW $(\omega_n = \omega)$ along the y axis and the residues are calculated for corresponding eigenwaves in the coordinate system of the qth transducer. It is assumed in (10) that $k_n(\omega_n = \omega_+) > 0$, $k_n(\omega_n = \omega_-) < 0$ and $k_n'(\omega_n = \omega_\pm) > 0$, where $k_{n\omega}' = \gamma \Delta H_k / V_{n\omega}$ is the imaginary part of the wavenumber at the frequency ω, which takes into account the loss in ferrite [20], ΔH_k is the half-linewidth of the ferromagnetic resonance (in a general case depending on the wavenumber k_n), and $V_{n\omega} = \partial \omega n / \partial k_n$ is the group velocity of the wave. Note that in a general case $k_{n\omega+} \neq k_{n\omega-}$ and $k_{n\omega+}' \neq k_{n\omega-}'$.

The system of singular integral (3), (8), and (10) for the unknown magnetic fields $\mathbf{h}_p$ and the magnetizations $\mathbf{M}_{pj}$ $(p = 1, 2, \ldots, k;\; j = 1, 2, \ldots, m)$ formulates the self-consistent electrodynamic problem for the structure. On the other hand, this system can be considered, as equations for unknown integral parameters $c_{\pm p}(z_p)$ (4) and (9), which give complex variable amplitudes of electromagnetic waves in transducers propagating along and opposite the z_p axis. If the distribution of the magnetization in eigenwaves is known, it is possible to resolve (3), (8), and (10) in a general case by reducing them to a system of linear algebraic equations. Response functions of the structure are then expressed in terms of these variables according to [12]. Substituting the magnetic field (1) into the system of singular integral equations and considering transmission lines with a dominant quasi-T wave with a transverse magnetic field, so that, $H_{\pm pz} = 0$, $H_{p0} = H_{-p0} = H_{-p0}^*$ [12], the solution of the system for a fundamental MSW mode, which has no variation of the magnetization over the z_p axes, can be obtained as given in (14), shown at the bottom of this page, and

$$\left.\begin{aligned} c_\nu(z_\nu) &= \left(e^{i\gamma_\nu z_\nu} - 1\right) c_{\nu 0}, \\ c_{-\nu}(z_\nu) &= \left(e^{-i\gamma_\nu z_\nu} - e^{-i\gamma_\nu L}\right) c_{\nu 0}, \end{aligned}\right\}, \quad \text{for } p = \nu \tag{15}$$

where the constants c_{p0} $(p = 1, 2, \ldots, k)$ satisfy the kth-order system of linear algebraic equations

$$\sum_{q=1}^{k} W_{qp} c_{q0} = \sum A_p. \tag{16}$$

Coefficients W_{qp} form a square matrix $\hat{W}$ with the following elements:

$$\text{for } q \neq \nu \quad W_{qp} = \begin{cases} i2s_q T_{qp}, & q < p \\ -2is_q R_{qp},, & q > p \\ \gamma_p \left(1 - \Gamma_{1p}\Gamma_{2p}\right) - i2s_p F_p, & q = p \end{cases} \tag{17}$$

and

$$\text{for } q = \nu \quad W_{\nu p} = \begin{cases} ip_\nu T_{\nu p}, & \nu < p \\ -ip_\nu R_{\nu p}, & \nu > p \\ \gamma_\nu - ip_\nu F_\nu, & \nu = p. \end{cases} \tag{18}$$

The column of absolute terms in (16) is proportional to the amplitude c of the incident electromagnetic wave at the port 1ν

$$A_p = \begin{cases} -q_\nu \gamma_\nu^{-1} T_{\nu p} c, & p > \nu \\ q_\nu \gamma_\nu^{-1} R_{\nu p} c, & p < \nu \\ q_\nu \gamma_\nu^{-1} F_\nu c, & p = \nu. \end{cases} \tag{19}$$

In (17)–(19), the following notations are introduced:

$$\begin{aligned} s_p = {} & L + \frac{1}{i\gamma_p}\left(e^{-i\gamma_p L} - 1\right) + \frac{\Gamma_{1p}}{2i\gamma_p}\left(e^{-i\gamma_p L} - 1\right)^2 \\ & + \frac{\Gamma_{2p}}{2i\gamma_p}\left(e^{i\gamma_p L} - 1\right)^2 \\ & - \Gamma_{1p}\Gamma_{2p}\left[L - \frac{1}{i\gamma_p}\left(e^{i\gamma_p L} - 1\right)\right] \end{aligned} \tag{20}$$

$$\begin{aligned} p_\nu = {} & \left[2\left(i\gamma_\nu L + e^{-i\gamma_\nu L} - 1\right) + \Gamma_{2\nu}\left(1 - e^{-i\gamma_\nu L}\right)^2\right] \\ & \times \left(i\gamma_\nu\right)^{-1} \end{aligned} \tag{21}$$

$$q_p = \left(1 - e^{-i\gamma_p L}\right)\left(1 + \Gamma_{2p} e^{-i\gamma_p L}\right) \tag{22}$$

$$\begin{aligned} T_{qp} = {} & -2\pi \frac{i\omega\mu_0}{N_p} \sum_{j=1}^{m} \sum_n \omega_{Mj} \\ & \times \left[\frac{I_{qnj} I_{pnj}^*}{V_n \Phi_{nj}} \cdot e^{-ik_n l_{qp} - k_n' l_{qp}}\right]_{\omega +} \end{aligned} \tag{23}$$

$$\begin{aligned} R_{qp} = {} & -2\pi \frac{i\omega\mu_0}{N_q} \sum_{j=1}^{m} \sum_n \omega_{Mj} \\ & \times \left[\frac{I_{pnj} I_{qnj}^*}{V_n \Phi_{nj}} \cdot e^{ik_n l_{qp} - k_n' l_{qp}}\right]_{\omega -} \end{aligned} \tag{24}$$

$$F_p = \frac{i\omega\mu_0}{N_p} \sum_{j=1}^{m} \sum_n \int_{-\infty}^{+\infty} \varphi_{nj} \left|I_{pnj}\right|^2 dk_n \tag{25}$$

where the "integrals of an excitation" [12]

$$I_{pnj} = \int_{S_j} \left(\mathbf{H}_{p0} \cdot \mathbf{m}_{nj}^*\right) dS_j \tag{26}$$

are calculated in the coordinate system of the pth transducer. The parameters s_p, p_ν, and q_ν depend only on the electrical lengths

$$\left.\begin{aligned} c_p(z_p) &= \left(e^{i\gamma_p z_p} - 1\right) c_{p0} + \Gamma_{1p}\left(1 - e^{-i\gamma_p z_p}\right) c_{p0}, \\ c_{-p}(z_p) &= \left(e^{-i\gamma_p z_p} - e^{-i\gamma_p L}\right) c_{p0} + \Gamma_{2p}\left(e^{i\gamma_p L} - e^{i\gamma_p z_p}\right) c_{p0}, \end{aligned}\right\}, \quad \text{for } p \neq \nu \tag{14}$$

of transducers and the reflection coefficients of terminations at corresponding ports. Diagonal components of the matrix $\hat{W}$ contain the complex "parameters of coupling" F_p (25), which completely characterize the coupling of each transducer to the film [12], while off-diagonal components describe the mutual coupling of transducers. The parameters T_{qp} [see (23)] and matrix elements below the diagonal determine the signal transit from the qth transducer to the pth one $(q < p)$ and R_{qp} [see (24)] with the matrix elements above the diagonal characterize their back coupling.

Response functions of the structure are easily expressed in the terms of (14) and (15) from the field representations (3) and (8). Transmission coefficients for the magnetic fields of incident electromagnetic waves from the port 1ν to the ports $1p$ and $2p(p \neq \nu)$ are

$$D_{1p} = \frac{c_{-p}(0)\,H_{p0}}{cH_{\nu 0}} \quad D_{2p} = \frac{c_p(L)\,H_{p0}}{cH_{\nu 0}} \tag{27}$$

and from the port 1ν to the port 2ν is

$$D_{2\nu} = \frac{[c + c_\nu(L)]\,e^{-i\gamma_\nu L}}{c}. \tag{28}$$

The reflection coefficient from the input port of the structure reads

$$\Gamma_{1\nu} = \frac{\left\{c_{-\nu}(0) + \Gamma_{2\nu} e^{-2i\gamma_\nu L}\,[c + c_\nu(L)]\right\}}{c}. \tag{29}$$

If different types of transmission lines are employed as transducers in the structure, the following normalization of transmission coefficients is required:

$$K_{ip} = D_{ip}\frac{H_{\nu 0}}{H_{p0}}\sqrt{\frac{N_p}{N_\nu}}, \qquad i = 1, 2 \tag{30}$$

so that the power transmission coefficient (in dB) is calculated as

$$b_{ip} = 20\log|K_{ip}| \tag{31}$$

while the phase of K_{ip} gives the phase shift between electromagnetic fields of waves incident at the ports ip $(i = 1, 2;\ p = 1, 2, \ldots, k;\ p \neq \nu)$ and the port 1ν. In the case when feeding transmission lines of the device and transmission lines forming the transducers (in the absence of a ferrite film) have different characteristic impedances, the resultant transmission coefficient for incident waves should take into account these step discontinuities, according to [12].

III. Phenomenon of Mutual Influence of Transducers in MSW Filters and Delay Lines

Consider a conventional MSW filter (delay line) containing an input and one output transducer. Denote by the index ν all the values relating to the input transducer and by the index μ relating to the output one. Resolving the algebraic system (16), we have (32) and (33), shown at the bottom of this page. Using these relations and (29) and (30), it is not difficult to obtain closed-form expressions for response functions of the device. In particular, the transmission coefficient for the magnetic fields of incident electromagnetic waves from port 1ν to port 1μ reads as given in (34), shown at the bottom of this page.

In the assumption of a negligible back coupling of the output transducer to the input one $(R_{\nu\mu} \equiv 0)$, this expression coincides with the earlier obtained result [12]. As it follows from (34), (23), and (24), the mutual influence of transducers is small in the case of the high magnetic loss in ferrite and the large distance between transducers, when $\exp\left[-\left(k'^{+}_{n\omega} + k'_{n\omega-}\right) l_{\nu\mu}\right] \ll 1$, and also under the significant asymmetry of the radiation of MSW, when

$$\left|I_{\mu n} I^*_{\nu n}\right|_{\omega-} << \left|I_{\nu n} I^*_{\mu n}\right|_{\omega+}. \tag{35}$$

This takes place for surface MSWs in the case of the strong coupling of transducers to a ferrite film. In low-loss filters using surface MSWs, when the optimum coupling is achieved by moving away transducers from a film [12], the condition (35) may not be satisfied. In all cases, the phenomenon of the mutual coupling of transducers appears to be more apparent for volume MSWs, when amplitudes of waves, radiated in opposite directions, are equal to each other.

Since the term $T_{\nu\mu}R_{\nu\mu} \sim \exp(-ik_{n\omega+} - i|k_{n\omega-}|)l_{\nu\mu}$ in the denominator of (34) varies rapidly with the frequency, the strong back coupling of transducers can cause passband ripples in the frequency responses of the devices. As seen from (34), this effect will be the strongest when

$$1 - \Gamma_{1\mu}\Gamma_{2\mu} \approx 0. \tag{36}$$

In the case of a large difference in characteristic impedances of a feeding transmission line and a line forming a transducer, this

$$c_{\nu 0} = \frac{ic}{\gamma_\nu} \cdot \frac{q_\nu\left\{F_\nu\left[i\gamma_\mu\left(1 - \Gamma_{1\mu}\Gamma_{2\mu}\right) + 2s_\mu F_\mu\right] + 2s_\mu T_{\nu\mu}R_{\nu\mu}\right\}}{\left(i\gamma_\nu + p_\nu F_\nu\right)\left[i\gamma_\mu\left(1 - \Gamma_{1\mu}\Gamma_{2\mu}\right) + 2s_\mu F_\mu\right] + 2s_\mu T_{\nu\mu}R_{\nu\mu}} \tag{32}$$

$$c_{\mu 0} = \frac{cq_\nu T_{\nu\mu}}{\left(i\gamma_\nu + p_\nu F_\nu\right)\left[i\gamma_\mu\left(1 - \Gamma_{1\mu}\Gamma_{2\mu}\right) + 2s_\mu F_\mu\right] + 2s_\mu p_\nu T_{\nu\mu}R_{\nu\mu}} \tag{33}$$

$$K_{11} = q_\nu q_\mu\sqrt{\frac{N_\mu}{N_\nu}} \cdot \frac{T_{\nu\mu}}{\left(i\gamma_\nu + p_\nu F_\nu\right)\left[i\gamma_\mu\left(1 - \Gamma_{1\mu}\,\Gamma_{2\mu}\right) + 2s_\mu F_\mu\right] + 2s_\mu p_\mu T_{\nu\mu}R_{\nu\mu}} \tag{34}$$

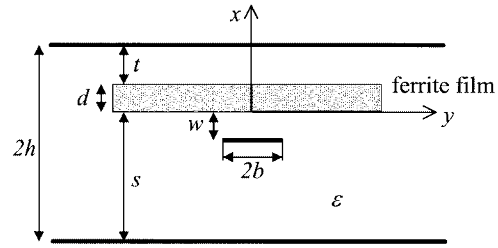

Fig. 2. Cross section of a strip-line MSW transducer.

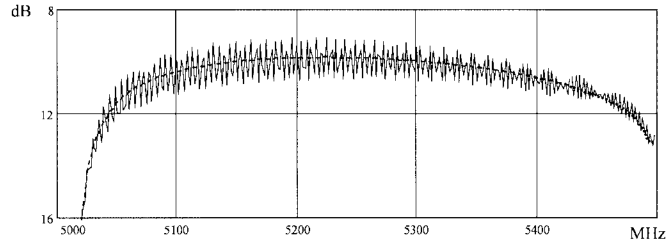

Fig. 3. Power transmission coefficient (in decibels) of a linear dispersive forward volume MSW delay line in a frequency band (dotted curve—back coupling of transducers neglected).

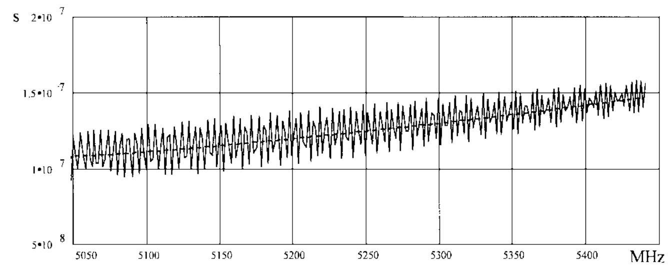

Fig. 4. Delay time (in seconds) of a linear dispersive forward volume MSW delay line in a frequency band (dotted curve—back coupling of transducers neglected).

condition is approximately satisfied for a short-circuited transducer of a resonant length $L \approx \lambda n/2,\ (n = 1, 2, 3, \ldots)$, where λ is the length of an electromagnetic wave in the absence of a ferrite film.

The delay time in delay lines is usually defined as the relation $\tau = l_{\nu\mu}/V_{n\omega}$ [1], which takes into account only the process of the propagation of waves between transducers. As seen from (34) and (23), the total phase shift in a delay line Φ_Σ is summed up of the phase shift $k_{n\omega} + l_{\nu\mu}$, associated with the propagation of waves, and an additional phase shift, caused by the processes of the excitation and the reception of waves. The latter has rapid variations with the frequency in the case of the strong back coupling of transducers. The correct determination of the time delay of a signal in a device requires the generalization of the definition of this notion as follows:

$$\tau = \frac{\partial \Phi_\Sigma}{\partial \omega} \tag{37}$$

where the total phase shift in a device is obtained from (34), as $\Phi_\Sigma = \arg(K_{11})$.

Numerical calculations were performed for a linear dispersive forward volume MSW delay line using YIG film and transducers on the base of a symmetrical strip-line with the dielectric filling of the relative permittivity $\varepsilon = 10$ (Fig. 2). Input and output transducers were supposed to be identical and connected to feeding transmission lines having the characteristic impedance of 50 Ω. Eigenwave functions of the magnetization, used in the computations, are presented in [9] and take into account the influence of metal screens at the distances t and s from the film. The electromagnetic field of a strip-line was used in the form obtained in [21].

Results of numerical calculations presented in Figs. 3 and 4 correspond to open-circuited (at the second ends) transducers and to the following set of the parameters of the structure: $h =$ 0.5 mm, $b = 30$ μm, $d = 15$ μm, $w = 0$, $L = 6$ mm, $l_{\nu\mu} =$ 12 mm, $M_s = 140 \cdot 10^3$ A/m, and $\Delta H_0 = 16$ A/m. The internal magnetizing field is $H_i = 143.2 \cdot 10^3$ A/m.

Figs. 3 and 4 show that the back coupling of transducers in MSW delay lines can cause passband ripples in frequency dependencies both of the power transmission coefficient and of the delay time, as in both cases amplitudes of these ripples are larger at the lower end of the frequency band. Note that these phenomena were earlier observed in [22] by Marcelli *et al.* The authors showed that the triple transit of the backoupled signal between transducers cause the passband ripples in [22].

IV. MSW Filters and Delay Lines With Two-Port Transducers

Consider an MSW filter (delay line) with a new type of transducer, which will be referred to as two-port transducers (Fig. 5). They are formed of two parallel transducers of an arbitrary type, with the second transducer terminating the first one through a transmission-line section. In such structures, unlike usual meander configurations [1], connecting transmission lines are located outside a ferrite film and can be comparatively long, while the coupling between transducers through a ferrite film is realized only by propagating MSW modes. The structure is excited at port 11 and has an arbitrary loading at output port 14. The reflection coefficients Γ_{12}, Γ_{13}, and Γ_{14} at the section $z = 0$ are supposed to be given.

The application of the general procedure, developed in Section II, to this structure requires the solution of a self-consistent problem for determination of the reflection coefficients $\Gamma_{2p}\ (p = 1, 2, 3, 4)$ at section $z = L$, since loadings of the first and third transducers depend both on the coupling of transducers 1,2 and 3,4 through a ferrite film and on their back coupling through transmission-line sections. Let us write down the equations of this self-consistent problem. Denote, as a_p, the amplitudes of electromagnetic waves, which propagate in connecting transmission lines at section $z = L$, as shown in Fig. 5. Then, the reflection coefficients can be written down in the form

$$\begin{aligned} \Gamma_{21} &= \frac{a_2}{a_1} e^{-i\gamma L_{12}} \\ \Gamma_{22} &= \frac{a_1}{a_2} e^{-i\gamma L_{12}} \\ \Gamma_{23} &= \frac{a_4}{a_3} e^{-i\gamma L_{34}} \\ \Gamma_{24} &= \frac{a_3}{a_4} e^{-i\gamma L_{34}} \end{aligned} \tag{38}$$

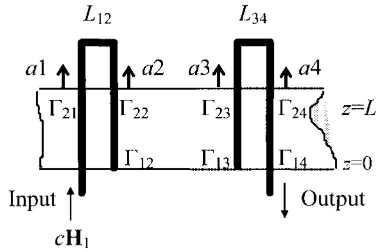

Fig. 5. MSW filter (delay line) with two-port transducers.

where L_{12} and L_{34} are the lengths of connecting transmission lines and γ is the propagation constant of a fundamental electromagnetic wave in these lines. It is obvious that these coefficients are related as follows:

$$\begin{aligned}\Gamma_{21} &= (\Gamma_{22})^{-1} e^{-2i\gamma L_{12}}\\ \Gamma_{23} &= (\Gamma_{24})^{-1} e^{-2i\gamma L_{34}}.\end{aligned} \tag{39}$$

Using (3), (8), (14), and (15), one can represent amplitudes of electromagnetic waves in connecting transmission lines through the amplitudes of electromagnetic waves incident at the ports $2p$ of the transducers at the section $z = L$ as

$$a_1 = c + c_1(L) = c + \left(e^{i\gamma L} - 1\right) c_{10} \tag{40}$$

$$a_p = c_p(L) = \left(1 - e^{-i\gamma_p L}\right)\left(e^{i\gamma_p L} + \Gamma_{1p}\right) c_{p0}, \quad p = 2, 3, 4. \tag{41}$$

Resolving the system of algebraic equations (38)–(41), we can find the reflection coefficients at ports $2p$ and then apply a general procedure, where these coefficients were considered to be known.

Numerical calculations were performed for a surface MSW filter (narrow-band delay line) with short-circuited strip-line two-port transducers connected to the feeding transmission lines with the characteristic impedance 50 Ω. Results of the calculations, shown in Fig. 6, correspond to the following set of parameters of the structure (the notations are the same as in Fig. 2): $h = 0.5$ mm, $b = 250$ μm, $d = 16$ μm, $w = 0$, $L = 3.5$ mm, $L_{12} = L_{34} = 2.5$ mm, $M_s = 60.8 \cdot 10^3$ A/m, $\Delta H_0 = 48$ A/m, and $\varepsilon = 9.8$. The internal magnetizing field is $H_i = 42.4 \cdot 10^3$ A/m. The distance between inner transducers is 2.5 mm.

In contrast to conventional MSW transducers, the presence of additional elements in two-port transducers provides the capability of adjusting frequency responses of filters and delay lines. Thus, the variation of lengths of connecting transmission lines leads to the appearance of additional minimums of a transmission coefficient (Fig. 6) and the frequency positions of the additional minimums depend on the lengths of these lines (in a general case these lengths may be different ones). This enables to realize much narrower passbands of filters and a higher steepness of their transmission characteristics. Of special interest are narrow-band MSW delay lines with two-port transducers, which can be used in magnetically tunable multi-octave oscillators as an element of a feedback loop of a transistor amplifier. Their specific phase characteristics enable us to realize the principle of the compensation of frequency-dependent phase shifts in an oscillator circuit [4] that results in the elimination of a typical defect of conventional MSW oscillators—frequency jumping during tuning.

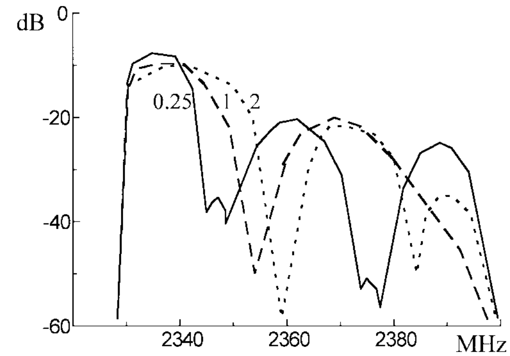

Fig. 6. Power transmission (in decibels) coefficient of a surface MSW filter (delay line) with two-port transducers in a frequency band for different lengths of connecting transmission lines (in millimeters).

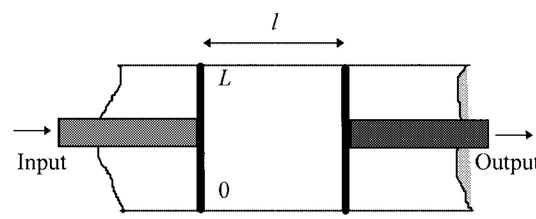

Fig. 7. MSW filter (delay line) with T-type transducers.

V. MSW Filters and Delay Lines With T-Type Transducers

Consider an MSW filter (delay line) with transducers of the T-type, which are connected symmetrically to feeding transmission lines, as shown in Fig. 7. Earlier such multiport structures were not analyzed and the comparison of their performance with devices, having conventional transducers, is of significant interest.

In the case of surface MSWs, with transducers magnetized along their axes, feeding transmission lines do not excite waves and their interaction with a ferrite film can be neglected. To apply a general approach of Section II to calculations of frequency responses of the MSW filter (delay line) with T-type transducers, we take into account a longitudinal symmetry of the device. For that, the transmission coefficient for incident electromagnetic waves can be computed, using (30), as for conventional transducers of a half-length, and with the loading at the first end of the output transducer in the form of the shunt impedance of the output transmission line and the input impedance of the transducer of a half-length.

In Fig. 8, a comparison is shown of frequency responses of surface MSW filters (delay lines) with short-circuited strip-line transducers of the T-type and conventional ones. The structures have the following parameters (notations correspond to Figs. 2 and 7): $h = 0.5$ mm, $b = 75$ μm, $d = 15$ μm, $w = 15$ μm, $L = 4$ mm, $l = 5$ mm, $M_s = 140 \cdot 10^3$ A/m, $\Delta H_0 = 20$ A/m, and $\varepsilon = 10$. The internal magnetizing field is $H_i = 89.2 \cdot 10^3$ A/m. The feeding transmission lines have the characteristic impedance 50 Ω.

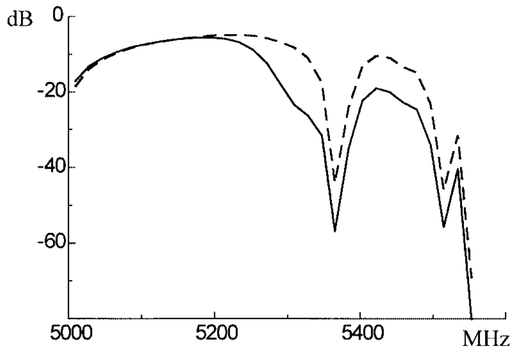

Fig. 8. Power transmission coefficient (in decibels) in a frequency band for a surface MSW filter (delay line) with T-type transducers (solid curve) and conventional transducers (dotted curve).

At lower frequencies of the band, the power transmission coefficients of both devices coincide, while at higher frequencies the filter (delay line) with T-type transducers has a smaller transmission coefficient. This enables us to draw a conclusion that T-type transducers can be employed for the realization of narrow-band devices.

VI. Conclusion

We have suggested a unified approach to the electrodynamic analysis of general multiport structures using MSWs in multilayered ferrite films. The solution holds for an arbitrary direction of a magnetizing field and for arbitrary loadings and types of transmission lines forming a structure. An interaction of magnetostatic modes through the magnetic field of a transducer, which excites them, and an interference of modes at output transducers are taken into account. Derived closed-form expressions for response functions of multiports enable us to use, if necessary, known numerical solutions for MSWs in ferrite-dielectric waveguiding structures and for an electromagnetic field in dielectric-filled transmission lines forming transducers. Obtained solutions provide the basis for modeling of a wide class of adjustable MSW passive and active devices. Numerical examples of the application of a general theory have been given for surface and forward volume MSW filters and delay lines using strip-line transducers. It has been shown that in conventional delay lines experimentally observed passband ripples in frequency dependencies of the transmission coefficient and the delay time are caused by back coupling of the output transducer to the input one. New types of MSW transducers for filters and delay lines (two-port and T-type) have been proposed, and their performance has been analyzed.

References

[1] *Proc. IEEE (Special Section on Microwave Magnetics)*, vol. 76, pp. 121–188, Feb. 1988.

[2] S. V. Zagriadski and A. G. Rezvanov, "Phenomenon of spatial synchronism of electromagnetic and magnetostatic waves and formation of wave beams in ferrite films," in *Proc. 29th Eur. Microwave Conf.*, vol. 1, Munich, Germany, 1999, pp. 20–23.

[3] K. W. Loh and P. R. Emtage, U.S. Patent 4.465.986, 1984.

[4] S. V. Zagriadski, "A new approach to the design of wide-band magnetostatic wave oscillators," in *Proc. 21st Eur. Microwave Conf.*, vol. 2, Stuttgart, Germany, 1991, pp. 1459–1464.

[5] T. Koike, T. Hirao, T. Ito, and T. Nagano, "Frequency control method for microwave oscillators using straight-edge MSW resonators," in *Proc. IEEE Int. Frequency Control Symp.*, 1996, pp. 776–781.

[6] J.-P. Castera and P. Hartemann, "Adjustable magnetostatic surface wave multistrip directional coupler," *Electron. Lett.*, vol. 16, no. 3, pp. 195–196, 1980.

[7] A. Kolodin, P. Kabos, and C. E. Patton, "Frequency and phase selective magnetostatic wave switch," in *IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 3, 1999, pp. 1173–1174.

[8] G. Volluet, "Unidirectional magnetostatic forward volume wave transducers," *IEEE Trans. Magn.*, vol. MAG-16, pp. 1162–1164, May 1980.

[9] A. G. Gurevich and G. A. Melkov, *Magnetic Oscillations and Waves*. New York: CRC Press, 1996.

[10] D. D. Stancil, *Theory of Magnetostatic Waves*. New York: Springer-Verlag, 1993.

[11] P. Kabos and V. S. Stalmachov, *Magnetostatic Waves and Their Applications*. London, U.K.: Cahpman & Hall, 1994.

[12] S. V. Zagriadski and S. Choi, "Excitation and reception of electromagnetic, magnetostatic and spin waves in ferrite films," *J. Electromagn. Waves Applicat.*, vol. 15, no. 12, pp. 1699–1701, 2001.

[13] T. W. O'Keeffe and R. W. Patterson, "Magnetostatic surface-wave propagation in finite samples," *J. Appl. Phys.*, vol. 49, pp. 4886–4895, 1978.

[14] P. R. Emtage and M. R. Daniel, "Magnetostatic waves and spin waves in layered ferrite structures," *Phys. Rev.*, vol. 29, no. 1, pp. 212–220, 1984.

[15] J. P. Parekh and K. W. Chang, "Nondispersive MSFVW propagation in a triple-YIG-film-layered structure," *IEEE Trans. Magn.*, vol. MAG-19, pp. 1868–1870, Sept. 1983.

[16] M. Koshiba and Y. Long, "Finite-element analysis of magnetostatic wave propagation in YIG film of finite dimensions," *IEEE Trans. Microwave Theory Tech.*, vol. 37, pp. 1766–1772, Nov. 1989.

[17] Y. Long, M. Koshiba, and M. Suzuki, "Finite-element solution of planar inhomogeneous waveguides for magnetostatic waves," *IEEE Trans. Microwave Theory Tech.*, vol. 35, pp. 731–736, Aug. 1987.

[18] K. S. Kunz and R. J. Luebbers, *The Finite Difference Time Domain Method for Electromagnetics*. Boca Raton, FL: CRC Press, 1993.

[19] A. Taflove, *Computational Electrodynamics: The Finite-Difference Time-Domain Method*. Norwood, MA: Artech House, 1995.

[20] A. G. Gurevich and G. A. Melkov, *Magnetic Oscillations and Waves*. New York: CRC Press, 1996.

[21] V. V. Rogozin and A. N. Averin, "Field of symmetric strip-line" (in Russian), *Radio Eng.*, no. 2, pp. 73–74, 1980.

[22] R. Marcelli, P. D. Gasperis, M. Rossi, and M. Guglielmi, "Design of tunable magnetostatic volume wave filters with dispersion and bandwidth control," *IEEE Trans. Magn.*, vol. 27, pp. 5471–5473, June 1991.

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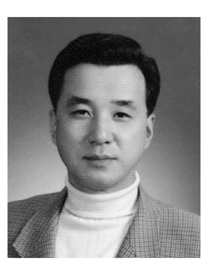

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