EFFICIENT MODELING OF MICROACOUSTIC STRUCTURES IN TERMS OF ELECTROMAGNETIC FIELD CONCEPTS

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Abstract – By way of two examples it is demonstrated that basic electromagnetic field theory may be efficiently applied to model anisotropic microacoustic phenomena. A phased array antenna model is used to analyze the guided–to–radiation mode conversion losses in metalic gratings, and microacoustic strip waveguides are characterized by use of the classical ray zigzag model.

1. Introduction

In recent years, surface acoustic wave (SAW) filters and resonators have become well–established components in modern mobile radio systems [1]. Matched filters (fixed–coded [2] as well as programmable [3]), sensors and radio sensors [4] are among the SAW devices which may play a key role in the upcoming commercial RF spread spectrum and radar systems. The SAW design usually combines methods based on signal theory, network theory, circuit theory, and microacoustic field theory. Since the advent of SAW technology, nearly the complete spectrum of electromagnetic methods and concepts have been applied to microacoustic problems [5]. However, full–wave analysis methods (e.g. modal analysis, method of moments, ...) as well as numerical methods (e.g. finite elements, finite differences, ...) incorporate immense computational efforts because the elastic fields and waves in piezoelectric (and thus anisotropic) media are governed by the tensor–based mechanical equations of motion which are in general coupled to Maxwell's equations. Even when analyzing simple boundary conditions such as free, metalized and periodically loaded surfaces [6], these methods require for too much computer resources to be applicable in a SAW design procedure. Nevertheless, basic electromagnetic concepts such as coupling–of–modes, Green's function, and Floquet analysis techniques are very powerful methods when the microacoustic problem can be formulated in a simplyfied way. In what follows, we will demonstrate the feasibility of two other classical and simple electromagnetic techniques.

2. Modeling of Mode Conversion Loss in Arbitrary Metalic Gratings

Most SAW devices incorporate reflecting gratings consisting of an array of metalic strips (fingers) having arbitrary positions and finger widths. The geometry of such reflectors can be rather complicated because weighting of the widths and positions of the fingers is often introduced in order to improve the stopband characteristic [7], and for broadband operation chirped gratings are used. One of the pertinent loss mechanisms in such gratings deposited on piezoelectric substrates is the surface–to–bulk wave conversion which occurs above a certain threshold frequency. This means that a considerable part of the traveling acoustic energy is radiated into the bulk causing an increase of the SAW device insertion loss. This effect plays an even more important role when the SAW device is operated at higher spatial harmonics which is often the case with center frequencies higher than 2.5 GHz. Any efficient design method for low–loss SAW filters has to account for that conversion loss.

A computationally simple method to evaluate the excitation of bulk waves has been introduced by *Fleischmann* and *Skeie* [8] who define bulk wave sources corresponding to the grating geometry. The sources are considered as elements of a phased array antenna as is shown in Fig. 1. The surface–to–bulk wave conversion loss is derived from the far field integral of the radiated power.

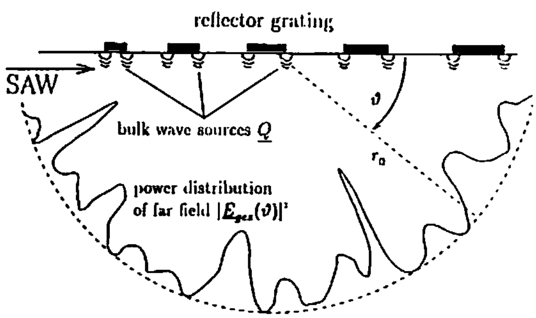

Fig. 1: Basic phased array antenna model

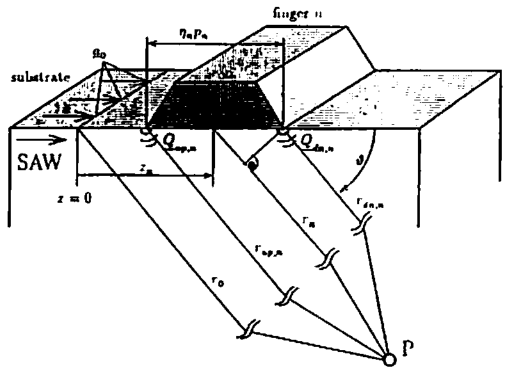

Fig. 2: Extented phased array antenna model (single finger of a grating)

In [8] only one isotropic bulk wave source located in the center of the strip was assumed and the anisotropy of the substrate was not accounted for. Moreover, *Fleischmann* and *Skeie* consider only uniform gratings. The source amplitudes are assumed to be constant and the source phases simply correspond to the distance covered by the surface wave. We found that most of these approximations can be dropped and worked out an extended phased array model. As is seen in Fig. 2, we introduce at first two bulk wave sources at the up– and down–step edges of each strip to model the physical conversion coefficients relating the incident surface wave to the sources at respectively the up–step and down–step. By doing so, we can account for varying finger widths. Next, we include the anisotropy of the substrate in the form of bulk wave velocity look–up tables with the angle in the sagittal plane as the argument. The far field integral is performed with respect to this angle. Furthermore, the sum formula describing the linear superposition of the far field (which is a closed–form expression in the case of a uniform grating) is evaluated numerically for each angle step of the integration. Thereby we are able to deal with arbitrary non–uniform grating geometries. As a last measure, we derive the source distribution exactly in modulus and phase from a network theory formalism which accounts for the relevant physical phenomena, and we describe the surface wave amplitudes in terms of scattering parameters. By doing so, we are able to predict the bulk wave losses at stopband frequencies very precisely. Our algorithm is as follows. The bulk wave loss is calculated from the far field integral of the radiated power

$$P_b = \int_0^{\pi} |E_{tot}(v_b(\Theta))|^2 \, r_0 \, d\Theta \, , \tag{1}$$

with the total complex wave amplitude E_{tot} being a function of the angle dependent bulk wave velocity $v_b(\Theta)$. Using the common far field approximations we then compute the complex bulk wave amplitude as a sum over all N strips as a function of strip height h, strip width $\eta_n p_n$ (η_n: local metalization ratio; p_n: local pitch) and frequency $f = v/\lambda$ (v and λ are respectively velocity and wavelength of the surface wave) as (n=1,2,...N)

$$E_{tot} = hf/(v\sqrt{r_0}) \sum_n \left(Q_{up,n}\, e^{j\varphi_{up}} + Q_{dn,n}\, e^{j\varphi_{dn}}\right) , \tag{2}$$

with $Q_{up,n}$ and $Q_{dn,n}$ being the complex bulk wave sources derived from the surface wave amplitudes and the corresponding conversion coefficients. The phase terms φ_{dn}

and φ_{up} are

$$\varphi_{up}^{dn} = \frac{2\pi}{\lambda_b(\Theta)} \left(z_n \pm \frac{\eta_n p_n}{2}\right) \cos\Theta \, . \tag{3}$$

As a numerical example, we will present the simulation results for a chirped reflector grating on an STX–quartz substrate incorporating 100 strips and having a 5 % bandwidth. We consider only the slow shear wave type because of its low cut–off frequency which affects the characteristics of chirped gratings even at the first stopband. Figs. 3 and 4 show the reflection factor and the conversion loss respectively near the first stopband and in a broader frequency range.

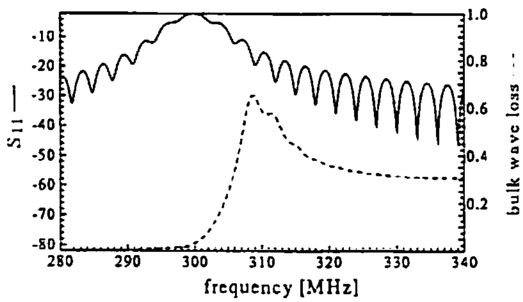

Fig. 3: Simulated reflection factor and bulk wave conversion loss

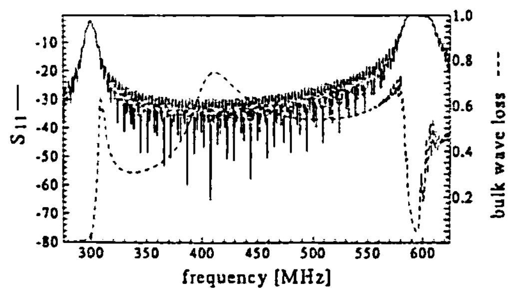

Fig. 4: Broadband simulation covering the first two stopbands

A new effect which we have discovered is an additional conversion maximum between the two stopbands caused by the anisotropy of the substrate. This maximum occurs at 410 MHz. Furthermore, at the second stopband an extremely low conversion loss due to the strongly decreasing surface wave amplitude has been found.

3. Modeling of Strip Waveguides

SAW waveguide convolvers incorporate the microacoustic strip waveguide as an essential part. A very accurate knowledge of the waveguide mode dispersion and attenuation is indispensable for the convolver design [3]. A very thin metal strip of width w and height h plated on the piezoelectric substrate acts as a waveguide because the surface–wave velocity for a metalized surface is slightly slower than that for a free surface. The velocity reduction is mainly due to the short–circuiting of the electric field associated with the surface acoustic wave. The guiding mechanism is basically similar to that of an electromagnetic waveguide comprised of a dielectric strip deposited on a dielectric substrate. The microacoustic strip corresponds in the electromagnetic case to a medium with a higher index of refraction. The wave field is guided by the strip, decaying away from it in all transverse directions. The microacoustic waveguide can support a series of propagating modes, each of which is dispersive. We present a computationally simple and efficient model which is based on a scalar potential representation of the SAW mode. Anisotropy in both the *low–velocity* strip region and the neighboring *high–velocity* unmetalized regions is fully included by using the accurately measured slowness of the piezoelectric substrate. The slowness is defined as the reciprocal of the propagation–angle–dependent phase velocity $v(\phi)$. Waveguide attenuation is accounted for by the introduction of complex propagation constants.

Let the crystal surface and the wave propagation direction be the x–z–plane and the z–direction, respectively. The starting point of our theory is the scalar potential $\psi(x,z)$ obeying the wave equation

$$\partial^2\psi/\partial x^2 + \partial^2\psi/\partial z^2 - \gamma_r\psi = 0 \quad (4)$$

with γ_r being the complex propagation constant. The dispersion relation is derived as

$$\tan(m\pi/2 + j\gamma_{xl}w/2) = -j(\gamma_{xh}/\gamma_{xl})(v_h/v_l)^2 \quad (5)$$

with indices h and l denoting respectively the high- and low-velocity regions, and with (i=h,l)

$$\gamma_{xi}^2 = \gamma_{ri}^2 - \gamma_z^2;\ \gamma_{ri}(\varphi_i) = \alpha_i + j\omega/v_i(\varphi_i);\ \varphi_i = \arctan(\gamma_{xi}/\gamma_z). \quad (6)$$

The zigzag-angle φ is a complex number but can be approximated as purely real in the low-velocity case and as purely imaginary in the high-velocity case.

Careful test-chip measurements of dispersion and attenuation of SAW modes both on free and on metalized surfaces have been made, parameterized, and included in the model. A very good agreement between computed and measured characteristics has been obtained for the phase nonlinearity (deviation from linear phase) as well as for the attenuation of strip waveguides on YZ-$LiNbO_3$ having different values of width w and metalization height h. The phase nonlinearity can be predicted with an accuracy of up to $\pm 0.5^0/\mu s$ over a relative bandwidth in the order of 50%.

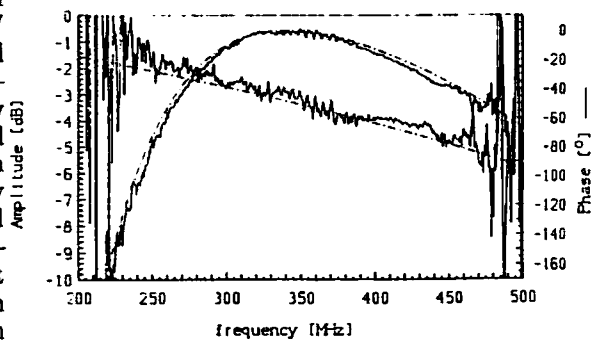

Fig. 5: Measured and computed phase nonlinearity and attenuation of an 8 μs-waveguide (w=29.3 μm, h=30 nm)

4. Conclusion

The present work demonstrates by way of two examples the usefulness of computationally simple electromagnetic field concepts for the modeling of microacoustic phenomena. The models require very short computing times so that they are suitable to be implemented in the overall design procedure of modern SAW devices.

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