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*Review Article*

# Microwave Magnetoelectric Devices

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Tunable microwave magnetoelectric devices based on layered ferrite-ferroelectric structures are described. The theory and experiment for attenuator, band-pass filter and phase shifter are presented. Tunability of the ME devices characteristics can be executed by application of an electric field. This electric tuning is relatively fast and is not power-consuming. The attenuator insertion losses vary from 26 dB to 2 dB at frequency 7251 MHz. The tuning range of 25 MHz of band-pass filter at frequency 7360 MHz was obtained. A maximum phase shift of 30–40 degree at the frequency region 6–9 GHz was obtained.

## 1. Introduction

Multiferroics are materials or structures where different ferroic orders such as ferroelectric, ferromagnetic, and ferroelastic coexist in one material. In multiferroic magnetoelectrics, a dielectric polarization can be induced by an external magnetic field (direct effect), a magnetic moment can be induced by an external electric field (inverse effect). This phenomenon is described as the magnetoelectric effect (ME). The ME effect in single crystals is generally described using the Landau theory by writing the expansion of free energy. The ME effect is observed in two classes of materials: single-phase multiferroic materials possessing simultaneously both ferroelectric and ferromagnetic properties and composites consisting of ferroelectric and ferromagnetic phases. The ME effect in the single-phase materials [1] arises from the long-range interaction between the atomic moments and electric dipoles in ordered magnetic and ferroelectric sublattices. The realizable ME coefficient in single-phase materials is, however, very small (1–20 mV/cm·Oe) and not sufficient for practical applications. Moreover, ME effect in most of these single-phase materials is observed only at low temperatures as either ferromagnetic (or antiferromagnetic) or ferroelectric transition temperature is very low.

Composite materials, on the other hand, provide an alternative strategy which makes use of indirect coupling via mechanical strain between the materials of two different phases: ferroelectric and ferromagnetic (or antiferromagnetic or ferrimagnetic such as ferrites). Mechanical deformation of the magnetostrictive phase results in polarization in the piezoelectric phase. ME effect in composites may be more than several orders of magnitude higher than that in single-phase materials, the latter forming actually the basis for multiferroics. Such composites can be implemented in the form of "multilayers" consisting of alternating layers of the ferroelectric phase and the ferro- or ferri-magnetic phase. Bilayers and multilayers of composites are especially promising due to their low leakage current and superior poling properties. In multilayers, however, the magnetoelectric coupling effect is weakened due to the clamping effect of the substrate, unless the multilayers are fabricated as free-standing membranes.

In the discussion of linear and higher-order magnetoelectrical coupling, we have so far ignored the effects of strain. However, inclusion of strain arising from piezomagnetism and magnetostriction would introduce cross-terms that are proportional to strain and have linear and/or quadratic dependence on magnetic field $H$. Similarly, additional terms would arise from piezoelectricity and electrostriction, having dependencies on electric field $E$. Even mixed terms involving strain, $H$ and $E$, would appear. The effects of strain are significant and in some cases may even dominate which actually desired in laminated two-layer composites. Most ferromagnetic materials exhibit magnetostriction, which describes a change in strain as a quadratic function of the applied magnetic field. This change in the strain may induce polarization in the ferroelectric phase, which is in contact with the ferromagnetic/antiferromagentic/ferromagnetic phase (ideally in atomic registry), through the piezoelectric effect.

For the two-phase systems, the physical properties are determined by the interaction between the constituents as well as by their individual properties. Some effects, which are already present in the constituents may be averaged or enhanced for the overall system. However, the ME effect is among the novel effects that arises from the product properties originating through the interaction between the two phases. The ME effect can be achieved by coupling a piezoelectric (or electrostrictive) material and a piezomagnetic (or magnetostrictive) material via a good mechanical contact. Such a two-phase system may be achieved in the form of composites, laminates, or epitaxial layers. In the direct ME effect, an applied magnetic field induces strain in the magnetostrictive constituent, which can be transferred to the piezoelectric constituent to the extent allowed by the mechanical coupling between the two, and the transferred strain induces an electric polarization in the piezoelectric constituent. The inverse effect is also possible, where an electric field may be applied to the piezoelectric material to induce strain, which is then transferred to the magnetostrictive material, where it affects the magnetization.

Even though the ferromagnetic resonance line broadening problem can be overcome by using layered structures such as laminated composites, the narrowest resonance linewidth is naturally possible in single-crystal material. Therefore, using single-crystal ferrites enhances the resolution in electric field tunability. Using ferroelectrics of single crystal form is also beneficial as they would produce higher strain, and therefore, a larger ME effect will result in ferroelectric/ferromagnetic bilayer structures. As single-crystal YIG has been shown to exhibit the narrowest ferromagnetic resonance linewidth, most of the efforts on ME coupling-based devices have focused on incorporating single-crystal YIG with different types of ferroelectric layers, either in single crystal or ceramic form. The two constituents are bonded using epoxy to form the bilayer structures.

An electric field-tunable yttrium iron garnet (YIG: $Y_3Fe_5O_{12}$) on gadolinium gallium garnet (GGG: $Gd_3Ga_5O_{12}$) substrate—lead zirconate titanate (PZT: $Pb(Zr, Ti)O_3$) microwave resonator based on ferromagnetic resonance for YIG has been demonstrated [2, 3] The finite lateral dimensions of the ferrite film, metallization on the ferroelectric surfaces, and the anisotropy fields in the ferrite may result in slight deviations from the theoretical estimates; however, FMR frequencies were shown to be in good agreement with measurements. The corresponding ME coefficients YIG/PZT structure extracted from magnetostriction-induced magnetic field shifts are $\alpha 1 = \delta H1/\delta E \approx 0.58\,\mathrm{Oe\cdot cm/kV}$ and $\alpha 2 = \delta H2/\delta E \approx 0.88\,\mathrm{Oe\cdot cm/kV}$ ($\delta H$ is the ferromagnetic resonance shift due to an applied electric field $\delta E$), for magnetic-bias field applied parallel and perpendicular to the sample plane, respectively, which are comparable to 1–4 Oe·cm/kV measured in similar YIG/PZT bilayers [4]. The range of the frequency tuned by electric field could potentially be increased by either decreasing the thickness of the GGG substrate or by using a ferroelectric with higher piezoelectric coefficients, such as single-crystal PMN-PT. The frequency shift, which arises due to magnetoelectric interactions, is relatively large for out-of-plane $H$ in accordance with the theory [5, 6]. From the analysis of data, it was concluded that the primary cause of the frequency shift was variation of the effective internal magnetic field in YIG with applied electric field.

To study the resonance effects, for simplicity the ME response is first measured as a function of the static magnetic field, for simplicity, to find the electric field value for peak response. This is the inverse of applications sought in microwave passive components. The static field is then set to this value, and the frequency of the AC magnetic field is swept to observe the resonance effect.

The theoretical analysis of the resonance ME effects may be performed for multilayers in the form of thin disks or thin plates, where the thickness is small compared to the wavelengths of the acoustic modes. The thickness is also assumed to be much smaller than the lateral dimensions of the multilayer (radius, $R$, in the case of disks) so that only the radial modes are considered. The outer surfaces of the multilayer system are coated with metallic contacts and are equipotential; therefore, only the normal component of the electric field is nonzero. Moreover, for thin layers the normal component of the stress tensor may be taken as zero in the entire volume.

Ferrites are an important class of materials for use in microwave passive devices such as phase shifters, circulators, filters, isolators, and resonators [5, 7]. Depending on the frequency of operation, different types of ferrite materials are needed and used. Because ferrites are magnetic dielectric materials (unlike magnetic metals which are conductive) they permit electromagnetic penetration and thus interaction between the electromagnetic wave and magnetization within the ferrite. Ferrite devices permit the control of microwave propagation by a static or switchable DC magnetic field. The devices can be reciprocal or nonreciprocal and linear or nonlinear. What is certain is that their development requires a good deal of knowledge of magnetic materials, electromagnetic theory, and microwave circuit theory.

Miniaturization of microwave devices is one of the fundamental requirements in communication systems. Therefore, small-size and high-performance devices are always necessary to reduce cost and enhance system performance. The advantages of such devices include rapid tuning with an electric field, zero power consumption, noise elimination due to the absence of magnetic tuning, compatibility with integrated circuit technology, and resistance to nuclear radiation. Today there is the diversified production engineering of forming of electric circuits with the small losses, manifesting frequency-dependent properties. It is possible to follow constructive alternatives.

Ferrites are used in tunable microwave and millimeter-wave devices, and the tunability is traditionally realized through the variation of a bias-magnetic field [5, 7, 8]. This magnetic tuning could be achieved over a very wide frequency range, but is relatively slow, noisy and requires high power for operation. Similar devices but with some unique advantages could be realized by replacing the ferrite with a ferrite-ferroelectric composite [9–11]. Such heterostructures are magnetoelectric due to their response to elastic and electromagnetic force fields.

Layered structure based on the magnetostrictive and piezoelectric materials are perspective for designing new microwave devices. The given materials contain magnetostrictive and piezoelectric components and possess magnetoelectric interaction that will allow creating essentially new devices on their basis.

## 2. Theory

*2.1. Coupling between ME Resonators.* For the estimation of device characteristics, we consider the transmission lines and the resonator as a single system that is described by a coupling coefficient. We follow a two-stage procedure for the device design: first, we obtain the coefficients of reflection, transmission, and absorption for the case when an ME resonator is considered as an irregularity in the transmission line. Then, we take into account the number of resonators to estimate the coupling coefficient.

Application of microstrip lines is usually limited to use of the basic wave of type $T$. The ratio for coupling coefficient in this case can be presented in the following:

$$k = \frac{2V\chi''_{+} z_0 \varepsilon}{\pi h^2 \lambda_\nu Z}\left(\operatorname{arc} tg\frac{Z}{z_0\sqrt{\varepsilon}} + \frac{1}{3}\operatorname{arc} tg\frac{3Z}{z_0\sqrt{\varepsilon}}\right)$$
$$\text{with } \chi''_{+} = \frac{8\pi M_0}{\Delta H}, Z = 120\pi \text{ Ohm}, z_0 = 50 \text{ Ohm}, \tag{1}$$

where $V$ is the ME resonator volume, $M_0$ *is* the YIG saturation magnetization, $\lambda_\nu$ is the wavelength in transmission line, $z_0$ is the microstrip impedance, $\varepsilon$ is the substrate permittivity, and $h$ is thickness of substrate.

*2.2. Insertion Loss and Phase Characteristic.* The insertion losses for single-resonator devices are determined by the following expression [12]:

$$L = -20 \log T1, \tag{2}$$

where

$$T1 = \frac{k}{\sqrt{(1+k)^2 + \xi^2}},$$
$$\xi = \frac{H_r - H + \delta H}{\Delta H}. \tag{3}$$

The phase characteristic is determined from the expression for the complex transmission coefficient

$$\varphi = \operatorname{arc} tg\left(\frac{2k\xi}{1 - k^2 + \xi^2}\right). \tag{4}$$

Here, $T1$ is transmission gain of the devices, $H_r$ is the resonance field, $\delta H$ is the FMR line shift due to the electrical field, $H$ is the dc magnetic field, $\Delta H$ is the half-width of FMR line, $\xi$ is the combined detuning, and $k$ is the single-resonator filter coupling coefficient.

The insertion loss is the sum of losses in the ferrite $\alpha_m$ and piezoelectric layer $\alpha_p$, metal conductors $\alpha_{\text{Me}}$, dielectric substrate $\alpha_d$, and due to less-than-ideal coupling between the resonator and transmission lines $\alpha_c$:

$$L = \alpha_m + \alpha_p + \alpha_{\text{Me}} + \alpha_d + \alpha_c. \tag{5}$$

Under ideal conditions, with $\alpha_m = 0,5$ dB, $\alpha_p = 1,0$ dB, $\alpha_{\text{Me}} = 0,5$ dB, $\alpha_d = 0,2$ dB, and $\alpha_c = 0,3$ dB, a total insertion loss as low as 2.5 dB could be realized in the one-cavity system.

*2.3. Two-Resonator Devices.* As model of the multiresonator devices the single-resonator devices joined by a piece of transmission line of nonresonance length, carrying out a role of connection device between parts is accepted.

Thus, the transfer ratio of the cascade-type devices was obtained from a transfer ratio for the single-resonator devices:

$$T2 = \frac{2KK^*}{\Delta}; \quad K = \frac{k}{1+i\xi}; \; K^* = -i\frac{k}{1+i\xi};$$
$$\Delta = (1+K)^2 - K^{*2}. \tag{6}$$

The insertion losses for two-resonator devices are determined by the following expression:

$$L_2 = -20 \log T2. \tag{7}$$

*Here,* $T2$ is transmission coefficient of two-resonator devices, $K$ and $K^*$ are complex coupling coefficients of two-resonator devices, $k$ is coupling coefficient of one-resonator devices.

A single-cavity device suffers from many disadvantages including insufficient pass-band and small out-of-band characteristics. It is, however, possible to improve the characteristics with increasing the numbers of resonators.

## 3. Magnetoelectric Attenuator

The electrically tunable attenuator designed on the base of microstrip transmission line with layer resonator based on YIG film and lead magnesium niobate-lead titanate (PMN-PT) is described. Figure 1 shows typical samples of PMN-PT. The chemical structural formula of PMN-PT is (1-$x$)[$Pb(Mg_{1/3}Nb_{2/3})O_3$]-$x$[$PbTiO_3$]. The PMN-PT is formulated to exhibit high piezoelectric coefficient, large electromechanical coupling coefficient, high dielectric constants and low dielectric losses. As a rule, the piezoelectric coefficient is higher than PZT ceramics, which results in improving band width, sensitivity, and source level in applications.

The unique combination of magnetic, electrical, and ME interactions, therefore, opens up the possibility of reciprocal and nonreciprocal ME signal processing devices. Here, we represent the ME microwave resonance attenuator. With the

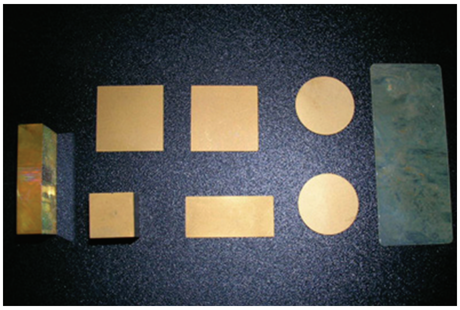

Figure 1: Typical samples of PMN-PT.

purpose of tuning the attenuator parameters in the given paper, the microwave ME effect is used [13, 14]. The effect consists in shift of FMR line on influence of an external electrical field. The used sample represents layered structure, which the magnetic part consists of the YIG thin film placed on the GGG film. A piezoelectric is the thin PMN-PT plate. For obtaining three-layer structure magnetic and piezoelectric parts are jointed together by a thin film of epoxy glue. For calculation of tunable filter parameters, it is necessary to take into account the microwave ME effect.

The base of the attenuator is a microstrip transmission line on dielectric substrate and ME resonator (Figure 2).

Work of the attenuator is based on resonance ME effect phenomena. Applying the control voltage to electrodes of the ME resonator results in a shift of FMR line due to the resonance ME effect, and so electrical tuning is realized.

*Theory and Experiment.* At the first approach, this curve can be written in the following expression:

$$P_{ab}(H) = P_0 \exp\left(-\frac{(H-H_0)^2}{2\sigma^2}\right), \tag{8}$$

where $P_{ab}(H)$ is the absorbed power versus bias magnetic field, $P_0$ is the accepted power at magnetic resonance, $H_0$ is the value of bias magnetic field at the magnetic resonance, and $\sigma$-parameter is determinated from the width of resonance line.

The electric field leads to the shift resonance line

$$\delta H = AE. \tag{9}$$

Substituting (9) into (8) we get

$$P_{ab} = P_0 \exp\left(-\frac{A^2E^2}{2\sigma^2}\right). \tag{10}$$

The insertion loss determined from ME resonator:

$$\begin{aligned} L_R &= 10\lg\left(\frac{P_{out}}{P_{in}}\right) \\ &= 10\lg\left(\frac{P_{in}-P_{ab}}{P_{in}}\right) \\ &= 10\lg\left(1-\frac{P_0}{P_{in}}\exp\left(-\frac{A^2E^2}{2\sigma^2}\right)\right). \end{aligned} \tag{11}$$

As the insertion loss, experimentally was measured parameter $s_{21}$.

The output power $P_{out} = P_{in} \cdot |s_{21}|^2$.

At the magnetic resonance

$$P_{ab} = P_0 = P_{in}\left(1-|s_{21}(H_0)|^2\right). \tag{12}$$

Using (12) we get

$$\frac{P_0}{P_{in}} = \left(1-|s_{21}(H_0)|^2\right). \tag{13}$$

Substituting (13) into (11) we get an equation for the insertion loss in the magnetoelectric resonator in following form:

$$\alpha_R = 20\lg\left(1-\left(1-|s_{21}(H_0)|^2\right)\exp\left(-\frac{A^2E^2}{2\sigma^2}\right)\right). \tag{14}$$

For the calculation was used the following experimental data $A = 4\,\text{MHz/kV}\cdot\text{cm}$, $\sigma = 14\,\text{MHz}$, and $s_{21} = 0,13$.

The insertion loss of attenuator is the sum of losses in the magnetoelectric resonator ($\alpha_R$) and piezoelectric layer $\alpha_p$, metal conductors ($\alpha_{Me}$), dielectric substrate ($\alpha_d$), and due to nonideal coupling between the resonator and transmission lines ($\alpha_c$)

$$L = \alpha_R + \alpha_p + \alpha_d + \alpha_{Me} + \alpha_c. \tag{15}$$

Under ideal conditions, with $\alpha_R = 0,5\,\text{dB}$, $\alpha_p = 1,0\,\text{dB}$, $\alpha_{Me} = 0,5\,\text{dB}$, $\alpha_d = 0,2\,\text{dB}$, and $\alpha_c = 0,3\,\text{dB}$, a total insertion loss as low as 2.5 dB could be realized in the one-cavity system.

If the attenuator is worked on a resonance frequency (Figure 3), thus the attenuator insertion losses on influence of control electrical field vary from 26 dB to 2 dB (Figure 4).

The obtained results proved the possibility for designing resonance ME attenuator with electrical control.

Thus, the microwave attenuator on the basis of composite ferrite-piezoelectric resonators are offered. Calculation of attenuator characteristics with single-crystal ME resonators on the trilayer GGG/YIG/PMN-PT is performed.

The electric tuning by ME device parameters opens new capabilities for their application and manufacturing of devices with high-speed operation, small size, and integral technologies.

## 4. Magnetoelectric Band-Pass Filter

A microstrip band-pass filter is designed and characterized [15]. The filter is based on ferromagnetic resonance (FMR) of ferrite component. The device operating at 5–10 GHz can be tuned over a wide frequency band with a bias magnetic field and over a narrow band with a voltage applied across piezoelectric component. The voltage tuning of the device is possible through ME interactions that are mediated by mechanical deformation and manifests as a shift in FMR. Data on tuning range, insertion loss, and device characteristics are presented for filters with single- and double-ME resonators.

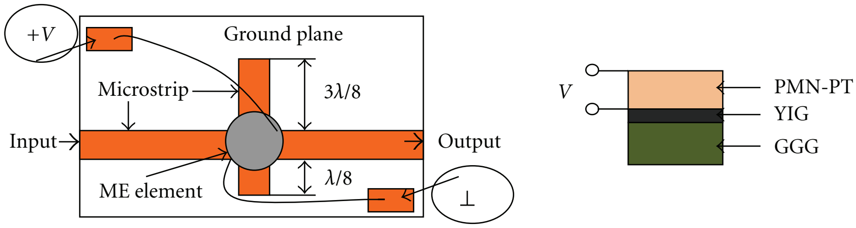


Figure 2: Design of microstrip ME attenuator and ME resonator.

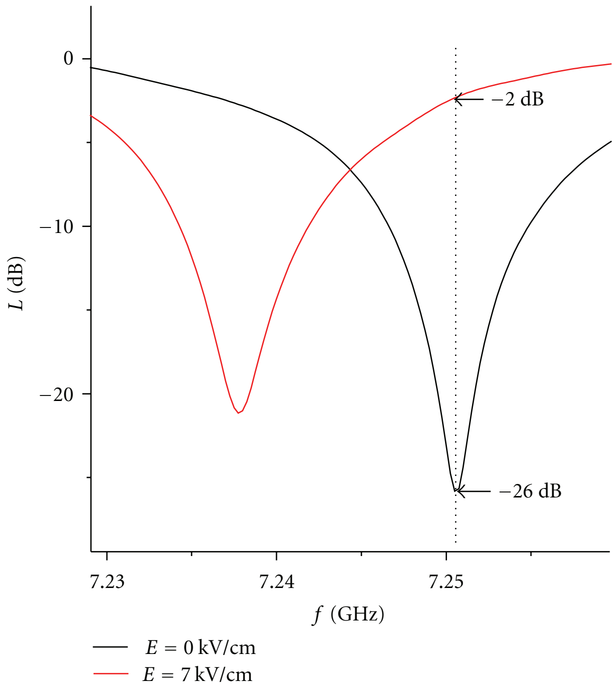


Figure 3: Experimental curves of insertion loss versus frequency. YIG [111] thickness is 110 $\mu$m; diameter is 2,5 mm; PMN-PT thickness is 0,5 mm; diameter is 6 mm; field $H = 1910$ Oe; field is parallel to plane of sample; central frequency is 7251 MHz.

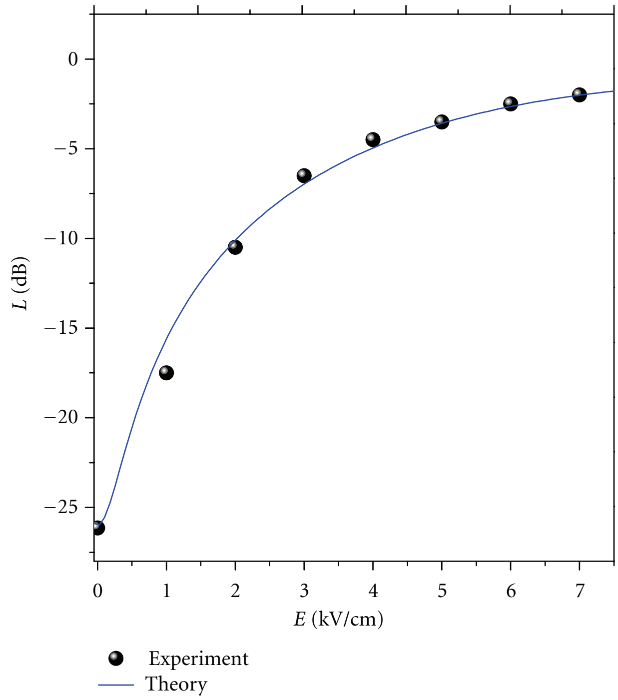


Figure 4: Comparison of experimental and theoretical data. YIG [111] thickness is 110 $\mu$m; diameter is 2,5 mm; PMN-PT tickness is 0,5 mm; diameter is 6 mm. Field $H = 1910$ Oe; field is parallel to plane of sample; average is 30; central frequency is 7251 MHz.

As objects used magnetostrictive component such as YIG and piezoelectric component such as PZT or PMN-PT. Designing band-pass filters utilizing ferrite-piezoelectric structure, which is ME involves calculations of the energy's coupling of the first ME resonator and of coupling of energy between resonators. Tunable resonators are widely used in microwave circuits as the frequency controlling element in reference oscillators, as the tunable resonator tank for voltage-control oscillators (VCOs), and as building elements for tunable filters. Traditional MMIC varactor-tuned resonators need to use GaAs MOSFETs or a *pin* diode as tuning elements, which will lead to high insertion loss and also small tuning range [16]. Microelectromechanical systems (MEMSs) technology has been successfully applied in developing a tunable resonator to achieve large tuning range from 3.5 to 7 GHz [12]. However, large difference in $S_{11}$ response is found in the whole tuning range. Tunable ME resonators are used in microwave range. The resonant frequency of the ME resonator can be shifted at a tunable range from 5 to 10 GHz. There are also various other considerations of importance such as temperature properties, unwanted mode suppression, and signal-handling capability. What is not available is adequate information on coupling between ME resonators and means of designing filters for the minimum excitation and coupling of unwanted modes. One of the goals of this paper was the determination of those physical parameters which affect the coupling of the fundamental mode.

We recently studied ME interactions at FMR in layered ferrite-ferroelectric composites [17, 18]. The studies were done on bilayers of single crystal YIG films and PMN-PT. Microwave ME measurements at 9.3 GHz were performed using a traditional FMR spectrometer. An electric field $E$ applied to the sample produced a mechanical deformation in the piezoelectric phase that in turn was coupled to the ferrite and manifested as a shift in FMR profiles. Profiles of

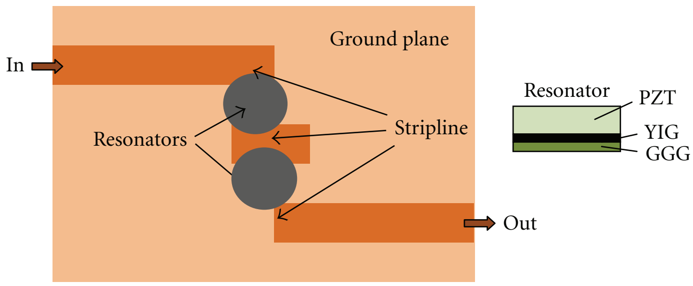


Figure 5: Schematic diagram showing a dual-cavity magnetoelectric filter and the ME resonators.

FMR absorption versus bias magnetic field $H$ were obtained for a series of electric voltage across PMN-PT. With the application of $E$, we measured a shift in the profile by $\delta H$ and the ME constant $A = \delta H/E$ varing over the range 1/5 Oe·cm/kV, depending on the YIG film thickness. Since 1 Oe of field shift corresponds to a frequency shift of 2.8 MHz, frequencies of FMR-based microwave devices can be tuned with $E$. For a nominal field of 30 kV/cm, an YIG/PMN-PT resonator, for example, can be tuned by 80/400 MHz. Thus, the ME effect would facilitate rapid voltage tuning for any FMR-based microwave device. Other advantages of microwave ME devices include miniaturization, near-zero power consumption, noise reduction and compatibility with integrated circuit technology.

This paper is on the fabrication and characterization of an electric field tunable microwave band-pass filter with single- and dual-ME resonators. Since the frequency of operation is 5/10 GHz, we choose low-loss single-crystal YIG films for the magnetic phase of the resonator and polycrystalline lead zirconate titanate (PZT) for the ferroelectric phase. Our studies indicate very good tunability due to a strong ME interaction and an acceptable insertion loss. Theoretical models account very well for the observations.

*Filter Design.* Both single- and dual-ME cavity filters were studied. Microstrip transducers fabricated by on substrates of foil-clad microwave material FLAN-10 (permittivity at 10 GHz, not more than 10,0 ± 0,8; tangent to dielectric loss angle at 10 GHz, not more than 0,0045; sheet thickness (including copper foil) from 1,0 ± 0,1 mm). It is a sheet microwave material manufactured from the composition on the basis of filled polyphenylene oxide with electrolytic galvano-resistant 35-micron copper foil bonded to both sides. It possesses a high stability value of dielectric permittivity and low dielectric losses in the microwave range. The base of the band-pass filter is a microstrip transmission line on dielectric substrate and ME resonators The single-cavity ME filter consisted of a 1 mm thick ground plane, input and output microstrips of nonresonance lengths, and an ME element. The microstrip transducers, 1 mm in width and 18 mm in length, were separated by 1.5 mm. The input-output decoupling is determined by this gap between the microstrips. Power is coupled from input to output under FMR conditions in the ME element. A dual-cavity filter shown in Figure 5 consisted of a 1 mm thick ground plane, input and output microstrips of nonresonance lengths equal 15 mm, section of a nonresonance length microstrip (3 mm), carrying out a role of an coupling element between two resonators, and a two-ME element. Width of all lines is 1 mm, gaps between microstrips are 1 mm also. The ME element consisted of epitaxial YIG film bonded to PZT. A YIG film grown by liquid-phase epitaxy on a (111) GGG was used. The ME element was made from YIG film 110 $\mu$m thick on GGG substrate 210 $\mu$m thick with diameter 1,5 mm and PZT plate (dimensions $4 \times 0{,}9 \times 0{,}5\ \mathrm{mm}^3$) with electrodes.

The film saturation induction $4\pi M$ of 1750 G, and FMR line width of 1 Oe. A PZT plate was initially poled at oil by heating up to 110°C within 5 hours and cooling back to room temperature in an electric field of 4 kV/mm perpendicular to the sample plane. The layered structure was made by bonding the YIG film surface to PZT with thin layer of ethyl cyanoacrylate, a fast-dry epoxy. The layered structure was placed between the transducers and was subjected to a field $H$ parallel and perpendicular to the sample plane.

*Results.* The device characterization was carried out with a vector network analyzer (PNA E-8361). An input continuous wave signal $P_{\mathrm{in}}(f) = 1\ \mathrm{mW}$ was applied to the filter. The frequency $f$ dependence of the insertion loss $L$, that is, the transmitted power through the ME element, was measured at 4/10 GHz as a function of $H$ and $E$ applied across the PZT layer.

First, we are interested in the frequency dependencies of the insertion losses were measured as a function of magnetic field $H$ (applied in the plane ferrite layer) and the electrical field $E$ (applied across the piezoelectric layer). Next, we measured other characteristics of filter like bandwidth of pass-band (level of −3 dB), stop-band, and shape factor—$K_p = \mathrm{BW}_{\text{stop-band}}/\mathrm{BW}_{\text{pass-band}}$, $Q$-factor.

The insertion loss dependencies of the frequency (Figure 6) were measured for one-resonator filter as compared to two-resonator filter. The data are for a central frequency $f = 7{,}36$ GHz (corresponding to $H_0 = 1915$ Oe, $H$ is parallel to plane of resonator).

Characteristics of one-resonator and two-resonator band-pass filters are presented in Table 1.

Figure 7 shows the dependence of the frequency shifts $\delta f$ on YIG/PZT structure versus intensity of $E$ for case that magnetic field is parallel and perpendicular to the plane of resonator.

TABLE 1: Characteristics of one-resonator and two-resonator band-pass filter.

| Type of filter | $f_{\text{central}}$ (MHz) | $\text{BW}_{-3\,\text{dB}}$ (MHz) | $\text{IL}_{\text{pass}}$ (dB) | $\text{IL}_{\text{stop}}$ (dB) | $K_p$ | Q |
|---|---|---|---|---|---|---|
| One-resonator YIG filter | 7356 | 8 | 3.0 | 12 | 3.0 | 660 |
| Two-resonator YIG filter | 7355 | 16 | 8 | 24 | 3.0 | 450 |
| Two-resonator YIG/PZT filter | 7370 | 30 | 11.4 | 25 | 4.0 | 240 |

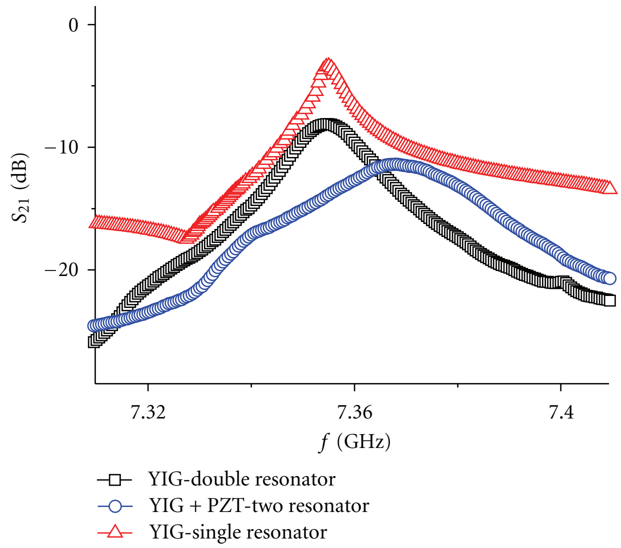


FIGURE 6: One-resonator filter as compared to two-resonator band-pass filter.

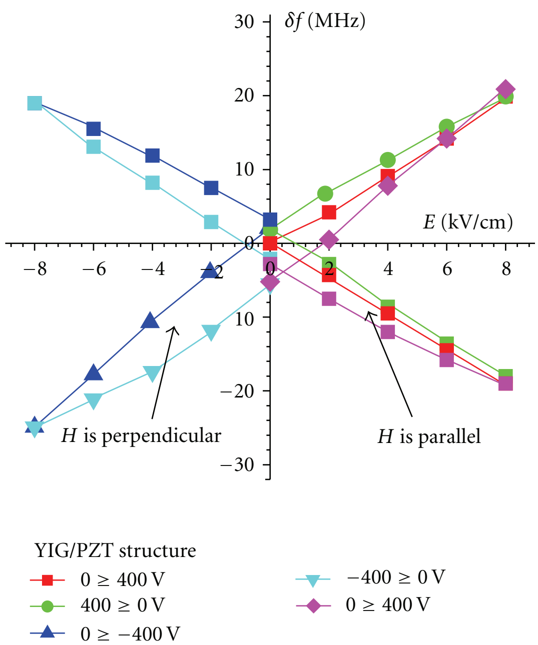


FIGURE 7: The dependence of the frequency shifts $\delta f$ versus intensity of $E$ in case that magnetic field is parallel and perpendicular to the plane of resonator.

One observes a near-linear variation in $\delta f$ with small hysteresis. The observable linear dependence of the shift may be explained by internal distortions of YIG films and indicates the necessity of application in practice of thin YIG films.

Thus, the electrically tunable band-pass filter designed in microstrip transmission line and with layer resonator based on YIG-film and PZT is described. First, we measured the insertion loss as a function of magnetic field $H$ (applied in the plane and across the ferrite layer) and the electrical field $E$ (applied across the piezoelectric layer). A minimum in the insertion loss is 3 dB for one-resonator YIG filter and 8/11 dB for two-resonator band-pass filter in the frequency region 5/10 GHz. The insertion loss is up to 25 dB for out-of-band frequencies.

Next, we measured the frequency shift $\delta f$ on YIG-PZT layered structure for different electric field $E$ (magnetic field $H$ is parallel and perpendicular to plane of resonator). In magnetoelectric composite frequency, tuning is realized through the variation of the applied electric field that creates effect of magnetostriction. Tunability of the devices characteristics in the chosen working range can be executed by application of an electric field. This electric tuning is possible in a narrow frequency range, but it is relatively fast and is not power-consuming. It will allow carrying out tuning in a range of 25 MHz. The electric tuning would facilitate high-speed operation, small size, and compatibility with integrated circuit technology.

## 5. Magnetoelectric Phase Shifter

Traditional ferrite phase shifter use magnetic tuning systems that are slow, demand high power, and are not miniature in size. Here, we discuss the design and characterization of a new electric field tunable phase shifter based on ferrite-piezoelectric layer composite. The electrical control of the phase shifter is realized through microwave magnetoelectric effect [19]. The phase shifter is capable of rapid tuning and miniature in size. This microwave phase shifter is of interest for phased array antenna system.

*Experiment.* Here, we presented the design and characterization of a new electric field tunable phase shifter based on ferrite-piezoelectric layer composite. The phase shifter is capable of rapid tuning and miniature in size. This microwave phase shifter is of interest for phased array antenna system. The base of the phase shifter is a microstrip transmission line on alumina substrate ($\varepsilon$ = 9.8, thickness = 1 mm) and ME resonator. At researches, there were used ME

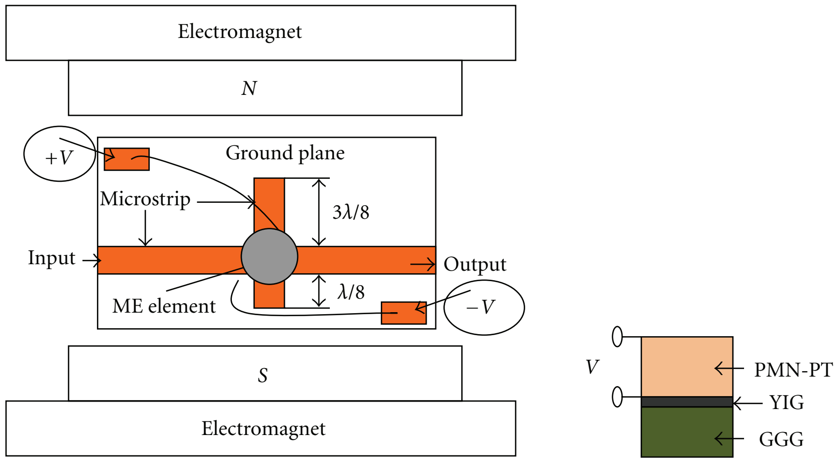


FIGURE 8: Schematics of a ME microwave phase shifter and ME resonator of YIG film on GGG bonded to PMN-PT.

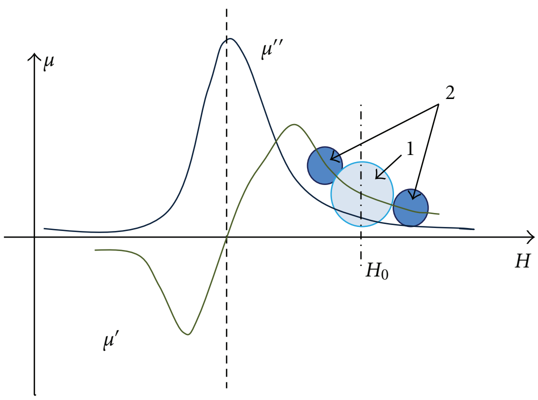


FIGURE 9: Typical ferromagnetic resonance regime for YIG showing the real and imaginary parts of the permeability as a function of bias field $H_0$. The change in the permeability and phase with $H_0$ or $E$ is linear in region 1 and is nonlinear in region 2.

composite consisting of epitaxial single-crystal (111) YIG films on GGG substrates and 0.44 mm thick (001) PMN-PT. Metal electrodes (200 nm in thickness of gold and 30 nm in thickness of chromium) were deposited on PMN-PT for electrical contacts, and the crystal was initially poled by heating to 373 K and cooling it back to room temperature in $E = 2$ kV/cm. A thin layer (<0.08 mm) of an epoxy, ethylcyanoacrylate, was used to bond YIG to PMN-PT. Circular polarization of the magnetic field in volume of the magnetoelectric resonator results from using the microstrip stubs of one eighth and three eighth wavelength. A bias field corresponding to FMR is applied to the ME resonator. Electrical tuning is realized with the application of a control voltage to electrodes leading to a shift of FMR line.

The ferrite film used is a 26.8 $\mu$m thick disk with diameter 4 mm. A PMN-PT plate with diameter 6 mm was used as piezoelectric element. Device characterization was carried out with a vector network analyzer PNA series E8361.

Schematics of a ME microwave phase shifter and ME resonator are represented in Figure 8.

Basis of a design of the phase shifter is the microstrip line on a substrate from an alumina and a magnetoelectric composite, serving by the resonator. In volume of the resonator by means of microstrip loops of $\lambda/8$ and $3\lambda/8$ wave lengths, a circular polarization of a magnetic field is created. In Figure 7, the design of the microstrip microwave phase shifter is presented.

The resonator is installed in area of circular polarization of a magnetic field, constant field $H$ in parallel to a sample plane. For reduction of losses the working point gets out on a dispersive curve outside of resonance area.

Work of the phase shifter is based on the following: for a choice of the working point laying on a dispersive curve, to the resonator on frequency close the external magnetic field is put to a resonance. Under influence of the operating voltage put to electrodes, located at end faces of the resonator, owing to microwave effect there is a shift of line FMR and electric management in parameters of the phase shifter is realized.

Calculation is lead for the phase shifter with resonant frequency of 6/9 GHz. The ME composite is included as heterogeneity in area of circular polarization of a magnetic making microwave of a field. Change of a phase corner is achieved by a choice of strong not mutual connection of the resonator with a line of transfer. Thus, regulation of a phase corner is achieved by change of an operating voltage on the electrodes put on the resonator.

The working condition of the phase shifter will be chosen on dispersion curve of permeability at external magnetic field. Figure 9 shows typical permeability $\mu$ versus $H$ profiles for YIG; FMR is expected for $H = H_r$. We choose $H_o > H_r$, to obtain low insertion loss and a linear or nonlinear variation of $\mu'$ with $H$.

*Results.* Representative data on *phase* versus *frequency* for different frequency and magnetic field $H$ are shown in Figures 10 and 11. We measured the phase characteristics as a function of the electrical field $E$ (applied across the piezo-electric layer) at frequency region 6/9 GHz. A magnetic field

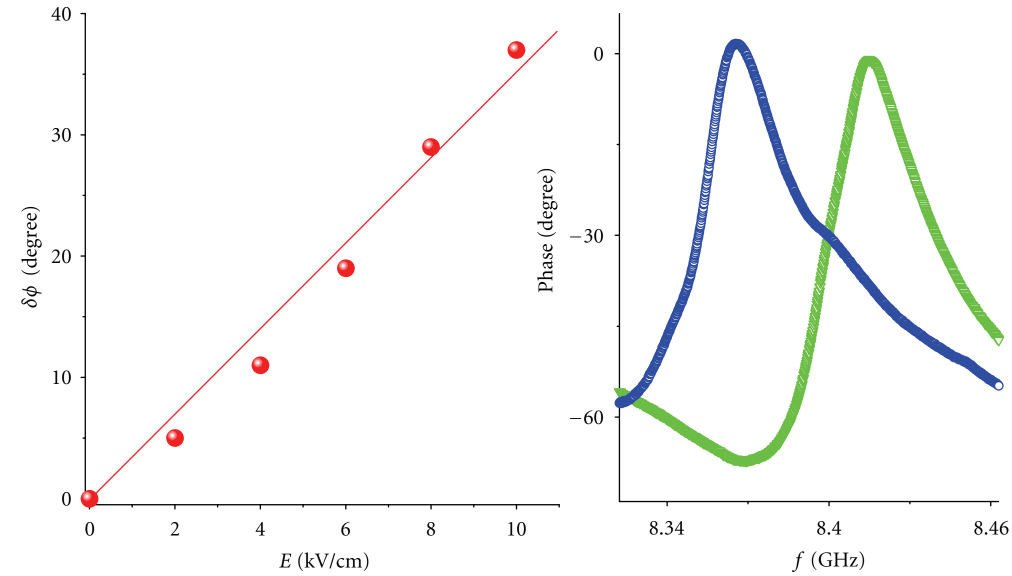


FIGURE 10: Differential phase shift versus electric field $E$ (left) and typical view of phase characteristics versus frequency (right). $H$ is parallel to sample $H_0 = 2277$ Oe; $L = -4$ dB; $f_0 = 8420$ MHz YIG 26.8 $\mu$m; $\phi = 4$ mm; $t_{GGG} = 440\,\mu$m; PMN-PT 0.5 mm; $\phi = 6$ mm.

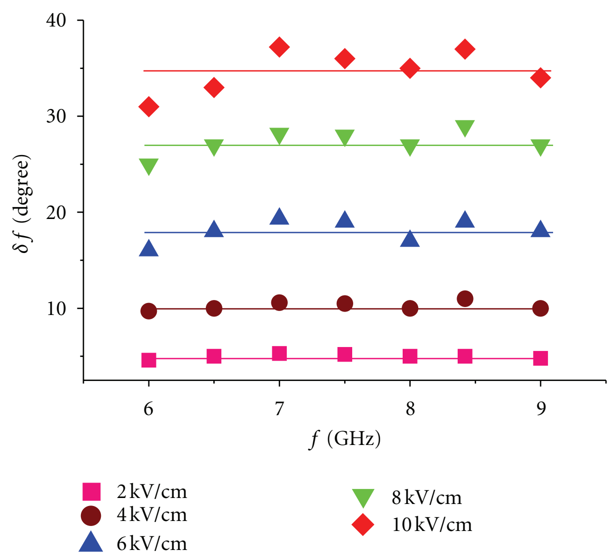


FIGURE 11: Differential phase shift versus frequency at different electric field $E$. $H$ is parallel to sample $L = -3/4$ dB; YIG 26.8 $\mu$m; $\phi = 4$ mm; $t_{GGG} = 440\,\mu$m; PMN-PT 0.5 mm; $\phi = 6$ mm.

$H$ was used for choice of working point. We used parallel magnetic field. The loss was 3/4 dB at the frequency region.

The data in Figures 10 and 11 constitute demonstration of the electrical tunability of the phase shift in a microwave ME microstrip device. The tunability is due to the influence of mechanical deformation on the piezoelectric phase. A near-linear variation in phase with $E$ is evident from the data. A maximum phase shift is 30/40 degree at the frequency region 6/9 GHz. Obtained data allow to speak about an opportunity of practical application of our device. Optimization of the material, both ferrite and piezoelectric phase, and device parameters will allow a further reduction in the insertion loss and provide a high electric tunability.

Thus, we designed, fabricated, and tested microwave ME phase shifter in a standard thin film passive technology. Experimental results showed that 30/40 degree phase shift was achieved at ranging the 6/9 GHz over a range from 0 to 10 kV/cm. Our technology can significantly reduce the cost of phased array systems, where the phase shifter circuit is a large fraction of the overall antenna cost.

## 6. Conclusion

Ferrite-piezoelectric heterostructures are ideal for studies on wideband magnetoelectric interactions between the magnetic and electric subsystems that are mediated by mechanical forces. Such structures show a variety of magnetoelectric phenomena including microwave ME effects. The phenomenon can be used for creating electrical tuning microwave ME resonators and devices on their basis. The specific focus is on ME effects when electromagnetic modes close to FMR are excited in the ferrite. In the presence of an electric field, the piezoelectric deformation in the bilayer structures will manifest as an internal magnetic field in the ferrite. Tunability of the ME devices characteristics in the chosen working range can be executed by application of an electric field. This electric tuning is relatively fast and is not power-consuming.

The central problem of microwave technology is the design of receiver, transmitter, and receiver-transmitter modules, phased array system (PAS) with specified characteristics. They must address diverse requirements of PAS in sensitivity, power, frequency range, and so forth. The use of electromechanical, ferromagnetic, and magnetoacoustic resonances, magnetic dipole and electric dipole transitions in these structures allows the design of new microwave

ME devices [6, 20] and manufacture on them through various modules [21]. The results of our research show promise reviewed microwave ME devices for phased array modules [22]. Development and application of new layered ferrite-piezoelectric structures with advanced properties will significantly improve the performance of modules and PAS as a whole.

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