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**“Microwave and THz Technologies and
Wireless Communications”**

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Preface

The International Conference “*Microwave and THz Technologies and Wireless Communications*” (IRPhE’2012) was held in Yerevan, Armenia, from October 16 to October 17, 2012. The IRPhE’2012 was directed to the revival of the traditional conferences organized by the Institute of Radiophysics and Electronics, National Academy of Sciences of Armenia, since 1968. In the framework of the Conference two sections were organized: “*Terahertz and Microwave Technologies*” and “*Wireless Communications*”.

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Terahertz Time-Domain Spectroscopy (THz-TDS) Applied to Non-destructive Dielectric Quality Evaluation of Industrial Materials

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The terahertz time-domain spectroscopy (THz-TDS) technique has recently been in progress with advantages over other conventional techniques (Raman and FTIR) and with a wide range of research and industrial applications. A versatile compact THz-TDS instrument has been newly developed with the advantages of wide wavenumber coverage and high dynamic range. The versatile THz-TDS measurements are now being applied for far-infrared spectrum measurements on industrial functional materials, polymorphous organic compounds, biomolecules, and crystalline morphology of industrial products. The THz-TDS investigation is also being applied for the quality control in pharmaceutical industry. Here, an overview of the newly developed instrument performances and some of the latest results of novel applications focusing on the THz-TDS non-destructive evaluation of dielectric properties of industrial functional material and also pharmaceutical products are described.

1. Introduction

The generation of coherent terahertz (THz) radiation through the femtosecond pulse laser irradiation on a photoconductive antenna has been effectively utilized for the promising new spectrometric technique of the terahertz time-domain spectroscopy (THz-TDS) [1]. The THz-TDS technique [2] with the advantage of better signal-to-noise-ratio than the conventional spectrometric techniques of Raman and FTIR makes it possible to measure not only the spectrometric transmission intensities $T(\omega)$ but also the intrinsic phase shifts $\Delta\phi(\omega)$ of propagating THz radiations through within a sample specimen. The exact measurements of both $T(\omega)$ and $\Delta\phi(\omega)$ make direct estimation of the real

part $\epsilon'(\omega)$ and the imaginary part $\epsilon''(\omega)$ of complex dielectric constant $\epsilon^*(\omega)$ free from the uncertainty caused by the Kramers-Kronig analysis. The intrinsic phase shifts $\Delta\phi(\omega)$ also enables to make analytical estimation of the dispersion relations for various elementary excitations coupled with the propagating THz radiations. Thus the THz-TDS measurements [3][4] are effectively applied to the far-infrared spectroscopic investigations of intermolecular vibration modes, crystalline phonon modes, crystalline morphology, photonic dielectric structures and glassy material boson peaks. The THz-TDS technique has recently been in progress with a wide expansion of fundamental research and industrial investigation [4] applications.

A versatile compact THz-TDS instrument has been newly developed with the advantages of wide wavenumber coverage and high dynamic range. The THz-TDS instrument is now being applied to the far-infrared spectrum measurements on dielectric functional materials, polymorphous organic compounds, biomolecules, and crystalline morphology of pharmaceutical reagents. The THz-TDS investigation is also being applied for the non-destructive quality evaluation of industrial products.

An overview of the newly developed instrument performances and some of the latest results of novel applications focusing on the THz-TDS non-destructive quality evaluation of functional dielectric materials and also of pharmaceutical products will be described.

2. Spectrometric Performances

On the Far-infrared spectroscopy, the THz-TDS technique has recently been in progress with advantages over other techniques (Raman and FTIR) and with a wide range of applications for far-infrared analyses of intermolecular vibrational modes and crystalline phonon modes [3].

A compact THz-TDS instrument (Fig.2) has been developed with the advantages of a wide wavenumber coverage applied for versatile measurements (transmission, reflection, liquid, gas, ATR, mapping,

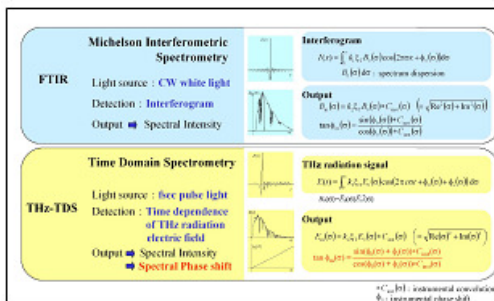


Fig.1 Output of the THz-TDS spectrometry compared with that of the FTIR spectrometry In this FTIR spectroscopy, the light source is continuous white light and its power intensity is detected as an interferograms, which does not exhibit an intrinsic phase shift. The light source in the THz-TDS spectrometry is a coherent femtosecond

pulsed source in which the detected time-domain signals have intrinsic phase shifts. The THz-TDS spectrometry makes output not only the spectral intensity of THz radiation but also the intrinsic phase shift $\Delta\phi(\omega)$ of THz light transmission. The ability to obtain the output with intrinsic phase shift is the most important advantage for spectroscopic applications.



Fig.2 Photographic view of the air evacuated bench unit (upper) of the newly developed versatile THz-TDS instrument. Aispec model: pulse IRS 2000/2000 (lower).

temperature dependence and etc.) [2]. For the THz-TDS, the sample preparation techniques used here are the same as those used in FTIR

and Raman spectroscopy. An absorption spectrum on air was measured and compared with the water vapour absorption spectra offered from the Jet Propulsion Laboratory (JPL, USA). As shown in Fig.3., the wavenumbers of the absorption lines agree with the absorption data of JPL. As shown in Fig.3 (b), the spectral resolution is less than $0.02[\text{cm}^{-1}]$ which was confirmed by the water vapour absorption line at $57.22[\text{cm}^{-1}]$ measured at less than $60[\text{Pa}]$. The spectral coverage depends on the femtosecond laser. The THz radiation source spectra measured by the laser (Femtolasers Productions GmbH: Integral Pro) which pulse duration is less than $10 [\text{fsec}]$ is shown in Fig.4 (a) and (b). The THz radiation is observed from about $1[\text{cm}^{-1}]$ to $230[\text{cm}^{-1}]$. Almost all compact type femtosecond lasers are possible to set on this instrument. The spectral coverage depends on the femtosecond laser as shown in Fig.4.

Between 230 and $340[\text{cm}^{-1}]$, the THz radiation is absorbed by the GaAs phonon absorption bands of the photoconductive antennas. It is expected that the high frequency limit of THz radiation will become higher than $440[\text{cm}^{-1}]$ by avoiding the absorption of the GaAs substrates. The recent progress in the research applications enables the potential application for the far-infrared measurements on new functional materials. A new instrument of THz-TDS has been developed with the advanced optical configuration which consists of a composite THz-TDS optics and a high throughput Michelson (Matin-Puplett configuration) interferometer. A photographic view of the optical configuration is shown in Fig.5. The instrument is for use in the qualitative analyses of optoelectric constants of materials in which the spectrum wavenumber coverage is expanded to the NIR region (Fig.5 (b)).

3. Non-destructive Quality Evaluation

The versatile compact THz-TDS instrument developed with the advantages of a wide wavenumber coverage and a high dynamic range has been in progress applied for far infrared measurements on industrial functional materials, polymorphous organic compounds, biomolecules, and crystalline morphology of pharmaceutical products, of which THz-TDS investigation is applied to the pharmaceutical industry.

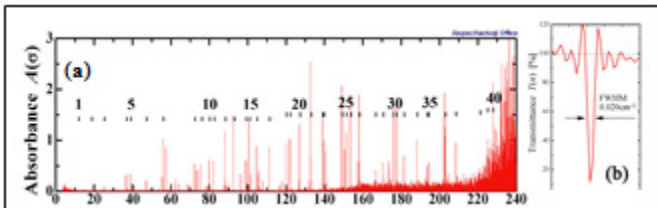


Fig.3. Wavenumber Calibration on the THz-TDS spectrum measurements. The water vapour in the air of 60 Pa absorption lines (a) and the spectral resolution (b).

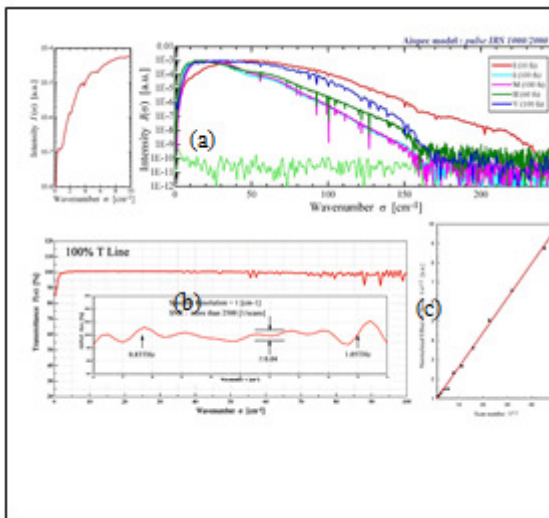


Fig.4 Spectrometric fundamental properties performed on the newly developed versatile the THz-TDS instrument (Aispec model :pulse IRS- 2000). The spectrum wavenumber coverage and spectrometric dynamic range of THz source radiation (a), the noise level and stability on 100% line (b) and the N 1/2 gain on random noise (c).

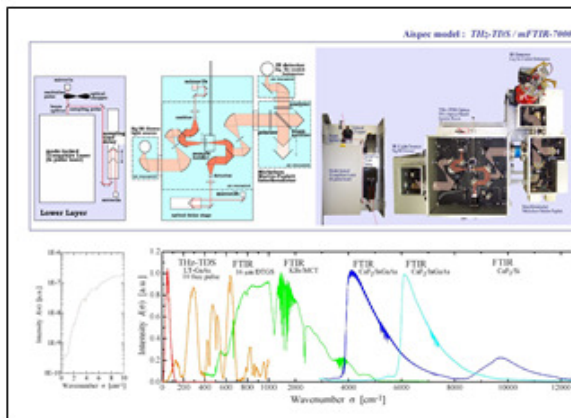


Fig.5 Advanced optics of the composite THz-TDS combined with a high throughput Michelson (Martin- Puplett configuration) interferometer (FTIR) (upper) and expansion of the spectrum wavenumber coverage to the NIR region (lower).

3.1 Dielectric Property Inspections of Ferroelectrics

In ferroelectric materials, the dielectric properties in the terahertz (THz) region are of great importance due to the physical and chemical properties of ferroelectricity dominantly originated in this region. The soft optic modes responsible for ferroelectricity appear in the far-infrared region below 100[cm⁻¹]. The measurements of complex dielectric constants at the THz frequencies are very useful. Since ferroelectric soft modes are infrared active and they propagate as polaritons, the polariton dispersion in the far-infrared region gives very important information for both fundamental and technical problems in ferroelectrics. The ferroelectric materials of current interest are mostly perovskite families with oxygen octahedral. The THz-TDS measurements are in focus on the typical perovskite families of current industrial interest, which includes bismuth titanate Bi₄Ti₃O₁₂ (BIT), lithium heptagermanate, Li₂Ge₇O₁₅ (LGO) and lithium niobate LiNbO₃ (LN). Their ferroelectricity originates dominantly from the instability of polar soft modes in a ferroelectric transition. So the THz dynamics is very important for the characterization of ferroelectric properties. As one of the great technological achievements in the 1990s, oxide ferroelectric thin films have attracted a great deal of attention for use in non-volatile memories. In which bismuth titanate Bi₄Ti₃O₁₂ (BIT) is one of the most important key materials for FeRAM. However, there are still some problems to overcome in order to measure very high-frequency dielectric properties accurately for thin films or for very thin plates. Consequently, an experimental method for determining THz dielectric properties is desired for fundamental and practical research.

As shown in Fig.6(a), the spectral transmission intensity $T(\sigma)$ and intrinsic phase shift $\Delta\phi(\sigma)$ of ac-plate of BIT were accurately measured frequency range from 100[cm⁻¹] down to 3[cm⁻¹]. Fig.6(b) shows the frequency dependences of dielectric constant $\epsilon'(\sigma)$ and loss $\epsilon''(\sigma)$ derived directly from the measured values of transmission intensity and intrinsic phase shift. For light polarization parallel to the a-axis ($E//a$) and b-axis ($E//b$), low frequency polariton branches of $A'(x, z)$ and $A''(y)$ symmetries were clearly observed down to 3[cm⁻¹], respectively. In the polariton dispersion, the wavevector $\kappa(\sigma)$ is reduced for the

observed values of the intrinsic phase shift $\Delta\phi(\sigma)$ through the equation, $\Delta\phi(\sigma) = dk(\sigma)$ where d is the sample crystal thickness.

The observed dispersion curves shown in Fig.6(c) by open circles were well reproduced by the phonon-polariton dispersion curves calculated through the Kurosawa formula [4]. The solid curves in Fig.6(c) denote the curve calculated by the Kurosawa formula using mode frequencies of $A'(x, z)$ symmetry, in which the values of fitting parameters are shown in the table in Fig.6(c). The good agreement concludes that the nonlinear relation reflects the dispersion relation of phonon-polariton. For the polariton coupled to a polar mode with $A'(x, z)$ symmetry at 28 [cm^{-1}], the nonlinear ω - κ relation of the lowest branch was clearly observed in the frequency range 3-26 [cm^{-1}]. The lowest limit of the static dielectric constants along the a- and b-axis: $\epsilon'a(0)=75.99$ and $\epsilon'b(0)=146.41$ were respectively estimated from the lowest polariton branches, whose values showed good agreement with those measured with LCR meter [3].

3.2 Pharmaceutical Quality Inspections of Medicament Products

The THz-TDS provides information on low-frequency intermolecular vibrational modes, and has a wide range of applications in Pharmaceutical industry including formulation, high throughput screening, and inspection in process. The different forms (polymorphs) have the same chemical formula but different crystalline structures that can lead to different physical and chemical properties of the material. The different forms may have different rates of dissolution or bioavailability, and may even effect the stability of the products. The formation of different

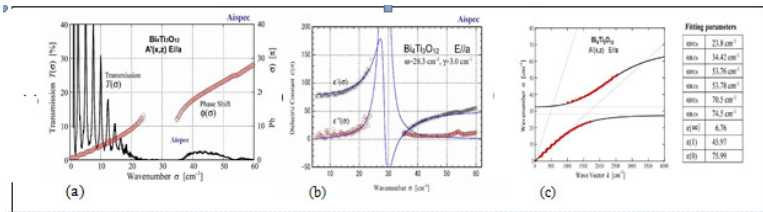


Fig.6 Far-Infrared phonon-polariton dispersion probed for the ferroelectric bismuth titanate Bi₄Ti₃O₁₂ (BIT) crystal by THz-TDS. The transmission $T(\sigma)$ and the intrinsic phase shift $\Delta\phi(\sigma)$ measured on the c-plate of high-quality BIT crystal (a), and the complex dielectric constants $\epsilon^*(\sigma)$ $\{\epsilon'(\sigma), \epsilon''(\sigma)\}$ estimated through $T(\sigma)$ and $\Delta\phi(\sigma)$ of

measured transmission spectra in the light polarization along the a-axis (b) are shown. In (b), the solid lines are the calculated curves using the constants of $\omega_{TO} = 28.3[\text{cm}^{-1}]$ and $\gamma = 3.0[\text{cm}^{-1}]$ obtained through Raman scattering measurements. It can be found that the estimated values of $\epsilon^*(\sigma)$ through THz-TDS transmission measurements are quite in agreement with the calculated values. The dispersion relations of phonon-polariton are also shown (c), in which the closed circles denote the estimated values through $\Delta\phi(\sigma)$ and the solid lines denote the calculated curves of phonon-polariton dispersion relation on the basis of the Kurosawa formula, and the table in (c) shows the values of fitting parameters for observed polariton.

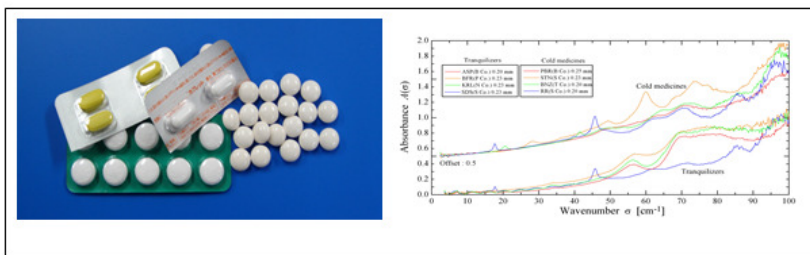


Fig.7 THz-TDS spectral fingerprints obtained by the non-destructive measurements on different kinds of marketed cold medicines and marketed tranquilizers.

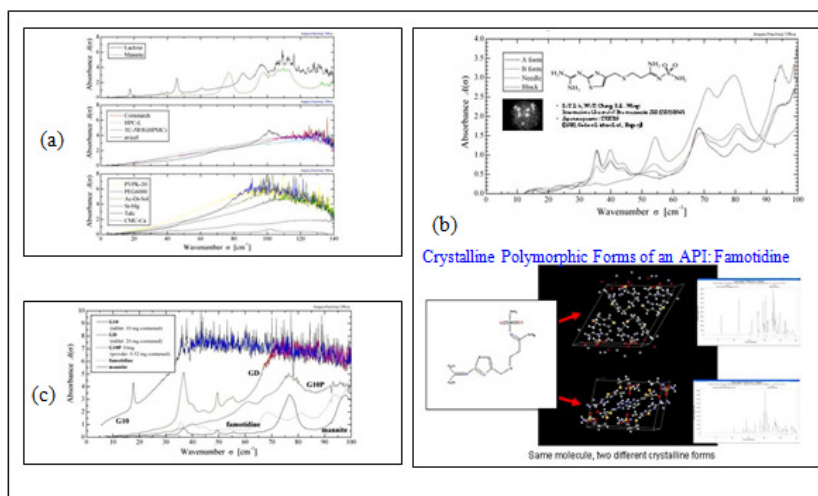


Fig.8 THz-TDS non-destructive evaluation in the pharmaceutical quality control interest. The THz-TDS absorption spectra are measured on the marketed stomach medicine tablets named G10, GD and G10P with the industrial interest in non-destructive quality control (a), the pharmaceutical main active component: API Famotidine, containing the different polymorphs of form A, form B and other forms (b), and also additives (c).

polymorphs can be controlled during crystallization by the solvent used, the rate of cooling, and the degree of super-saturation of the solution. Once in the desired crystalline form the polymorphic state may be changed by incorrect storage or even during tablet preparation. At present, there are no quick and convenient methods for confirming the polymorphic state of products while in storage or during manufacture.

The THz-TDS spectra are sensitive to the difference in the crystalline structure, and thus are applied to search the polymorphs of medicines. In Fig.7, the THz-TDS transmission spectra of some marketed pharmaceutical products are shown with the industrial interest in non-destructive quality evaluation. Fig.8 shows the transmission spectra measured on the tablets over the counter medications named G10, GD and G10P identified with the pharmaceutical main active component, namely API famotidine containing the polymorphs of form A, form B and other forms, and additionally the main additives of lactose and mannite.

A precursory instrument for the quality control on pharmaceutical products [4] has been newly

developed with the advantages of high dynamic range, sensitivity, high throughput screening, and high speed inspection in pharmaceutical product line.

4. Summary

The new instruments of THz-TDS have been developed with different optical bench configurations. One is a composite THz-TDS instrument combined with a high throughput Michelson (Martin-Puplett) interferometer applied for research measurements with wide wavenumber coverage from far-infrared to near-infrared. By the THz-TDS instrument, both measurements with THz-TDS and FTIR are possible without moving a sample specimen. Another is a compact THz-TDS instrument with the adaptation of routine quantitative measurements for characterizing industrial products. This compact instrument enables to measure with various sampling optics units inserted in the sample room.

The versatile THz-TDS instruments have applied for accurate

quantitative measurements and the spectral intensities $T(\omega)$ and the intrinsic phase shifts $\Delta\phi(\omega)$ of transmission spectra have carefully been measured on the ferroelectric materials and a glassy material. From the spectral intensities $T(\omega)$ and the intrinsic phase shifts $\Delta\phi(\omega)$, the complex dielectric constants of the materials were estimated without the uncertainty caused by the Kramers-Kronig analysis in conventional infrared spectroscopy. The dispersion relations of the ferroelectric materials were also derived from the intrinsic phase shifts. The obtained dispersion relations are well reproduced by the phonon-polariton dispersion relations calculated on the basis of Kurosawa's formula [5]. By the THz-TDS instruments, the lowest branch of phonon-polariton dispersion was determined down to the very low frequency of 3cm^{-1} , which could not be attained by other experimental methods. The anisotropy of polariton was also clearly observed simply by rotating the light polarization of the incident THz radiation. The boson peak of a glassy material was also observed by the THz-TDS instruments. Up to now most of properties on boson peaks were discussed only on imaginary parts χ'' (δ) of dynamic susceptibility. Therefore, the determination of both the real part and imaginary part of dielectric constant is very important. We believe these observations are the first reported observations of the dispersion relation of phonon-polariton. The novel applications to pharmaceutical products were described. The THz-TDS has a wide range of applications for industrial especially pharmaceutical quality control including formulation, high throughput screening, and inspection in the product process. It is also suggested that THz-TDS has the technological advantage of non-contact measurement on far-infrared dielectric properties, which is usefully applied to both fundamental study of various far-infrared elementary excitations and utilitarian development of electronics devices, and further to quality control of industry products in the near future.

References

- [1] K. Sakai ed., "Terahertz Optoelectronics", Springer-Verlag Berlin Heidelberg (2005).
- [2] S. Nishizawa, T. Iwamoto, K. Shirawachi, M. W. Takeda, M. Tani and S.

- Sakai, Special Review issues, Proc.1999 IEEE 7th Inter. Conf. on Terahertz Electronics, 308 (1999), S. Nishizawa, T. Iwamoto, S. Kojima and M.W. Takeda, Abstracts of The 2nd International Conference on Advanced Vibrational Spectroscopy (ICAVS-2) CL11 (2003) 48.
- [3] S. Kojima, N. Tsumura, M. W. Takeda and S. Nishizawa, Phys Rev. B67 (2003) 035102
- S..Nishizawa, N. Tsumura, H. Kitahara, M. W. Takeda and S. Kojima, Special Review issues, Phys. Med. Bio. 47 (2002) 3771
- S. Kojima, W. Takeda and S. Nishizawa, J .Molecular Structure 651-653 (2003) 285
- [4] S. Nishizawa, T. Iwamoto, M. W. Takeda and M. Tani, Special Review issues, Abstracts, The 5th International Conference on Advanced Vibrational Spectroscopy (ICAVS-5) P901 (2009) 31
- [5] T. Kurosawa, J. Phys. Soc. Jpn. 16 (1961) 1298.

Frequency tunable meta-material element based on superconducting quantum interference filter

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We present results on digital simulation of a meta-material element (meta-atom) based on thin film superconducting helical resonator which is weakly coupled with the thin film metal resonator. Superconducting resonator contains frequency tunable Josephson multi-element structure. Being in superconducting state its resonance frequency can be controlled by external magnetic field by means of change in inductances of Josephson junctions in the structure. Characteristics of one-dimensional negative media built from proposed meta-atoms are discussed taking into account experimental data of multi-element structure: superconducting quantum interference filter.

1. Introduction

Recently theoretical and experimental studies of meta-materials based on linear passive reciprocal structure were reported (see [1] and references therein). These studies identified a number of problems to be solved for realization of unique features of meta-materials: heat losses, large sizes of elements in microwave frequency band, very complex approaches of implementation using linear structures, and so on. At these circumstances the superconducting resonant structures (SRS) with extremely low losses look promising, particularly SRS based on superconducting Josephson structures [2]. Unlike ordinary passive resonators [3] thin film Josephson SRS can be tuned under applying external magnetic field which changes the resonance frequency. An artificial environment with negative refractive factor at microwave frequencies requires a combination of electric and magnetic dipoles [4, 5]. Usually, electric dipoles are linear dipoles, while the magnetic ones are rings with narrow cut-slot [3]. Following this approach we made

numerical modeling of the basic cell element of meta-material (meta-atoms) for one-dimensional and two-dimensional planar negative media based on Josephson SRS.

2. Meta-atom: multi-element superconducting structure

Single element of negative media (meta-atom) is shown in Figure 1. Meta-atom has two weakly coupled resonators: “magnetic dipole” – a superconducting spiral double line with a superconducting quantum interference filter (SQIF) [6], and electric dipole – a comb shape resonator made from thin low loss metal (gold) film. SQIF is placed in the central part of the spiral resonator and consists of series-connected superconducting rings (dc SQUIDs) each one with two Josephson junctions (JJs) connected in parallel (see inset in Figure 1). The length of the spiral was chosen corresponding to half-wave resonance within frequency range $f=1-3$ GHz. Experimental SQIFs were fabricated using bicrystal NdGaO₃ substrates which dielectric constant is $\epsilon=25$. Critical currents (and self-inductances L_J) of Josephson junctions and SQUIDs are very sensitive to the external magnetic field [6, 7]. The radiating resonator in this circuit is the electric dipole with high Q-factor required for sufficiently well interaction with the superconducting spiral resonator. Figure 2 shows results of spectral characteristics of simulation carried out for frequency range $f=0-3$ GHz when electric dipole is loaded by $50\ \Omega$ line. Single standing electric dipole shows a minimum around 2 GHz. Inserting a magnetic dipole we observe a change in response: narrow peaks appear at $f=1.2$ GHz (low peak) and $f=2.013$ GHz. Because of differences between the radiation losses for electric and magnetic dipoles their Q-factors differ significantly. In a system of two coupled resonators at $f=2.013$ GHz we see an increase in the transparency for the whole structure. A similar behavior was observed earlier for response from two rings with a cut-slot, coupled with the electric dipole [7]. When magnetic field is applied the resonance frequency of magnetic dipole shifts due to change of inductances of Josephson junctions in the SQIF. An important parameter for this meta-atom is the inductance modulation depth. According to our calculations the resonance frequency of magnetic dipole varies with the inductance

of Josephson junctions in the SQIF with a rate 2 1015 Hz/H. Josephson junction inductance is determined by critical current I_C and phases difference φ :

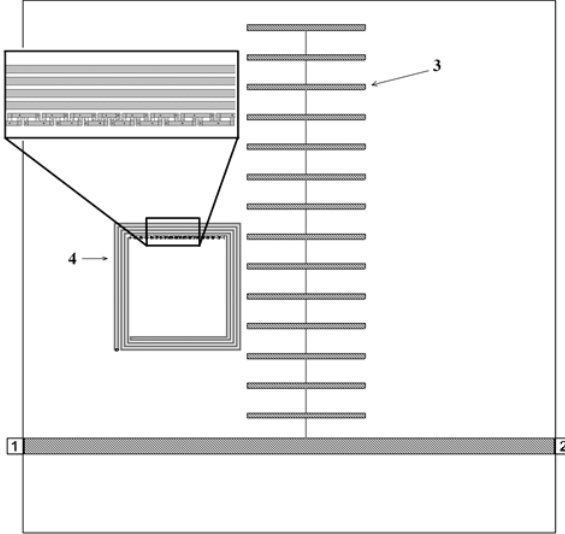


Fig.1. The thin film topology of the meta-atom on NdGaO3 substrate. 1 and 2 are input/output of microwave strip-line, 3 – radiating resonator, 4 – spiral superconducting resonator with SQIF. Inset shows SQIF structure.

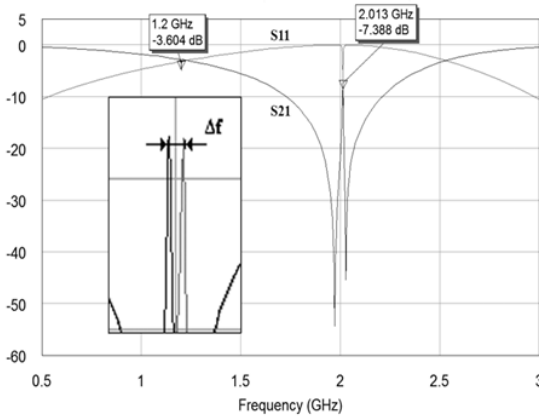


Fig.2. Simulated parameters S21 and S11 for meta-atom, shown in Fig.1. Inset shows 54 MHz change in the resonance frequency when total inductance of Josephson junctions in SQIF changed by 27 pH.

$$L_J = \Phi_0 (2\pi I_C \cos \varphi)^{-1} \quad (1)$$

where $\Phi_0 = h/2e$ - quantum of magnetic flux. Total inductance of the i^{th} SQUID in SQIF includes also the geometric inductance L_{Ki} . Assuming

that the parameters of JJ are identical, for SQUID with two JJ connected in parallel the sum inductance is $L_J/2 + L_{Ki}/2$. Total inductance of SQIF with N serially connected SQUIDs with different L_{Ki} is:

$$L_{\Sigma} = 1/2 \sum_{i=1}^N (L_{Ki} + L_J) \quad (2)$$

From (1) and (2) the change in inductance of the SQIF under the influence of external magnetic flux Φ_e :

$$\frac{dL_{\Sigma}}{d\Phi_e} = \frac{NdL_J}{2d\Phi_e} = \frac{N}{2} \left(-\frac{\hbar \frac{dI_C}{d\Phi_e}}{2eI_C^2 \cdot \cos \varphi} + \frac{\hbar \cdot \sin \varphi \cdot \frac{d\varphi}{d\Phi_e}}{2eI_C \cdot \cos^2 \varphi} \right) \quad (3)$$

Here phase difference in JJ is changed by magnetic field. Total magnetic flux Φ in i th SQUID with the screening current I_s , induced by the external magnetic flux Φ_e is [9]:

$$\Phi = \Phi_e + I_s \cdot L_{ki} \quad (4)$$

For the case of large $L_{Ki} \gg L_J$, which usually takes place in experiment for most SQIF structures, the SQUIDs act as superconducting rings with very minor influence of JJs, conserving magnetic flux unchanged. In this case external flux Φ_e is compensated by current I_s , and the phases of JJ in SQUIDs become independent [9]. This allows us to neglect the second term in (3) and use simple Fraunhofer dependence for critical current of single JJ:

$$I_C = I_C^0 \frac{\sin \pi \frac{\Phi_e}{\Phi_0}}{\pi \frac{\Phi_e}{\Phi_0}} \quad (5)$$

We see that in the SQIF with N serially connected SQUIDs with large inductive loop inductances the variation L_{Ki} is $dL_{\Sigma}/d\Phi_e$ is due to change in the critical current by magnetic field, $dI_C/d\Phi_e$. For thin-film JJ magnetic field and magnetic flux we use well known relation $\Phi_e = \mu_0 H w/2$ [10], where w is width of bicrystal JJ. Taking $\cos \varphi \sim 1$ for the external flux $\Phi_e = \Phi_0/2$ the change of the critical current I_C in (5),

according to (1-3), is $dL\Sigma/d\Phi_e = N/4IC$. Thus, in the case of the SQIF structure with $N = 30$ $w = 10 \mu\text{m}$, and the critical current $IC = 100 \mu\text{A}$ we have $dL\Sigma/d\mu_0H = 7.5 \cdot 10^{-6} \text{ H/T}$.

According to the measurements data for the SQIF -structure with $N = 30$ SQUIDs [6, 7], we can change the slope of the critical current in SQIF structure with magnetic field $dIC/d\mu_0H = 100 \text{ A/T}$. Thus, we obtain $dL\Sigma/d\mu_0H = \Phi_0/(2\pi IC^2) dIC/d\mu_0H = 3 \cdot 10^{-6} \text{ H/T}$, which is slightly less than the estimation made in simulation. Then for typical value of the critical current $IC = 100 \mu\text{A}$ obtain the magnitude of $df/d\mu_0H = 3 \cdot 10^{-6} \cdot fr/L\Sigma \text{ Hz/T}$, where fr is the resonant frequency of SQIF -structure.

Thus, for realistic change $\mu_0H = 100 \mu\text{T}$ of magnetic field we obtain a shift of resonance frequency of SQIF -structure about 600 MHz for $fr = 2 \text{ GHz}$ and $L\Sigma = 1 \text{ nH}$. Here is assumed that the Fraunhofer dependence (5) dominates in SQIF structure based on bicrystal JJs with relatively large superconducting loops. At the same time thin film superconducting loops increase sensitivity to external magnetic field even in the absence of flux transformer. Note also, in SQIF -structure spread of critical currents is less critical, since the main contribution of inductance change comes from the Fraunhofer dependence of the critical current vs. magnetic field of individual bicrystal JJs designed $10 \mu\text{m}$ in widths, which may differ no more than $\pm 0.1 \mu\text{m}$.

3. Concluding remarks

Then, using approach [11] for wave propagation in media with negative permittivity and permeability the existence of backward waves have been shown in narrow frequency band $f = 2.97 - 2.98 \text{ GHz}$ in 1D meta-environment built from series-connected meta-atoms. Calculated bandwidth of backward waves was 1 MHz, and the tuning range of magnetic resonator with SQIF - 54 MHz. Thus, meta-atoms with frequency tunable superconducting SQIF structures are suitable for meta-environment implementation.

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4. References

- [1] R. Marques, F. Martin, M. Sorolla. *Metamaterials with Negative Parameters: Theory, Design and Microwave Applications* (Wiley-Interscience, 2008).
- [2] S.M. Anlage. *J. Opt.* 13 (2011) 024001.
- [3] J. B. Pendry, A. J. Holden, D. J. Robbins, and W. J. Stewart. *IEEE Trans. MTT*, 47, (1999) 2075.
- [4] D.R. Smith, W.J. Padilla, D.C. Vier, S.C. Nemat-Nasser, and S. Schultz. *Phys. Rev. Lett.* 84 (2000) 4184.
- [5] A.V. Shadrin, K.Y. Constantinian, G.A. Ovsyannikov, et al. *Appl. Phys. Lett.* 93 (2008) 262503.
- [6] S.E. Bankov. *Electromagnetic crystals* (Moscow: Phys.Math.Lit. 2010) (in Russian).
- [7] A.V. Shadrin, K.Y. Constantinian, and G.A. Ovsyannikov, *Tech. Phys.Lett.* 33, 192 (2007).
- [8] C. Kurter, Ph. Tassin, L. Zhang, et al. *Phys. Rev. Lett.* 107 (2011) 043901.
- [9] K.K. Likharev. *Dynamics of Josephson junctions and circuits* (New York: Gordon and Breach, 1986).
- [10] P.A. Rozenthal, M.R. Beasley, K. Char, et al. *Appl. Phys. Lett.*, 59 (1991) 3482.
- [11] R.W. Ziolkowsky, E. Heyman. *Phys. Rev. E.* 64 (2001) 056625.

THz radiation in the LiNbO₃ crystal tapered waveguide

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Results of terahertz (THz) radiation generation in the 0.1-2.5 THz band via optical rectification of femtosecond laser pulses in LiNbO₃ crystal tapered are presented. A simulation study for visualization on terahertz wave propagation in LiNbO₃ crystal for developing effective novel THz waveguide system has been performed. The finite-element method is used in the computational technique. It is shown that during propagating of the THz radiation in a wedge shaped crystal the mode structure and phase velocity are changed.

I. Introduction

Due to nonlinear optics achievements, in particular nonlinear frequency conversion of the femtosecond optical laser pulses in THz range and development of THz sources and detectors, the interest in effective waveguide systems, i.e. devices of various forms limiting and guiding electromagnetic waves has considerably grown. The absence of effective, large bandwidth at THz frequency waveguides with low losses and dispersion is a basic barrier hinder for the development and implementation of a device for applications of THz radiation in different fields of science.

The THz waveguide is necessary for their application in fields such as THz non-contact distance sensing [1], time-domain spectroscopy, allowing definition of the refraction indices and material absorption in unique wide bandwidth frequency [2], visualization of latent and malignant tumors in medicine [3], biomedical imaging [4], detection of weapons, explosives and drugs [5, 6], radiation transmission from the radiation source to the antenna or receiver, for developing scanning near-field THz microscope and T-ray tomography [7, 8]. In contrast to

X-ray techniques THz radiation is safer because of non ionized nature, since energy of quant for THz radiation is much less than the energy of X-ray radiation quant.

Recently diverse novel wave guide structures have been examined in the THz region based on traditional metal-cylindrical waveguides [9,10], dielectric fibers [11], plastic ribbon THz waveguides [12], plastic and teflon photonic crystal fibers [13,14] etc. However these waveguides show both strong attenuation and dispersion of the group velocity during propagation of THz waves, preventing THz wave transmission over long distances and limiting the length of the waveguide. To minimize the dispersion in papers [15, 16], parallel-plate metal guides were used. But due-to inadmissibly high weakness and large area of cross section these waveguides cannot be used in different fields including diagnosis in medicine. A waveguide made of stainless steel wire 0.9 mm in diameter and 15 cm in length, as an endoscope for application in medicine was used in [17]. However the effectiveness of radiation coupling with the waveguide was very low. Only 0.1 % linearly polarized THz radiation power generated by photoconductive GaAs antenna is coupled with the waveguide.

Perspective and wide application of THz pulses in spectroscopy and latent image visualization systems impose additional requirements to the waveguides. The waveguides should have not large dispersion in multimode operation, while in case of a quasi-monochromatic wave, with a small spectral band, there is no concept of dispersion.

In this paper the results on generation of THz radiation in band 0.1-2.5 THz via optical rectification of femtosecond laser pulses in LiNbO₃ crystal, tapered at the end, are presented.

For investigation of THz radiation propagation and its visualization in LiNbO₃ crystal waveguide the finite-element method is applied in simulation. The results can be used for creating effective THz broadband active waveguide system or frequency scanning antenna.

2. Laser driven Terahertz LiNbO₃ wedge antenna in free space

Results of THz radiation generation driven by femtosecond laser pulses in a nonlinear dielectric crystal tapered at the end are presented.

As far as we know, this is for the first time that this type of results are discussed. THz pulses were generated via optical rectification of laser pulses in the nonlinear LiNbO_3 crystal. During propagation of an optical laser pulse in a medium with the second order of nonlinear susceptibility a conversion of energy into sum and difference frequency radiation owing to coherent mixing spectral components of the optical pulse occurs. To effectively generate sum or difference frequency radiation, it is necessary to provide the phase-matching condition for all spectral components of the optical pulse participating in a nonlinear process. The possibility of increasing light conversion efficiency into GHz-THz range with a nonlinear crystal partially filling the metallic rectangular waveguide was suggested, theoretically grounded and experimentally realized in [18]. This approach broadens the choice of material for THz generation. A waveguide antenna can be formed from a nonlinear material transparent in both the THz and optical ranges.

A simplified scheme showing the main configuration of the experimental instrumentation used to generate and detection of THz pulse is given in [19]. Optical excitation of THz dielectric antenna was performed by 50 or 200 fs pulses of a Ti:sapphire laser operating with repetition rate of 82 MHz at the near-infrared wavelength 800 nm (Fig.1). The average power on the crystal was not greater than 750 mW. The crystals had cross-section from $0,3 \times 1 \text{ mm}^2$ to $0.1 \times 0.8 \text{ mm}^2$. The optical field strength E and nonlinear polarization P vectors, as well as the optical axis of the crystal were parallel to the narrow wall of the rectangular LiNbO_3 waveguide.

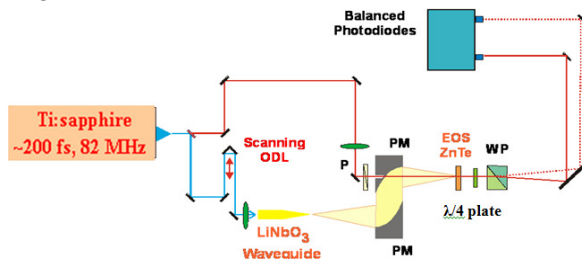


Fig.1. Schematic diagram demonstrating the principles of THz pulse generation and detection.

A small fraction of the pump infrared beam (split from the incident

pump beam) is used for coherent detection. The THz beam is shaped and collimated with the set of parabolic off-axis mirrors (PM). Coherent detection of THz pulses from the LiNbO₃ crystal placed in the free space is performed with a dynamic electro-optical sampling cell consisting of the ZnTe crystal with orientation (110) and thickness 1 mm as well as the quarter-wave plate providing optical bias and the Wollaston prism (WP), which separates the *s* and *p* polarizations of the probe beam as an analyzer. Thus, the whole system represents a version of a coherent pump-probe spectroscopic setup.

The THz output signal from the current amplifier is fed to a lock-in amplifier and then is directed to computer to display the THz-pulse waveform. Two spectra, THz field and phase, are processed via a fast Fourier transform from the originally obtained time-domain dependence.

The temporal waveform (a) of the THz radiation in the far zone from a nonlinear dielectric LiNbO₃ wedge antenna in the free space and its corresponding spectra (b) after fast Fourier transform are shown in Figure 2. The dielectric tapered antenna with a cross-section of 0.3x1 mm² and a length of 8 mm is shown in the Figure 2 (a, b). The cross section of the LiNbO₃ waveguide is constant up to 4 mm length and the taper length is 4 mm as well. The maximum spectral density of the THz radiation in the band up to 2.5 THz is 270 GHz. Because the group velocity of the THz wave in the LiNbO₃ crystal is slower than that of the optical wave ($\epsilon_{\text{THz}}^e = 28$, $\epsilon_{\text{opt}}^e = 4.73$), the THz radiation that emerges is not a stretched derivative laser pulse, but a signal with a large number of cycles, the number being determined by the difference between group velocities.

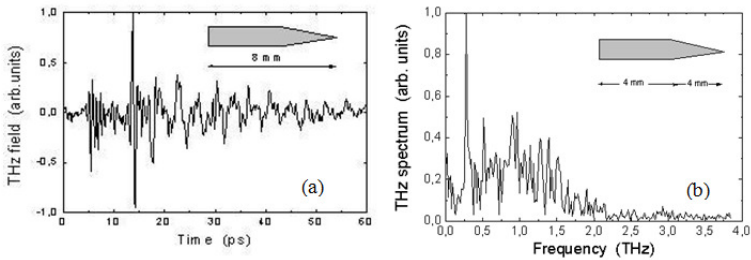


Fig.2. Time dependence of the THz field (a) and its spectrum (b) when the LiNbO₃ crystal is placed in free space.

The crystal with the tapered end provides the broadband matching of the crystal with free space resulting in greater of THz radiation from the crystal compared with its rectangular form. The lengths of the wedge were varied. As it increases, the beam-width decreases, the directional diagram becomes sharper and the relative gain of the LiNbO_3 antenna increases. The tapered form also decreases the lateral lobes of the antenna as well, that is, the condition for propagating the main wave is satisfied. It was experimentally demonstrated that the THz radiation of the LiNbO_3 crystal tapered at the end is 5 to 10 times larger than the THz radiation obtained from a rectangular crystal [20]. At the end of the waveguide the THz mode in the near-field zone is TEM type.

3. Simulations of guided wave phenomena at THz frequencies in iNbO_3 crystal

The ability to model and simulate THz wave propagation aids in the development, visualization, and understanding of novel terahertz devices and phenomena. To simulate and excite THz wave propagation in the rectangular tapered waveguide made of LiNbO_3 crystal placed in free space, the program “COMSOL Multiphysics” is used. A computing apparatus based on finite element method (FEM) was applied for imitating and imaging THz wave propagation. The FEM modeling has been applied to study electromagnetic propagation in both the microwave and optical bands of the spectrum and results are given in [21-23]. The FEM method provides a method for approximating the solution of the differential equations systems in partial derivatives.

The finite element method allows one to define losses and dispersion in THz wave propagation, as well as THz field distribution depending on its spatial location along the waveguide axis and in waveguide cross-section depending on frequency and time. The application of the finite element method is performed in several stages. At first the geometry of the model is specified. The simulation domain is the crystal plate located in free space surrounded by air. After selecting the model geometry and space dimensionality the domain of simulation is divided into small spatial volumes known as subdomains and then physical properties are given. The dielectric permittivity, the magnetic

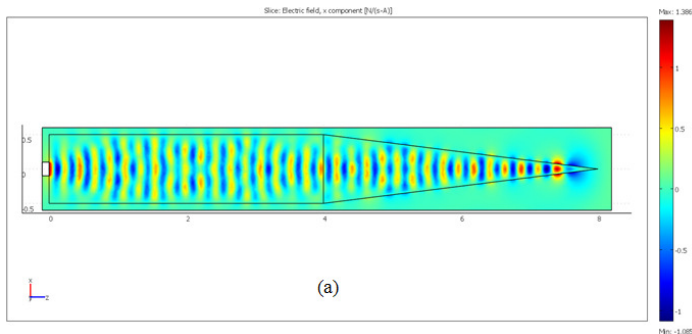
permeability, the refractive index are defined for each subdomain. The simulation domain is subdivided into numerous discrete structural elements called a mesh that are much smaller than the subdomains that make up the simulation domain. The character of the geometric programming mesh depends on the domain of simulation, in the case of a 2D simulation it is triangular and rectangular cells, but in the case of a 3D simulation it is tetrahedral or hexahedral cells [24].

The boundaries of the waveguide separate the spatial volume of simulation, in our case it is 3D space. The simulation domain is divided into tetrahedral cells. The system of elements together, with given physical parameters and boundary conditions, forms a system of linked partial differential equation. Solving the partial differential equation allows visualization of the detection of the electromagnetic field in THz waveguides of different form and materials both in frequency and time domains [25,26]. Time necessary for solving the system of equations by finite element method depends on the simulation size cells number and modeling electromagnetic phenomena. To solve correctly the equation system which describes the distribution of THz wave field, the largest size of the cell should not be greater then one fifth of the wave length [27]. In our case it is 50 μm . Dielectric susceptibility value (ϵ), absorption (α), power for given THz wave frequency are introduced into the program [30].

The real and imaginary parts of the complex dielectric constants of LiNbO_3 crystal were determined for both $\epsilon_z(\omega)$ and $\epsilon_y(\omega)$ in [28] by using THz time domain spectroscopy (THz –TDS). The imaginary part of the complex dielectric constant increases when frequency increases. The real part of the complex dielectric constant $\epsilon_z(\omega)$ is constant in the frequency range from 0.1 THz to 2 THz. Therefore for different frequencies in this range ϵ_z has same real value $\epsilon_z=27,96$ but different values of the imaginary part of the complex dielectric constant $\epsilon'(240 \text{ GHz})=0.091$, $\epsilon'(270 \text{ GHz})=0.102$, $\epsilon'(307 \text{ GHz})=0,116$. These values calculated through the formula given in [29] were inputted into program. The THz wave source was located on the input surface of LiNbO_3 plate with cross section $0.3 \times 1 \text{ mm}^2$.

The propagation of THz waves in LiNbO_3 waveguide with fre-

quencies equal to the most intensive spectral radiation lines in THz pulse spectra Fig.2.(b) at 240 GHz, 270 GHz, 307 GHz [28] generated via optical rectification by femtosecond pulses of a Ti:sapphire laser was studied. To provide the absence of the THz wave reflections from output surface of the LiNbO₃ crystal due to an impedance mismatch with free space, it was given a wedge shape. The distribution of THz electric field E_x and E_y components at 270 GHz frequency ($\lambda = 1.1$ mm) in propagation along the z axis, as well as in 13 planes of the LiNbO₃ waveguide are given in Fig.3. The red color indicates positive values and blue indicates negative values of the THz electric field on Fig.3 (a, b). As the THz wave propagates in the narrowing part of the waveguide the THz field is being focused and passes from the multimode regime to single-mode. The radiation appearing from the tip of the wedge-shaped waveguide demonstrates the possibility to use the LiNbO₃ waveguide as a THz antenna driven by optical laser pulses. As the length of the tip increases, the beam-width decreases. The dependence of the $E_x(z)$ -component of the THz electric field at 270 GHz frequency on the crystal length is presented in the graph in Fig.3. Fig.4 demonstrates stable propagation of the THz wave modulated oscillation with increasing amplitude that is the evidence of transition process. Because of the continuously variable height of the narrow wall of the waveguide is a change of the critical frequency and the speed of THz wave. 2-D images of the calculated spatial distribution of the THz electric field in plane xy at the end of the waveguide ($L=7.9$ mm) and out of the waveguide in the near-field zone ($L=8,05$ mm) are shown in Fig.5.



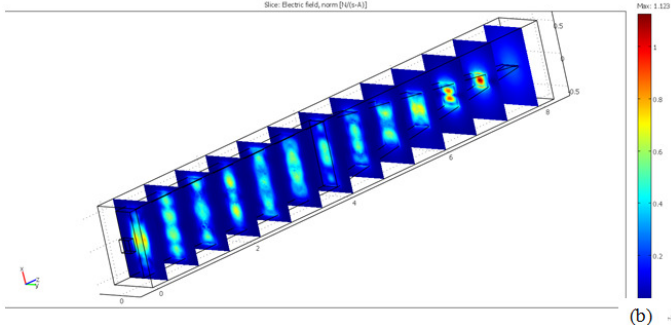


Fig.3. Distribution of THz electric field E_x component with 270 GHz frequency in propagation along the z axis: (a) lateral view in (xz) plane, (b) spatial distribution of THz field in 13 planes of the LiNbO_3 waveguide.

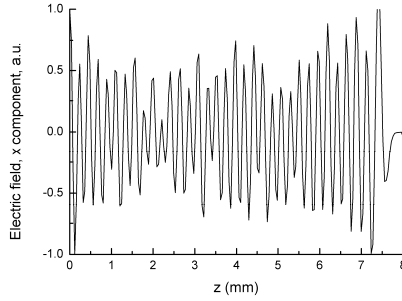


Fig.4. Graph of the $E_x(z)$ -component of the THz electric field at 270 GHz frequency depending on crystal length $L=8\text{mm}$.

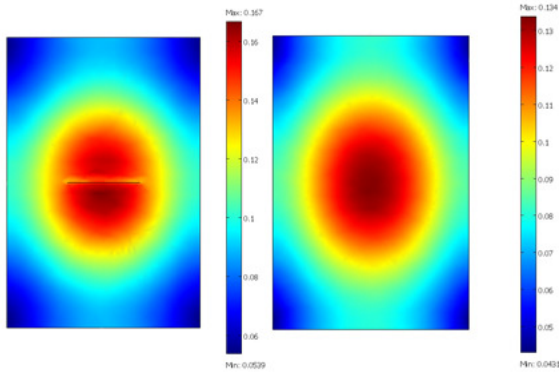


Fig.5. The spatial distribution of the THz electric field in plane (xy plane shown) at the end of the waveguide ($L=7.9\text{ mm}$) and out ($L=8.05\text{ mm}$) at 270 GHz.

4. CONCLUSIONS

We employ the finite-element method to modeling and simulate propagation of THz waves in LiNbO_3 wedge waveguides to understand experimental results and visualize the influence the form of the waveguide on the THz radiation inside waveguide and outside in the near-field zone. As far as we know, no results have yet been reported concerning experimental and simulation with LiNbO_3 wedge-shaped waveguides. We investigated how tapering of the output surface of the crystal waveguide effects on the THz mode type and the spatial extent. The spatial extent of the THz electric field becomes narrower as the length of wedge increases and as its cross-section decreases, since the propagating mode is compressed. The compression of THz field along the wedge-shaped waveguide explains the result stated experimentally that the THz radiation (in the range $0.1 \div 3$ THz) from LiNbO_3 crystal tapered at the end $\sim 5 \div 10$ times exceeds [20] the THz radiation obtained from a rectangular crystal. Alternation of the mode structure during propagation of the THz radiation in the wedge-shaped crystal is accompanied by a change of the phase and group velocities. There is variation of the phase and group velocities in the wedge-shaped LiNbO_3 waveguide we have used [19] for the efficient generation of THz pulse at the condition of phase matching. The simulation results show that at the end of the waveguide, the spatial shape of the THz field is broader at lower 240 GHz and 270 GHz frequencies because the diffraction is stronger for the low frequencies than for the high 307 GHz frequency.

The simulation shows the ability to focus the spatial extent of the THz field provides a promising direction for the use of wedge waveguides in various applications, for instance in THz microscopy and imaging or using them as THz dielectric antennas.

REFERENCES

1. D. Mittleman. (ed.) Sensing with Terahertz Radiation. Springer, Heidelberg (2002).

2. D. Grischkowsky, S. Keiding, M. van Exter, and Ch. Fattinger. Far-infrared time-domain spectroscopy with terahertz beams of dielectrics and semiconductors, *J. Opt. Soc. Amer. B*, **7**, 2006–2015, (1990).
3. N.N. Zinov'ev, A.S. Nikoghosyan, J.M. Chamberlain. *Proc. of SPIE*, **6257**, 62570P1–62570P8, (2006).
4. R.M. Woodward et al., Terahertz pulse dimaging of skin cancer in the time and frequency domain, *J. Biol. Phys.*, **29**, 257–261, (2003).
5. M.R. Leahy-Hoppa et al., Wideband terahertz spectroscopy of explosives, *Chem. Phys. Lett.*, **434**, no. 4–6, 227–230, (2007).
6. K. Kawase, Y. Ogawa, and Y. Watanabe, Non-destructive terahertz imaging of illicit drugs using spectral fingerprints, *Opt. Express*, **11**, no. 20, 2549–2554, (2003).
7. N.N. Zinov'ev, A.V. Andrianov. Terahertz confocal imaging. *Appl. Phys. Letters*, **95**, 011114-3, (2009).
8. S. Wang, X.-C. Zhang. Pulsed terahertz tomography. *J. Phys. D*, **37**, R1–R36, (2004).
9. R. W. McGowan, G. Gallot, D. Grischkowsky. Propagation of ultrawideband short pulses of THz radiation through submillimeter-diameter circular waveguides. *Opt. Lett.* **24**, 1431–1433, (1999).
10. G. Gallot, S.P. Jamison, R.W. McGowan, and D. Grischkowsky. Terahertz waveguides. *J. Opt. Soc. Am. B* **17**, 851–863, (2000).
11. S.P. Jamison, R.W. McGowan, D. Grischkowsky, S.P. Jamison, R.W. McGowan, D. Grischkowsky. Terahertz Single-mode waveguide propagation and reshaping of sub-ps terahertz pulses in sapphire fiber. *Appl. Phys. Lett.* **76**, 1987–1989, (2000).
12. R. Mendis, D. Grischkowsky. Plastic ribbon THz waveguides. *J. Appl. Phys.* **88**, 4449–4451, (2000).
13. H. Han, H. Park, M. Cho, J. Kim. Terahertz pulse propagation in a plastic photonic crystal fiber. *Appl. Phys. Lett.* **80**, 2634–2636, (2002).
14. M. Goto, A. Quema, H. Takahashi, S. Ono, N. Sarukura. Teflon photonic crystal fiber as terahertz waveguide. *Jap. J. Appl. Phys.* **43**, L317–L319, (2004).
15. R. Mendis, D. Grischkowsky. Undistorted guided-wave propagation of subpicosecond terahertz pulses. *Opt. Lett.* **26**, 846–848, (2001).
16. R. Mendis, D. Grischkowsky. THz interconnect with low loss and low group velocity dispersion. *IEEE Microwave Wireless Comp. Lett.* **11**, 444–446, (2001).

17. K. Wang, D.M. Mittleman. Metal wires for terahertz wave guiding. *Nature*, **432**, 376-379, (2004).
18. E.M. Laziev, A.S. Nikoghosyan. "Phaze Synchronizm and Ultrashort SHF Pulses Generation in Waveguides". SPIE, Mode-Locked Lasers and Ultrafast Phenomena, v.**1842**, 113-118, (1991).
19. A.S.Nikoghosyan, P.M. Martirosyan, A.A. Hakhoumian, J.M. Chamberlain, R.A. Dudley, N.N. Zinov'ev. Generation of THz Radiation in Waveguides Partially Loaded with Nonlinear Crystal. *Inter. Journal Electromagnetic Waves and Electronic System*. 11, N4, 47-55, (2006).
20. A.S. Nikoghosyan. Ultrashort pulses generation in the millimeter and centimeter bands in a waveguide partially filled with a nonlinear crystal. *Quantum Electronic*. 15, 969-971, (1988).
21. R. Coccioli et al. *IEEE Antennas Propag. Mag.*, **38**, 34, (1996).
22. B. M. A. Rahman, F. A. Fernandez, J. B. Davies, *Proc. IEEE*, **79**, 1442, (1991).
23. I. Tsukerman, F. Cajko, A. P. Sokolov, *Proc. SPIE, Plasmonics: Metallic Nanostructures and Their Optical Properties III*, **5927**, 128, (2005).
24. J.A. Deibel and D.M. Mittleman Designing and simulating THz wire waveguides using FEM modeling. www.rfdesign.com. 36-43, August (2006).
25. D. Jiao, J.-M. Jin. *IEEE Trans. Antennas Propag.*, **50**, 1624, (2002).
26. J.-F. Lee, R. Lee, A. Cangellaris. *IEEE Trans. Antennas Propag.*, **45**, 430, (1997).
27. J. Jin. *The Finite Element Method in Electromagnetics*. New York: John Wiley & Sonc, Inc., (2002).
28. S. Kojima, N. Tsumura, H. Kitahara, M.W. Takeda, S. Nishizawa. Terahertz Time Domain Spectroscopy of Phonon-Polaritons in Ferroelectric Lithium Niobate Crystals. *Jpn. J. Appl. Phys.* **41**, 7033-7037 (2002).
29. N.N. Zinov'ev, A.S. Nikoghosyan, J.M. Chamberlain. Terahertz Radiation from a Nonlinear Slab Traversed by an Optical Pulse. *Physical Review Letters*, **98**, 044801-1 044801-4, (2007).

Generation of difference frequency radiation in the field of few-cycle laser pulse propagating in periodically-poled Lithium Niobate

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In this paper we study the generation of difference frequency radiation (DFR) in wavelength range $(3 - 4) \mu\text{m}$ that occurs during propagation of an ordinary linearly polarized laser pulse of a few optical cycles in periodically-poled LiNbO_3 (PPLN) in the direction normal to the optical axis of a nonlinear uniaxial crystal of 3m symmetry group. We describe propagation of the laser pulse by solving Maxwell's equations with use of the finite-difference time domain (FDTD) method. As an example we calculate the spectrum profile of DFR generated during the propagation of laser pump pulse with duration 10 fs at 810 nm wavelength. We apply our model to study the propagation of a femtosecond laser pulse in a PPLN consist from 20 domain. The thickness of considered domain equal $6.3\mu\text{m}$, which provides the phase quasi synchronism for DFR generation.

1. Introduction

In recent years considerable successes have been achieved in generation of pulsed DFR by optical methods, among which the most widespread became the techniques based on exploiting electro-optical materials [1-3]. The considerable activity in using infrared region is motivated by the utility of such sources in fundamental and applied studies in optoelectronics and optical fibre communications $(1-2) \mu\text{m}$, material sciences and time-resolved chemistry, as well as in biology $(3-10) \mu\text{m}$. Consequently, the problem of broadband pulsed DFR generation in the field of few-cycle laser pulse, propagating in PPLN, is of considerable practical interest.

In this paper, we study the generation of DFR in wavelength range $(3-4) \mu\text{m}$ that occurs during the propagation of an ordinary linearly polarized laser pulse of a few optical cycles with duration 10 fs at

central wavelength $0.81 \mu\text{m}$ in the direction normal to the optical axis of a PPLN. The considered PPLN consist from 10 periods where the length of period is $12.6 \mu\text{m}$. We solve Maxwell's equations to describe the propagation of the laser pulse using the FDTD method. The time profile and spectrum distribution of extraordinary linearly polarized pulse, arising due to nonlinear interaction of laser pulse with PPLN, are obtained.

2. Mathematical model for the description of nonlinear interaction of linearly-polarized few-cycle laser pulses with PPLN

Let us consider a linearly polarized laser pulse of a few optical cycles with field components E_x and H_z with a plane wave front propagating along the y -axis, normal to the optical axis of a uniaxial crystal of $3m$ symmetry group. Due to nonlinear polarization term during propagation will generated the E_z and H_x also. Maxwell's equations in this case are written as

$$\frac{\partial D_z}{\partial t} = -\frac{\partial H_x}{\partial y}, \quad (1),$$

$$\frac{\partial H_x}{\partial t} = -\frac{1}{\mu_0} \cdot \frac{\partial E_z}{\partial y}, E_z = \frac{D_z - P_{zL} - P_{zNL}}{\epsilon_0}$$

$$\frac{\partial D_x}{\partial t} = \frac{\partial H_z}{\partial y}, \quad (2),$$

$$\frac{\partial H_z}{\partial t} = \frac{1}{\mu_0} \cdot \frac{\partial E_x}{\partial y}, E_x = \frac{D_x - P_{xL} - P_{xNL}}{\epsilon_0}$$

where D_z , D_x is the electric induction components, ϵ_0 is the free-space permeability, μ_0 is the free-space permittivity, and P_{zL} , P_{xL} , P_{xNL} and P_{zNL} are the induced linear and nonlinear electric polarizations, respectively. The components D_z , D_x of the electric induction are determined from constitutive equations, where linear dispersion as well as nonlinear polarization of the medium is successively taken into account,

$$D_z = \epsilon_0 \cdot E_z + P_{zL} + P_{zNL} \quad (3) \quad D_x = \epsilon_0 \cdot E_x + P_{xL} + P_{xNL} \quad (4)$$

The induced linear electric polarization can be written as

$$P_{xL}(t) = \epsilon_0 \cdot \int_{-\infty}^t \chi_x^{(1)}(t-\tau) \cdot E_x(\tau) \cdot d\tau, \quad P_{zL}(t) = \epsilon_0 \cdot \int_{-\infty}^t \chi_z^{(1)}(t-\tau) \cdot E_z(\tau) \cdot d\tau \quad (5):$$

In the nonlinear part of the polarization of a uniaxial optical crystal of 3m symmetry group, such as LiNbO₃, which is responsible for second order-nonlinearities, we shall confine ourselves to the quasistatic approximation

$$P_{xNL}(t) = 2 \cdot \epsilon_0 \cdot \tilde{d}_{31}(y) \cdot E_x(t) \cdot E_z(t), \\ P_{zNL}(t) = \epsilon_0 \cdot \tilde{d}_{33}(y) \cdot E_z^2(t) + \epsilon_0 \cdot \tilde{d}_{31}(y) \cdot E_x^2(t) \quad (6),$$

where the nonlinear coefficients of a PPLN crystal with periodically-inverted sign of second-order susceptibilities with a duty cycle of 50%, which can be presented in the form of a Fourier series

$$\tilde{d}_{33(31)}(y) = d_{33(31)} \sum_{k=0}^K \frac{\sin(2\pi y[2k+1]/\Lambda) \sin[\pi(k+1)/K]}{(2k+1) \pi(k+1)/K} \quad (7)$$

$d_{31} = 5.44 \cdot 10^{-12}$ m/V, $d_{33} = 2.76 \cdot 10^{-12}$ m/V are the coefficients of nonlinear susceptibility of the LiNbO₃ crystal, Λ period of LiNbO₃ second-order susceptibility sign inversion. The corresponding frequency-domain response functions $\chi_x^{(1)}(\omega), \chi_z^{(1)}(\omega)$ are related to the linear index of refraction of the ordinary $n_o(\omega, T)$ and extraordinary $n_e(\omega, T)$ waves in the medium by the equations

$$\chi_x^{(1)}(\omega) = n_o^2(\omega, T) - 1 \quad (8)$$

$$\chi_z^{(1)}(\omega) = n_e^2(\omega, T) - 1. \quad (9)$$

For optical frequencies that are far from the resonant frequencies of the medium ((0.35 – 5.5) μm), the indexes of refraction of the LiNbO₃ uniaxial nonlinear crystal can be approximated by the Sellmeier

equations [4]. These equations are based on the classical Lorentz model of an atom and have the following form:

$$\epsilon_{or}(\omega, T) = n_{or}^2(\omega, T) = 4.913 + \frac{\frac{(0.1173 + 1.65 \cdot 10^{-8} \cdot T^2) \cdot \omega^2}{(2\pi \cdot c)^2}}{1 - \frac{(0.212 + 2.7 \cdot 10^{-8} \cdot T^2) \cdot \omega^2}{(2\pi \cdot c)^2}} - 2.78 \cdot 10^{-2} \cdot \frac{(2\pi \cdot c)^2}{\omega^2} \quad (10),$$

$$\epsilon_e(\omega, T) = n_e^2(\omega, T) = 4.5567 + 2.605 \cdot 10^{-7} \cdot T^2 + \frac{\frac{(0.097 + 2.7 \cdot 10^{-8} \cdot T^2) \cdot \omega^2}{(2\pi \cdot c)^2}}{1 - \frac{(0.201 + 5.4 \cdot 10^{-8} \cdot T^2) \cdot \omega^2}{(2\pi \cdot c)^2}} - 2.24 \cdot 10^{-2} \cdot \frac{(2\pi \cdot c)^2}{\omega^2} \quad (11),$$

where T is the temperature and c is the speed of light in vacuum in $\mu\text{m}/\text{sec}$. The index of refraction was calculated at $T = 293$ K. According to (10) and (11) the linear frequency-domain responses functions $\chi_x^{(1)}(\omega), \chi_z^{(1)}(\omega)$ can be represented in the following form:

$$\chi_x^{(1)}(\omega) = \left(a_o - \frac{b_o}{c_o^2} \right) + \frac{(b_o/c_o^2)}{1 - c_o^2 \cdot \omega^2} - \frac{q_o}{\omega^2} \quad (12),$$

$$\chi_z^{(1)}(\omega) = \left(a_e - \frac{b_e}{c_e^2} \right) + \frac{(b_e/c_e^2)}{1 - c_e^2 \cdot \omega^2} - \frac{q_e}{\omega^2} \quad (13),$$

where

$$\begin{aligned} a_o &= 3.913, \quad b_o = \left(\frac{0.1173}{(2\pi \cdot c)^2} + \frac{1.65 \cdot 10^{-8} \cdot T^2}{(2\pi \cdot c)^2} \right) \\ c_o &= (0.212 + 2.7 \cdot 10^{-8} \cdot T^2) / (2\pi \cdot c), \quad q_o = 2.78 \cdot 10^{-2} \cdot (2\pi \cdot c)^2 \\ a_e &= (3.5567 + 2.605 \cdot 10^{-7} \cdot T^2), \quad b_e = \left(\frac{0.097}{(2\pi \cdot c)^2} + \frac{2.7 \cdot 10^{-8} \cdot T^2}{(2\pi \cdot c)^2} \right) \\ c_e &= (0.201 + 5.4 \cdot 10^{-8} \cdot T^2) / (2\pi \cdot c), \quad q_e = 2.24 \cdot 10^{-2} \cdot (2\pi \cdot c)^2 \end{aligned} \quad (14).$$

As shown in [5] the nonlinear susceptibilities describing the difference-frequency Ω generation can be presented as

$$\chi_{x,z}^{(2)}(\Omega; \omega_0 + \Delta\Omega/2, \omega_0 - \Delta\Omega/2) \sim \chi_{x,z}^{(1)}(\Omega) \chi_{x,z}^{(1)}(\omega_0 + \Delta\Omega/2) \chi_{x,z}^{(1)}(\omega_0 - \Delta\Omega/2) \quad (15)$$

where $\omega_0 \pm \Omega/2$ the spectral component shifted on $\Omega/2$ oscillation frequency contained in the pump pulse

spectrum bandwidth. For difference-frequency radiation bandwidth $\Delta\nu = 1/\tau_0 \sqrt{\ln(2)} = \Delta\Omega/2\pi = 120\text{THz}$ ($\Delta\lambda = 270\text{nm}$) in accordance with (8) and (9) it can be shown that $\left(\partial\chi_{x,z}^{(2)}/\partial\Omega\right)_{\omega_0} \Delta\Omega/\left(\chi_{x,z}^{(2)}\right)_{\omega_0}$ around of $3\text{ }\mu\text{m}$ wavelength of difference frequency radiation and at $\lambda_0 = 0.81\text{ }\mu\text{m}$ (Gaussian pulse duration $\tau_0=10\text{ fs}$) is no more than 0.1. Consequently in infrared range of wavelengths $(3\div 4)\text{ }\mu\text{m}$ may be used quasistatic approximation, which corresponds to instantaneous nonlinear response of the medium. At the same time susceptibilities describing the summary-frequency (SF) ω_{SF} generation can be presented as

$$\chi_{x,z}^{(2)}(\omega_{\text{SF}}; \omega_s, \omega_{\text{SF}} - \omega_s) \sim \chi_{x,z}^{(1)}(\omega_{\text{SF}}) \chi_{x,z}^{(1)}(\omega_s) \chi_{x,z}^{(1)}(\omega_{\text{SF}} - \omega_s) \quad (16)$$

where $\omega_{\text{SF}} = \omega_s + \omega_l$, $\omega_s = 2\pi c/\lambda_s$, $\omega_l = 2\pi c/\lambda_l$, λ_s - short wavelength component (signal wavelength) λ_l long wavelength component (idler wavelength) contained in the spectrum bandwidth of the few-cycle duration pulse. For SF radiation, generated due to mixing of signal and idler wavelength components inside of pulse bandwidth $\Delta\lambda=270\text{ nm}$, in accordance with (8) and (9) it can be shown that $\left(\partial\chi_{x,z}^{(2)}/\partial\omega_{\text{SF}}\right)_{\omega_s} \Delta\Omega/\left(\chi_{x,z}^{(2)}\right)_{\omega_s}$ around of $0.38\text{ }\mu\text{m}$ wavelength of summary frequency radiation and at $\lambda_0 = 0.81\text{ }\mu\text{m}$ (Gaussian pulse duration $\tau_0=10\text{ fs}$) is no more than 0.075. Consequently in the range of wavelengths $(0.38\div 0.44)\text{ }\mu\text{m}$ may be used quasistatic approximation, which corresponds to instantaneous nonlinear response of the medium. The three terms for the linear response function in (11), (12) lead to the following decompositions for the corresponding time-dependent polarizations:

$$\begin{aligned} P_{xL}(t) &= \varepsilon_0 \cdot \varepsilon_{co} \cdot E_x(t) + F_x(t) + G_x(t) \\ P_{zL}(t) &= \varepsilon_0 \cdot \varepsilon_{cc} \cdot E_z(t) + F_z(t) + G_z(t) \end{aligned} \quad (16),$$

where $\varepsilon_{ce} = a_e - b_e/c_e^2$, $\varepsilon_{co} = a_o - b_o/c_o^2$ and $F_{x,z}(t)$ and $G_{x,z}(t)$ are the solutions of the following ordinary differential equations:

$$c_o^2 \cdot \frac{\partial^2 F_x(t)}{\partial t^2} + F_x(t) = \epsilon_o \cdot \frac{b_o}{c_o^2} \cdot E_x(t, y), \quad c_e^2 \cdot \frac{\partial^2 F_z(t)}{\partial t^2} + F_z(t) = \epsilon_o \cdot \frac{b_e}{c_e^2} \cdot E_z(t, y), \quad (17),$$

$$\frac{\partial^2 G_x(t)}{\partial t^2} = \epsilon_o \cdot q_o \cdot E_x(t, y) \quad \frac{\partial^2 G_z(t)}{\partial t^2} = \epsilon_o \cdot q_e \cdot E_z(t, y)$$

The system of equations (17) describes the linear dispersion properties of the medium in the transparency region according to the Lorentz classical model. Taking into account (1)-(4), and (6) the electric-flux densities D_z , D_x may be represented as

$$D_z = E_z + \epsilon_{ce} \cdot E_z(t) + F_z(t) + G_z(t) + \tilde{d}(y)_{33} \cdot E_z^2(t) + \tilde{d}(y)_{31} \cdot E_x^2(t) \quad (18),$$

$$D_x = E_x + \epsilon_{co} \cdot E_x(t) + F_x(t) + G_x(t) + 2 \cdot \tilde{d}(y)_{31} \cdot E_x(t) \cdot E_z(t) \quad (19).$$

The classical model of interaction between a laser pulse of a few optical cycles and a nonlinear dispersive medium described above was used for the description of the parametric near infrared radiation generation [6]. In this paper we use this model for the analysis of DFR and SFR generation by femtosecond laser pulse of a few optical cycles propagating in nonlinear dispersive uniaxial crystal of $3m$ symmetry group. The above decrypted second-order nonlinear wavelength conversion technique related to the difference frequency generation (DFG), which is called down-conversion. Optical waves of two different frequencies are mixed to generate the third optical wave with a frequency which is the difference of the two input frequencies. Due to second-order nonlinear wavelength conversion efficient generation of the sum-frequency occurs also which is called parametric up-conversion or sum frequency generation (SFG). SFG, DFG require phase matching to be efficient. For wavelength conversion phase matching is essential for increasing the efficiency of wavelength conversion. This means that a proper phase relationship between the interacting waves (for maximum wavelength conversion) is maintained along the propagation direction, so that the amplitude contributions from different locations to the resultant wave are all in phase. This leads to the condition that phase mismatch has to be zero. A broadband few cycle laser pulse, which consists of several frequencies such as ω_1 and ω_2 , can induce nonlinear

polarization with a difference frequency, $\omega_{\text{DF}} = \omega_1 - \omega_2$, assuming $\omega_1 > \omega_2$ and summary frequency $\omega_{\text{SF}} = \omega_1 + \omega_2$ in a periodically poled nonlinear crystal. For the *ooe* types of interaction the energy and the momentum conservation laws for generating difference frequency can be described as

$$E_{\omega_1} = E_{\omega_2} + E_{\omega_{\text{DF}}}, \quad \bar{K}_{\omega_1} = \bar{K}_{\omega_2} + \bar{K}_{\omega_{\text{DF}}} + \bar{K}_{\text{AD}} \quad (20),$$

and for generating summary frequency can be described as

$$E_{\omega_{\text{SF}}} = E_{\omega_1} + E_{\omega_2}, \quad \bar{K}_{\omega_{\text{SF}}} = \bar{K}_{\omega_1} + \bar{K}_{\omega_2} + \bar{K}_{\text{AS}} \quad (21)$$

where E_{ω} is the photon energy, K is the wave vector at each frequency, and $\bar{K}_{\text{AD}} = 2\pi/\Lambda_{\text{DF}}$, and $\bar{K}_{\text{AS}} = 2\pi/\Lambda_{\text{SF}}$, are the gratings wave vectors. With taking into account the above mentioned waves interaction geometry, which described by $\epsilon_0 \cdot \tilde{d}_{31}(y) \cdot E_x^2(t)$ term in (6), equation (20), (21), can be written more detail as

$$\frac{1}{\lambda_1} = \frac{1}{\lambda_2} + \frac{1}{\lambda_{\text{DF}}}, \quad \frac{n_o(\lambda_1)}{\lambda_1} = \frac{n_o(\lambda_2)}{\lambda_2} + \frac{n_e(\lambda_{\text{DF}})}{\lambda_{\text{DF}}} + \frac{1}{\Lambda_{\text{DF}}} \quad (22)$$

for DFG, and

$$\frac{1}{\lambda_{\text{SF}}} = \frac{1}{\lambda_1} + \frac{1}{\lambda_2}, \quad \frac{n_e(\lambda_{\text{SF}})}{\lambda_{\text{SF}}} = \frac{n_o(\lambda_1)}{\lambda_1} + \frac{n_o(\lambda_2)}{\lambda_2} + \frac{1}{\Lambda_{\text{SF}}} \quad (23)$$

for SFG.

According to (22) via mixing of short wavelength component (pump wavelength) $\lambda_1 = \lambda_p$ with long wavelength component (signal wavelength) $\lambda_2 = \lambda_s$, contained in the spectrum bandwidth of the few-cycle duration pulse, the difference-frequency radiation at λ_{DF} wavelength produced. Phase-matched interaction of considered waves takes place at the λ_p , λ_s and λ_{DF} wavelengths for which fulfilled the conservation laws conditions (20) and (22). According to (23) via mixing of short wavelength component (signal wavelength) $\lambda_1 = \lambda_s$ with long wavelength component (idler wavelength) $\lambda_2 = \lambda_i$, contained in the spectrum bandwidth of the few-cycle duration pulse, the summary-

frequency radiation at λ_{SF} wavelength produced. Phase-matched interaction of considered waves takes place at the λ_{S} , λ_{I} and λ_{SF} wavelengths for which fulfilled the conservation laws conditions (21) and (23). It's known that when the ratio $\Lambda_{\text{DF}}/\Lambda_{\text{SF}}$ is an integer odd number or is the ratio of odd numbers than take place the simultaneous the phase quasisynchronism for summary and difference frequencies radiations [7]. In particular, at difference-frequency radiation wavelength $\lambda_{\text{DF}} = 3 \mu\text{m}$, short wavelength component (pump wavelength) $\lambda_{\text{p}} = 0.776 \mu\text{m}$ quasisynchronism for DR will be at $\Lambda_{\text{DF}} = 12.6 \mu\text{m}$. According to the numerically obtained dependence of the ratio $\Lambda_{\text{DF}}/\Lambda_{\text{SF}}$ vs. λ_{SF} , and according to calculation results at λ_{SF} about $0.44 \mu\text{m}$ the $\Lambda_{\text{DF}}/\Lambda_{\text{SF}} \approx 1$ and at $\Lambda_{\text{DF}}/\Lambda_{\text{SF}} = 5/3; 7/5; 9/5; 9/7; 11/7; 13/7; 11/9; 13/9; 15/9$ and $17/9$ will be take place the simultaneous the phase quasisynchronism for SFR at $\lambda_{\text{SF}} = 0.3981 \mu\text{m}; 0.4116 \mu\text{m}; 0.3895 \mu\text{m}; 0.4218 \mu\text{m}; 0.4037 \mu\text{m}; 0.3860 \mu\text{m}; 0.4265 \mu\text{m}; 0.4116 \mu\text{m}; 0.3981 \mu\text{m}$ and $0.3841 \mu\text{m}$ respectively and for DFR at $\lambda_{\text{DF}} = 3 \mu\text{m}$.

3. Results of Numerical Simulation and Discussion

The model of modified finite-difference solution of nonlinear Maxwell equations is used for the numerical integration of this system. For the numerical modeling of the processes described by equations (1), (2), (6), (17)-(19), it's passed to the mesh functions for E_x , E_z , H_z and H_x fields, electric inductions D_x , D_z , linear and nonlinear responses, for which the grids are set over the coordinate $k \cdot \Delta y$ and time $n \cdot \Delta t$. The step of spatial grid Δy was chosen equal to $\lambda_0/400 = 2.025 \text{ nm}$. The step of time grid is determined by Kurant's [8] condition and is equal to $\Delta t = \Delta y/2c = 3.375 \cdot 10^{-3} \text{ fs}$. For such a time step the linear part of the scheme has a dispersion that approaches the Lorentz dispersion of the medium as much as possible. The numerical estimation shows that relative error of phase velocity is 0.039% and relative error of group velocity is 0.046%. Figure 1 shows the evolutions of time-profiles (a), (b) and (c) for the initial x-ordinary polarized $E_x(t, y)$ pulse, z-extraordinary polarized pulse $E_z(t, y)$ (SFR and DFR) and z-extraordinary polarized pulse $E_{z\text{DF}}(t, y)$ (only DFR) at PPLN output at $E_{x0} = 600 \text{ MV/m}$ and at the different time step values n .

Numerical simulation was performed for the following initial conditions:

$$E_x(t, y=0) = E_{x0} \cdot \exp(-t^2/\tau_{x0}^2) \cdot \cos(2 \cdot \pi \cdot c \cdot t/\lambda_{x0}), E_z(t, y=0) = 0 \quad (24)$$

where $E_{x0}=600$ MV/m - amplitude of x polarized pulse, $\tau_{x0} = 10$ fs are the durations of the pulse, $\lambda_{x0} = 0.81 \mu\text{m}$ central wavelength.

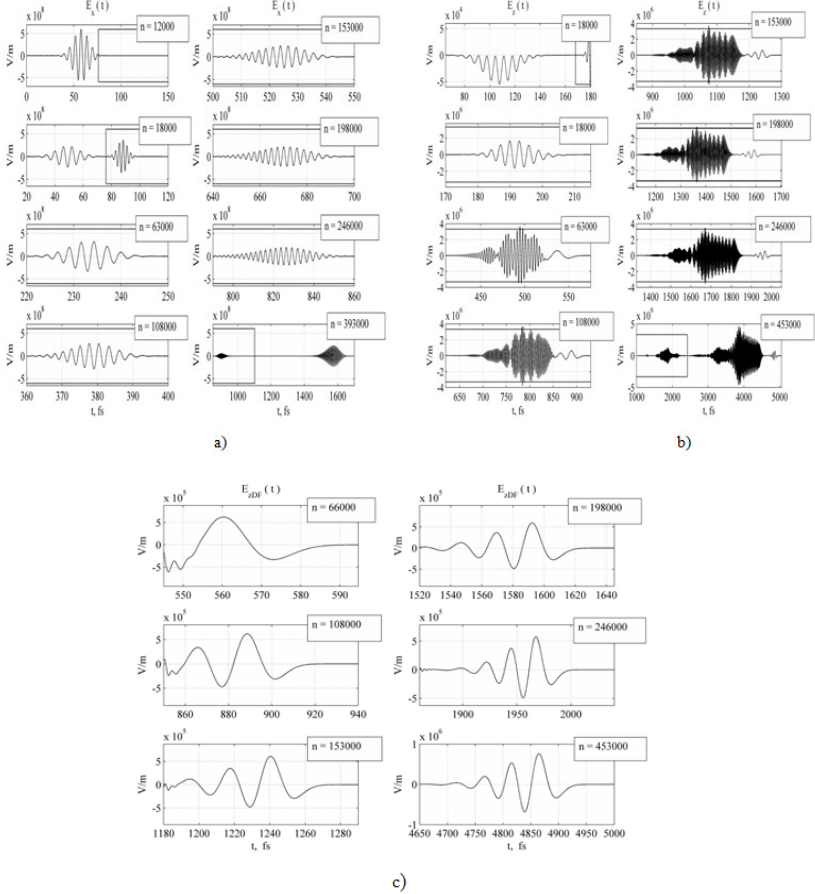


Figure 1. The evolutions of time-profiles of x-polarized and z-polarized pulses.

Figure 2 shows the evolutions of spectrum-profiles for the initial

x-ordinary polarized $E_x(t,y)$ pulse (dashed line - 1) and z-extraordinary polarized pulse $E_z(t,y)$ (solid line - 2).

As seen from figure the spectral distribution for the DFR monotonically increase from 1.1 μm up to 4 μm with the -49.77 dB normalized spectral distribution level at 3 μm and for SFR we observe spectral peaks at 0.4147 μm , 0.3591 μm , 0.3307 μm , 0.3115 μm wave lengths and at -22.41 dB, -30.38 dB, -48.92 dB, -67.99 dB levels respectively. The generation of spectral components with wavelengths less than 0.38 μm can be explained by cascade generation of high order harmonics.

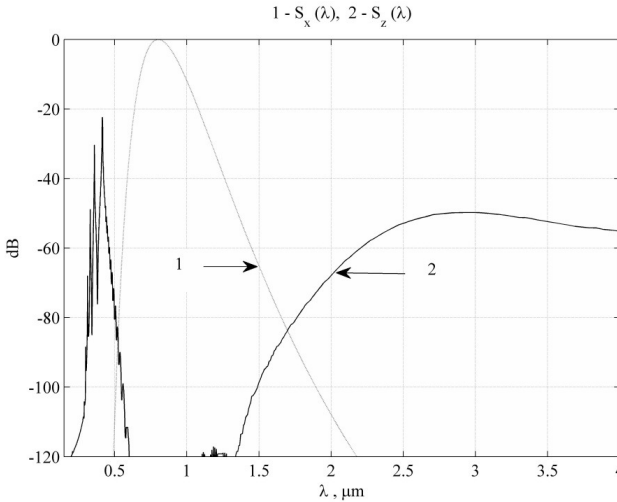


Figure 2. The spectral profiles for the initial x-ordinary polarized $E_x(t,y)$ pulse (dashed line - 1) and z-extraordinary polarized pulse $E_z(t,y)$ (solid line - 2).

4. Conclusion

The generation of DFR in wavelength range (3–4) μm that occurs during the propagation of an ordinary linearly polarized laser pulse of a few optical cycles with duration 10 fs at central wavelength 0.81 μm in the direction normal to the optical axis of a PPLN where studied. Considered the PPLN consisted from 10 period's with 12.6 μm thicknesses. We solve Maxwell's equations to describe the propagation of the laser pulse using the FDTD method. The time profile and spectrum distri-

bution of extraordinary linearly polarized pulse, arising due to nonlinear interaction of laser pulse with PPLN, are obtained.

References

- [1] Y.Lee, Principles of Terahertz Science and Technology (Springer, 2009).
- [2] S. L. Dexheimer, Terahertz Spectroscopy Principles and Applications (Taylor & Francis Group, LLC., 2008).
- [3] K. Sakai, Terahertz Optoelectronics (Springer-Verlag Berlin Heidelberg, 2005).
- [4] T.Skauli, P.S.Kuo, K.L.Vodopyanov, T.J.Pinguet, O.Levi, L. A.Eyres, J. S.Harris, M. M.Fejer, E.L.Ginzton, B.Gerard, L.Becouarn, E.Lallier, Journal of Applied Physics, 94, 10, (2003) 6447.
- [5] R. W. Boyd, Nonlinear Optics (Academic Press; 3 edition, 2008).
- [6] D. L. Hovhannisyan, A. A. Hakhoumian, R. M. Martirosyan, A. S. Nikoghosyan, E. M. Laziev, G. D. Hovhannisyan, Journal of Modern Optics, 57, 12, (2010), 1075.
- [7] V. G. Dmitriev, L. V. Tarasov, Applied Nonlinear Optics (Moscow Fizmatlit, 2 edition, 2004).
- [8] A.Taflov, S.C.Hagness, Computational Electrodynamics the Finite-Difference Time-Domain Method (Artech House, Boston, 1995).

THz radiation in LiNbO₃ rectangular plate waveguide

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A simulation study for visualizing the wave propagation in LiNbO₃ plate waveguide aimed to develop an effective active THz waveguide system has been performed. The finite-element method is used in the computational technique. It is shown that when the THz radiation propagates 4 mm in a rectangular crystal plate with frequency of 0.27 THz its mode structure is changed.

I. Introduction

The absence of effective, large bandwidth at THz frequency waveguides with low losses and dispersion is a basic barrier hinder for the development and implementation of a device for applications of THz radiation in different fields of science because the THz waveguides are key components of many electronic circuits operating in the frequency range from $0.3 \div 10$ THz. A variety of methods to THz wave guiding have been demonstrated over the past years. Researchers have studied several THz waveguides: hollow metal tubes [1,2], parallel metal plates [3], porous-core honeycomb fiber [4], metal wire [5,6], plastic ribbon waveguides and photonic crystal fibers [7,8] hollow polymer waveguides with inner metallic layers [9] plastic photonic crystal fibers (PPCF) [10], subwavelength plastic fiber [11], waveguide for guiding superconducting carriers or HTSC waveguide [12], LiNbO₃ plate tapered waveguide [13,14].

Dispersion of the waveguide can cause extreme distortions of the input ultrashort THz pulse. Since the waveguide is a linear system, one may represented the ultrashort pulse as a sum of quasi-monochromatic spectral components using Fourier transform. One can research and analyze the propagation of ultrashort pulses in waveguide by studying the propagation of the spectral components of the pulse.

In this paper the finite-element method is applied in simulation to investigate the wave propagation with frequency of 0.27 THz in a

rectangular LiNbO₃ plate waveguide. The results obtained are compared with the findings done for the rectangular LiNbO₃ tapered waveguide [14]. In the both cases the cross-section of the crystals was the same - 0,27x1mm². The results can be used to create an effective THz broadband active waveguide system.

2. Simulations of guided wave propagation at 0.27 THz frequency in rectangular LiNbO₃ plate

A computing apparatus based on finite element method (FEM) was applied for imitating and imaging THz wave propagation. To simulate and excite THz wave propagation in the waveguide made of LiNbO₃ rectangular plate placed in free space, the program “COMSOL Multiphysics” is used [14]. Dielectric plate is a surface wave line (for example, Sommerfeld single-wire line, reflective line, etc.) has the least attenuation and the best frequency dependence.

The boundaries of the waveguide separate the spatial volume of simulation, in our case it is 3D space. The simulation domain is divided into tetrahedral cells. The system of elements together, with given physical parameters and boundary conditions, forms a system of linked partial differential equation. To solve correctly the equation system which describes the distribution of THz wave field, the largest size of the cell should not be greater then one fifth of the wave length. In our case it is 50 μm. The following values were inputted into the program: the real and the imaginary parts of the dielectric susceptibility ($\epsilon_z=27,96$ and $\epsilon'(270 \text{ GHz})=0.102$, correspondingly) and the power for the given THz extraordinary wave. These values calculated through the formula given in [15] were inputted into the program. The THz wave source was located on the input surface of LiNbO₃ plate with the cross-section of 0.27x1mm² and the length of 4 mm. We assumed that THz waves have the polarization parallel to the optical axis (X-axis) of the anisotropic crystal LiNbO₃. This takes place when an optical source generates THz radiation in LiNbO₃ plate due to the difference frequency generation or the optical rectification of the femtosecond laser pulse. When the optical radiation polarization direction is the same as the direction of the optical axis of the non-linear crystal, one

uses the largest second-order nonlinear tensor d_{33} in the nonlinear conversion process, and the THz radiation generated has the highest power. Any non-linear optical crystals (e.g., GaP, InP, GaAs, GaSe, DAST, etc.) with selected high non-linear second-order susceptibility and having low absorption coefficients and low dispersion may also be used as active waveguides for ultra-high-speed electronic integrated circuits.

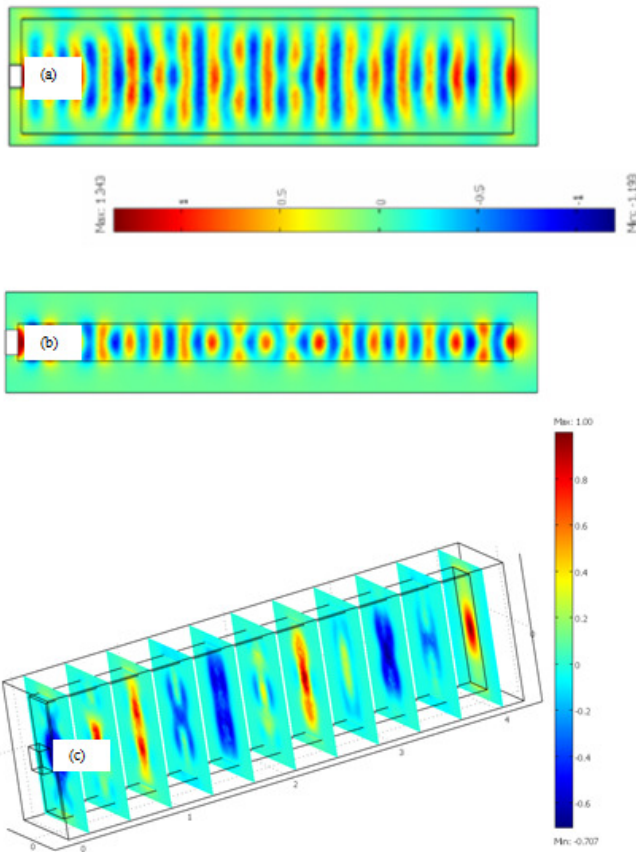


Fig.1. Distribution of the THz electric field E_x component during propagation along the z axis with frequency of 270 GHz: (a) lateral view in (xz) plane, (b) up view in (yz) plane, (c) spatial distribution of THz field in 11 cross-sections of the LiNbO_3 waveguide.

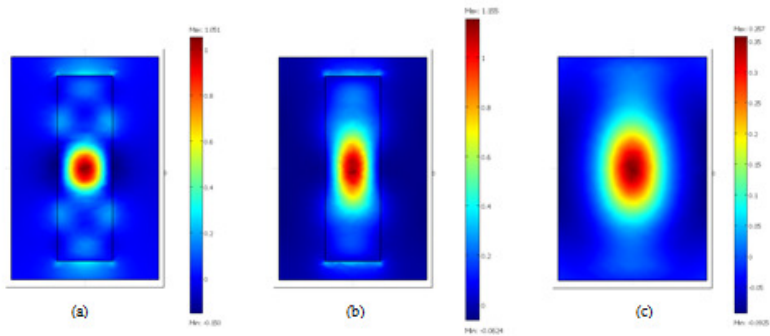


Fig.2. The spatial distribution of the THz electric field in (xy) cross-section is shown (a) at the entry of the waveguide ($L=0.05$ mm), (b) at the end of the waveguide ($L=3.99$ mm) and (c) out of the waveguide ($L=4.1$ mm).

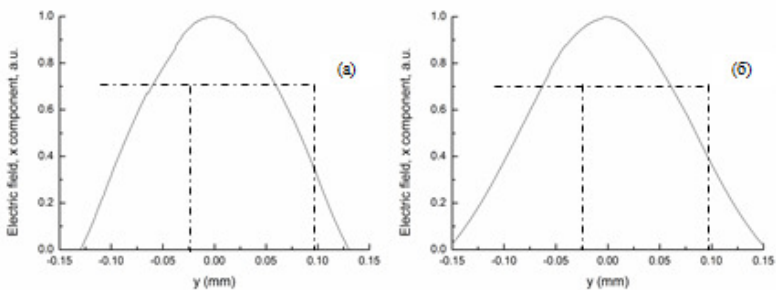


Fig.3. Distribution of the THz electric field $E_x(z)$ -component along the Y axis, demonstrated on the distance of (a) $z = 0.05$ mm and (b) $z = 4$ mm from the entry surface of the LiNbO_3 slab.

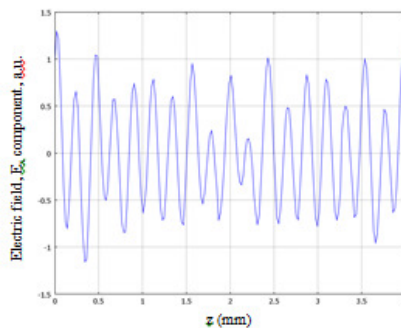


Fig.4. Graph of the $E_x(z)$ -component of the THz electric field (frequency - 270 GHz, crystal length $L=4$ mm)

The mode structure of the THz electric field along the z axis, as well as in 11 cross-sections of the LiNbO_3 waveguide is given in Fig.1.(a, b, c). Fig.1 demonstrates how the mode changes its type along the propagation direction, i.e. it shows transition from the main mode E_{11}^x to the highest mode. In the cross-sections, distributions of fields are symmetric relative to the center of the plates ($z=2$ mm), as well as relative to the X , Y , Z axis. Figure 2 shows the spatial distribution of the THz electric field in (xy) plane inside and out of the waveguide. On Fig.1 and Fig.2 the red color indicates positive and the blue color indicates negative values of the THz electric field. Distribution of the THz electric field $E_x(z)$ -component along the Y axis at frequency of 270 GHz is presented on Fig.3. While comparing the spatial sizes of the THz field (at $1/\sqrt{2}$ level of the $E_x(y)$'s maximum) by the entry and exit surfaces of the waveguide (see Fig.2 and Fig.3), one can see that the width of the $E_x(y)$ component distribution along the Y -axis is the same but the width of the field distribution along the X -axis increases as the distance increases, and takes the form of ellipse in the (xy) cross-section. Fig.4 demonstrates stable propagation of the amplitude-modulated THz wave. Therefore, comparing the results of the present work and the results of [14], we conclude that in order to concentrate the THz field, provide a good impedance match, as well as get almost all energy out of the waveguide one has to use a tapered- to- the-point waveguide.

3. CONCLUSIONS

In the paper a finite-element method is applied in simulation to investigate the wave propagation with frequency of 0.27 THz in a rectangular LiNbO_3 plate waveguide. The results may be used to construct an efficient active THz waveguide system. The full energy of the THz radiation propagating along straight lines parallel to the Z -axis of the plate has the both – external (outside of the plate) and internal fields (Fig.1, Fig.2). Such a way of the full-field transmission allows to efficiently generate the THz radiation in nonlinear optical crystal waveguide if for given optical and THz frequencies the collinear phase-matching condition is satisfied. Phase-matching may be achieved by

choosing the thickness of the plate equal to value obtained from the transcendent dispersion equation in [13]. The thickness of the plate uniquely determines the phase matching condition. However, one can provide the phase matching condition for THz waves at other frequencies as well by choosing appropriate a/b ratios. Less is the thickness of the crystal, higher are frequencies, at which the phase-matching condition is satisfied. The ratio b/a of the waveguide affects the propagation loss. Larger is the ratio, less is the propagation loss. Influence of diffraction on the THz wave propagation in waveguide depends on the size and form of the crystal's cross-section. The diffraction degrades the quality of the THz pulse beam at the exit face of the crystal by spatially chirping it. In order to mitigate the detrimental effects of diffraction, studies in paper [16] have shown that the pump spot size should be roughly equal to the peak generated THz wavelength inside the crystal.

Excitation of the THz radiation in the waveguide with the help of optical laser pulse, by optical rectification [17,18], allows resolving problems connected with input/output coupling – mode matching and single mode propagation. In case of a rectangular waveguide, since the LiNbO_3 crystal has a high reflection factor, about 46% of the THz radiation reflects from the exit surface of the LiNbO_3 plate. One can provide full output coupling from the waveguide by shaping it tapered to the point [13, 14].

In order to use a nonlinear crystal as an efficient active plate waveguide to obtain the highest possible THz radiation (continues or pulse) power in the near-field zone, one should use a crystal with: (a) high nonlinear second-order susceptibility ($P \sim \chi^2$); (b) low absorption coefficient, that restricts the waveguide's length and as a result - the THz radiation power ($P \sim L^2$); (c) optimal thickness-to-height ratio of the waveguide (a/b), (d) low dispersion and broad velocity-matching bandwidth. And finally, in order to concentrate the THz field, to provide a good impedance match, as well as to get almost whole energy out of the waveguide, one should use a waveguide with tapered-to-the-point exit face.

REFERENCES

1. R.W. McGowan, G. Gallot, and D. Grischkowsky. Propagation of ultra-wideband, short pulses of THz radiation through submillimeter-diameter circular waveguides. *Opt. Lett.*, **24**, 1431–1433, (1999).
2. G. Gallot, S.P. Jamison, R.W. McGowan, and D. Grischkowsky. Terahertz waveguides. *J. Opt. Soc. Am B* **17**, 851–863, (2000).
3. R. Mendis, D. Grischkowsky. THz interconnect with low loss and low group velocity dispersion. *IEEE Microwave Wireless Comp. Lett.* **11**, 444–446, (2001).
4. H. Bao, K. Nielsen, H.K. Rasmussen, P.U. Jepsen, O. Bang. Fabrication and characterization of porous-core honeycomb bandgap THz fibers. *OPTICS EXPRESS*, **20**, No. 28, 29508, (2012).
5. K. Wang and D.M. Mittleman. Metal wires for terahertz wave guiding. *Nature*, **432**, 376–379, (2004).
6. Jeon, J. Zhang, and D. Grischkowsky. THz Sommerfeld wave propagation on a single metal wire. *Appl. Phys. Lett.*, **86**, 161904/1–3, (2005).
7. R. Mendis and D. Grischkowsky. Plastic ribbon THz waveguides. *J. Appl. Phys.*, **88**, 4449, (2000).
8. S.P. Jamison, R. W. McGowan, D. Grischkowsky. Terahertz single-mode waveguide propagation and reshaping of sub-ps terahertz pulses in sapphire fibers. *Appl. Phys. Lett.*, **76**, 1987–1989, (2000).
9. J. Harrington, R. George, P. Pedersen, E. Mueller. Hollow polycarbonate waveguides with inner Cu coatings for delivery of terahertz radiation. *Optics Express*, **12**, Issue 21, pp. 5263–5268, (2004).
10. H. Han, H. Park, M. Cho, J. Kim. Terahertz pulse propagation in a plastic photonic crystal fiber. *Appl. Phys. Lett.* **80**, 2634–2636, (2002).
11. L.J. Chen, H.-W. Chen, T.-F. Kao, J.-Y. Lu, and C.-K. Sun, “Low-loss subwavelength plastic fiber for terahertz waveguiding,” *Opt. Lett.* **31**, 308–310 (2006). C.-K. Sun, *Opt. Lett.*, (Oct. 2005)
12. H.P. Roeser, A. Bohr, D.T. Haslam, J. S. López, M. Stepper, F.M. Huber, A.S. Nikoghosyan. Correlation between crystal structure and resistivity of high-temperature superconducting cuprates *Armenian Journal of Physics*, **4**, issue 2, 109–133, (2011).

13. E.M Laziev, A.S. Nikoghosyan. Phaze Synchronizm and Ultrashort SHF Pulses Generation in Waveguides. SPIE, Mode-Locked Lasers and Ultrafast Phenomena, **1842**, 113-118, (1991).
14. A.S. Nikoghosyan, H.P. Roeser, A.S. Akopyan, A. Bohr, D.T. Haslam, J.S. López, M. Stepper. THz radiation in the LiNbO₃ crystal tapered waveguide. Proceeding of IRPhE'2012 Confer., Yerevan, Armenia.
15. N.N. Zinov`ev, A.S. Nikoghosyan, J.M. Chamberlain. Terahertz Radiation from a Nonlinear Slab Traversed by an Optical Pulse. Physical Review Letters, **98**, 044801-1 044801-4, (2007).
16. G. Chang, C. J. Divin, J. Yang, M.A Musheinish, S.L. Williamson, A. Galvanauskas, T.B. Norris. GaP waveguide emitters for high power broadband THz generation pumped by Yb doped fiber lasers. Optics Express, **15**, No. 25, 16308-16315, (2007).
17. A.S. Nikoghosyan. Laser Driven Terahertz Dielectric Wedge Antenna Placed in Free Space or in Hollow Metallic Waveguide. Inter. Conf. IRMMW-THz 2010 at Rome, Italy, September 5-10, Tu-P.54-55 (2010).
18. A.S. Nikoghosyan, M.A. Kazaryan, V.I. Sachkov. Generation of THz pulse in a nonlinear crystal GaSe, partially filling the waveguide. Atmospheric and Oceanic Optics, **25**, 264-267, (2012).

Detection of laser radiation in optically transparent ferromagnet

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In this paper we have experimentally investigated the detection of laser radiation in the transparent ferromagnet, obtained due to the excitation of the magnetic moment of a magnetized ferromagnet by induction of the magnetic field of the laser radiation. As a ferromagnet YIG crystal was used, which has transparency window in the near infrared range. The femtosecond laser as a modulated radiation source was used. The polarization of the laser radiation was linear, the tuning range of the wavelength of 710-950 nm. The effective detection is obtained only in the range of relative transparency (900 - 950 nm) of a ferromagnetic crystal. In the absence of an external magnetic field, detected signal is also absent. Signal reaches the maximum value in the region of maximum change of the slope of the magnetization curve of a ferromagnetic sample. The reversing of the external magnetic field changes the sign of the detected signal.

1. Introduction

Properties of ferromagnets, due to the nonlinearity of the magnetic susceptibility of the medium in the low-frequency and microwave fields have been well investigated [1,2]. These phenomena have great practical and scientific value in the areas of signal detection, frequency conversion and radiation control. The first such phenomenon were observed by Bloembergen and Damon [3] in nickel ferrite powder with pellets of 0.5 mm diameter in the radiation field of the magnetron in the 10 GHz frequency band.

There are also a lot of transparent ferromagnetic materials in the infrared and visible range that have been successfully used to control radiation [4,5]. However, the nonlinear properties of ferromagnetic

materials in the infrared and visible ranges have not been investigated. It is believed [6,7] that in this range magnetic permeability of the medium is equal to unity and therefore the medium does not exhibit nonlinear magnetic properties in the field of the electromagnetic wave.

In the recent years phenomena of reorientation of the magnetic moment in ferromagnetic materials by laser radiation have been investigated. The main reason for magnetic reorientation is considered the local heating of the sample or the inverse Faraday effect [8,9]. It has been shown that excitation of a magnetically ordered material with ultrashort femtosecond laser pulses may result in demagnetization and spin-reorientation on a timescale of sub-picoseconds [10].

In [11] it was shown that the magnetization of ferromagnetic materials can be fully controlled by circularly polarized femtosecond laser pulses, without application of an external magnetic field. This is explained by the inverse Faraday effect. The emerging magnetic field because of this effect can reorient the magnetic moment much faster than the above mentioned thermal effect. In another study [12] it is shown, that the magnetization is much better controlled by linearly polarized pulse.

In [19] nonlinear interaction of laser radiation with optically transparent ferromagnet has been investigated. In the result of such interaction the optical rectification was obtained.

In this paper we have investigated the detection of infrared laser radiation in a ferromagnetic crystal Yttrium-Iron-Garnet (YIG).

2. The equation of motion of the magnetic moment in a variable magnetic field

The magnetic polarizability of the medium is described by equations, based on the equation of motion of the magnetic moment in a *constant* magnetic field [13] (Larmor's theorem)

$$\frac{d\mathbf{M}}{dt} = -\gamma[\mathbf{M} \times \mathbf{H}] \quad (1)$$

where $\omega_L = \gamma H = (q/2mc)H$ – Larmor angular velocity of rotation. If we multiply equation (1) on $\mathbf{M}$ scalarly, we get $\mathbf{M}^2 = \text{const}$, and it becomes obvious that this equation does not describe the excitement of the moment in a variable magnetic field.

Nonlinear magnetic properties of ferromagnets are described by the nonlinear dependence of the magnetic susceptibility on the applied magnetic field $\mathbf{H}$. To describe the properties of ferromagnets in an variable magnetic field, the equation of motion of the magnetic moment in a *constant* magnetic field (1) with various dissipative terms [14-17] is used, which are essentially non-linear. Part of the equation is a modified Landau-Lifshitz equation[14]

$$\frac{d\mathbf{M}}{dt} = -\gamma[\mathbf{M} \times \mathbf{H}] - \gamma \frac{\alpha}{|\mathbf{M}|} [\mathbf{M} \times [\mathbf{M} \times \mathbf{H}]] \quad (2)$$

and others - modified Bloch equation [15]

$$\frac{d\mathbf{M}}{dt} = -\gamma[\mathbf{M} \times \mathbf{H}] - \frac{\mathbf{M} - \mathbf{M}_0}{\tau} \quad (3)$$

Bloembergen modified Bloch equation [16]

$$\frac{d\mathbf{M}}{dt} = -\gamma[\mathbf{M} \times \mathbf{H}] - \frac{\chi_0 \mathbf{H} - \mathbf{M}_0}{\tau} \quad (4)$$

where $\mathbf{M}$ in the last term is presented the dependence on the applied magnetic field in the form $\mathbf{M} = -\chi_0 \mathbf{H}$.

Gilbert modified Landau-Lifshitz equation[17]

$$\frac{d\mathbf{M}}{dt} = -\gamma[\mathbf{M} \times \mathbf{H}] + \frac{\alpha}{|\mathbf{M}|} \left[\mathbf{M} \times \frac{d\mathbf{M}}{dt} \right] \quad (5)$$

obtained replacing the expression $-\gamma[\mathbf{M} \times \mathbf{H}]$ by $d\mathbf{M}/dt$ in the last term of equation (2).

Cullen modified Landau-Lifshitz equation[18]

$$\frac{d\mathbf{M}}{dt} = -\gamma[\mathbf{M} \times \mathbf{H}] - \chi \mathbf{M} - \lambda [\mathbf{M} \times [\mathbf{M} \times \mathbf{H}]] \quad (6)$$

in essence differs from equation (2) addition of the linear relaxation term in the form $-\chi \mathbf{M}$.

In [20] proposed an equation of motion of the magnetic moment in an *variable* magnetic field, in which appears excitation of the magnetic moment $\mathbf{M}$ by induction of the magnetic field $\mathbf{H}$

$$\frac{d\mathbf{M}}{dt} = -\gamma[\mathbf{M} \times \mathbf{H}] - \gamma^2 I \frac{d\mathbf{H}}{dt}, \quad (7)$$

where I -moment of inertia. Taking the scalar on $\mathbf{M}$, we obtain an expression

$$\frac{d}{dt} \left(\frac{\mathbf{M}^2}{2} \right) = -\gamma^2 I \mathbf{M} \frac{d\mathbf{H}}{dt}, \quad (8)$$

where $\mathbf{M} = (q/2mc)\mathbf{L}$ - magnetic, and $\mathbf{L}$ - mechanical momentum of the particle. This shows why and how the rotation kinetic energy $\mathbf{L}^2/2I$ of particles with magnetic moment $\mathbf{M}$ in an variable magnetic field $\mathbf{B}$ changes. As we see, in the case of variable magnetic field in the equations is missing a member that is responsible for the excitation of the magnetic moment by induction of variable magnetic field.

3. Experimental setup and measurement

We have experimentally investigated detection of laser radiation in a transparent ferromagnet YIG, obtained due to the excitation of the magnetic moment of a magnetized ferromagnet by the induction of magnetic field of the laser radiation.

As a source of modulated radiation we used femtosecond Ti:sapphire laser Mai-Tai, from Spectra-Physics. Average power of linearly polarized laser radiation was 1.1-1.8 W (depending on wavelength), laser pulse duration of ~100fs and a repetition rate of ~80MHz. The tuning range of laser wavelength is 710-950nm.

To study the interaction of optical radiation with a ferromagnet it is necessary that the test material is transparent enough for such a band. As a material was used ferromagnetic crystal YIG, which is transparent in the infrared range.

Crystal YIG (chemical formula $\text{Y}_3\text{Fe}_5\text{O}_{12}$) is a perfect two-sublattice ferromagnet, prototype of ferrogarnet. It has the following electrical and magnetic properties: saturation magnetization – $4\pi M_0 = 1750$ Gs, crystalline anisotropy field – $H_A = 42$ Oe, Curie temperature – $T_C = 556$ K, the resistivity – $\rho = 10^{14}$ $\Omega\cdot\text{cm}$. At 300 K the thermal conductivity of the crystal YIG in the directions [110], [211], [111], [100] is equal to 8.0, 7.2, 7.8, 6.7 W·m·K, respectively. Specific heat – 432.41 J·mol·K [21].

Optical properties are: transparency band - 1.1-5.5 μm , the refractive index in the transparency band of 1.4 to 5.5 μm varies from 2.209 to 2.103, and in the 0.6 to 1.0 μm - 2.2 to 2.4. Faraday effect in the transparent window– 1.08 Rad·cm⁻¹. The absorption coefficient in the transparent band – $\gamma = 0.03\text{-}0.1$ cm⁻¹. Dependence of the absorption

coefficient on the length of the electromagnetic wave is shown in Fig. 1 [21].

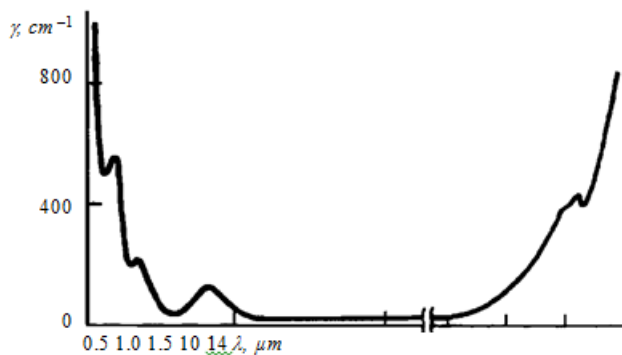


Fig. 1. Dependence of the absorption coefficient of YIG on the wavelength.

Schematic view of the experimental setup is shown in Fig. 2.

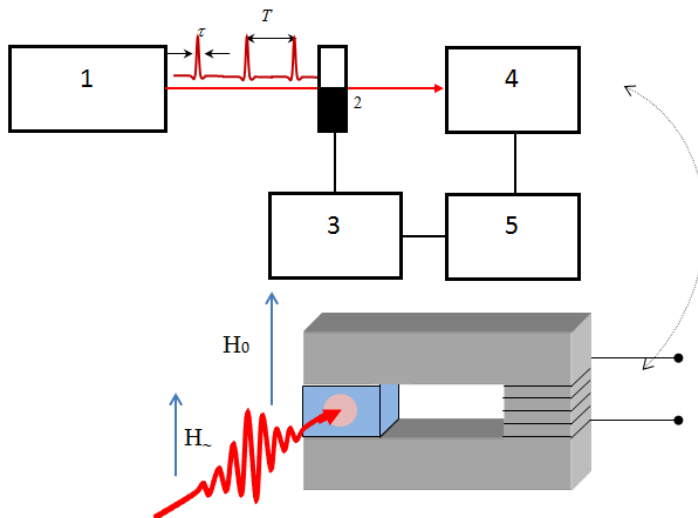


Fig. 2. The experimental setup.

Ferromagnetic sample is located in the path of the laser light in such a way that the direction of the magnetic moment of the magne-

tized ferromagnetic in an external field H_0 coincides with the direction of the magnetic field of the linearly polarized laser light. As follows from the equations of motion of the magnetic moment (7), such an arrangement could lead to non-linear change of the magnetic moment of the magnetized ferromagnetic by electromagnetic waves. In particular, when the ferromagnet is magnetized to saturation, excitation by electromagnetic wave (see equation (8)) can only lead to a decrease in the magnetic moment of the ferromagnet due to partial reorientation of the magnetic moment in the opposite direction with respect to H_0 . As a result, the average value of the magnetic moment decreases.

To register the change in the average value of the magnetic moment of the sample magnetic sensor 4, in the form of a horseshoe with a coil inductance wrapped around a ferrite was used. The sensor is attached to the magnetized sample YIG, as shown in Fig. 1.

The change of the magnetic moment in the crystal YIG under laser radiation leads to a change of the magnetic flux in the magnetic sensor, which induces a voltage in the coil. Voltage across the inductor (detected signal) is recorded by oscilloscope 5 Agilent Technologies DSO7012B (100 MHz, 2 Gs/s).

The repetition rate of the laser pulse is high ($\sim 80\text{MHz}$) and for the registration of the change of the magnetic moment with the above-described sensors low-frequency (1 kHz) modulation of the laser radiation with a mechanical chopper 2 was used (see Figure 1).

As show measurement, the effective detection is obtained only in the range of relative transparency (900 - 950 nm) of a ferromagnetic crystal (Fig. 3).

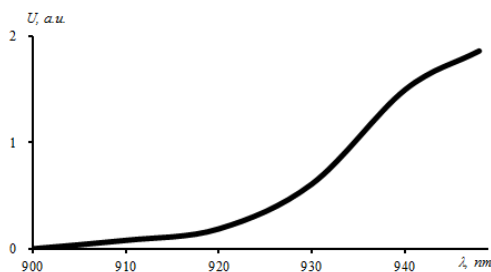


Fig. 3. The dependence of the detected voltage across the inductance of the laser wavelength.

In the absence of an external magnetic field, detected signal is also absent. Signal reaches the maximum value in the region of maximum change of the slope of the magnetization curve of a ferromagnetic sample (see Figure 4.). Changing the direction of the external magnetic field to the opposite leads to a change in the sign of the detected signal (Figure 4.).

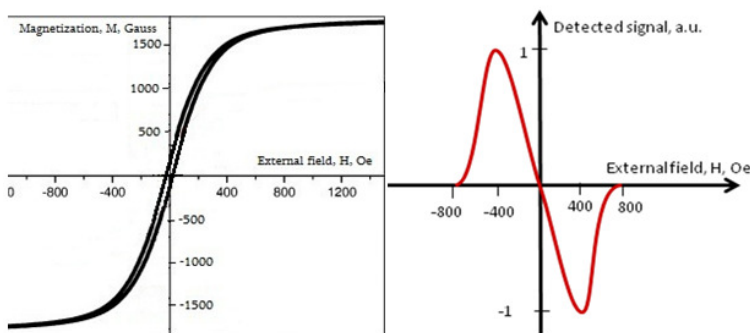


Fig. 4. Magnetization curve (a) and the dependence of the detected signal on the external magnetic field (b).

4. Discussion of the results and conclusions

Experimentally detection of linearly polarized amplitude-modulated laser radiation in infrared transparent ferromagnetic YIG is obtained. It is assumed that the reorientation of the magnetic moment of the medium occurs by induction of variable magnetic field of the laser.

The dependence of the detected signal from the laser wavelength (Figure 3.) indicates that the effective detection is obtained only in the relative transparency of the ferromagnet, and therefore the interaction occurs in the sample and is not thermal in nature. This is indicated by the dependence of the detected signal on the polarization of the laser radiation.

It is shown that the effective nonlinearity is seen in the case of coincidence of the magnetic field of a linearly polarized laser beam

with the direction of the magnetic moment of a magnetized ferromagnetic.

The presence of maxima of detected signal and changes of signal badge, depending on the external magnetic field H_0 (Figure 4.) correlate well with the magnetization curve of the crystal YIG (Fig. 4b). Maximum detected signal observed when the magnetization of the ferromagnet is approaching saturation magnetization. At full saturation probability of reorientation of the magnetic moment under the action of laser radiation is greatly reduced, thus reducing the detected signal. When the ferromagnet is not magnetized, changes in the average value of the magnetic moment under the action of laser radiation do not occur, since reorientation in both directions is the same.

In conclusion we remark that nonlinear interaction of laser radiation with ferromagnets can find practical application in frequency conversion of laser radiation, recording and storage of information, etc.

REFERENCES

- [1] *Ферриты в нелинейных СВЧ устройствах*, Сб.-к статей под. ред. А.Г. Гуревича (Москва, ИИЛ, 1961).
- [2] *Нелинейные свойства ферритов в полях СВЧ*, Сб.-к статей под. ред. А.Л. Микаэляна (Москва, ИИЛ, 1963).
- [3] N. Bloembergen, R.W. Damon, Phys. Rev., **85**, (1952), 699.
- [4] Г.С. Кринчик, М.В. Четкин, УФН, т. **98**, 1, (1969), 3.
- [5] С. Крупичка, Физика ферритов, т.т. 1-2, (Москва, МИР, 1976).
- [6] Л. Д. Ландау, Е. М. Лифшиц, Электродинамика сплошных сред (Москва, Физматлит, 2005).
- [8] Л.Д. Ландау, Е.М. Лифшиц, Статистическая физика, часть 2 (Москва, Физматлит, 2004).
- [9] J. Hohlfeld, Th. Gerrits, M. Bilderbeek, Th. Rasing, H. Awano, and N. Ohta, Phys. Rev., B 65, (2001), 012413.
- [10] A. V. Kimel, A. Kirilyuk, P. A. Usachev, R. V. Pisarev, A. M. Balbashov, and Th. Rasing, Nature, **435**, (2005), 655.
- [11] W. A. Challener, T. W. McDaniel, C. D. Mihalcea, K. R. Mountfield, I. K. Pelhos and Sendur, Jpn. J. Appl. Phys., **42**, (2003), 981.
- [12] A. Kirilyuk, Journal of Science and Arts Year 11, 3(16), (2011), 353-356

- [13] A.M. Kalashnikova, A.V. Kimel, R.V. Pisarev, V.N. Gridnev, A. Kirilyuk, and Th. Rasing, Phys. Rev. Lett., **99**, (2007), 167205.
- [14] Л. Д. Ландау, Е. М. Лифшиц, *Теория поля* (Москва, Физматлит, 2003).
- [15] L. D. Landau, E.M. Lifshitz, Phys. Zs., Sowjetunion, **8**, (1935), 153.
- [16] F. Bloch, Phys. Rev., **70**, (1946), 460.
- [17] N. Bloembergen, Phys. Rev., **78**, (1950), 572.
- [18] T. L. Gilbert, Phys. Rev., **100**, (1955), 1243.
- [19] Н. В. Callen, J. Phys. Chem. Solids, **4**, (1958), 256.
- [20] А. О. Макарян, Молодой научный сотрудник, ЕГУ, 2(36), (1982), 134-136
- [21] В. М. Мыхитарян, Известия НАН Армении, Физика, **47**, 6, (2012), 379.
- [22] Ю.М.Яковлев, С.Ш.Генделев, *Монокристаллы ферритов в радиоэлектронике* (Москва, Сов. Радио, 1975).

Microwave-photonic receivers – new conception in radio-communication

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The reasons of advent of new type combined microwave-photonic receivers in the well established radio-frequency communication area are analysed. Advantages of microwave-photonic receivers' architecture against traditional full electronic radio-frequency receivers are discussed. Existing realisation of microwave-photonic receivers with the key-element as a highly sensitive electro-optical modulator based on LiNbO₃ microdisk is described. An alternative electro-optical modulator based on Fabry-Perot type microresonator is considered and its main advantages are presented.

1. Introduction

Microwave (MW) communication systems are important part of telecommunication which expand rapidly the frequencies of their operation. Increase of operating frequencies is advantageous on the subject of communication speed; however it revealed the serious problem in contemporary receivers' construction. Now the most part of communication systems are exploiting superheterodyne radio-frequency (RF) receivers. They best satisfy the requirements of modern communication systems, as inherently possess higher selectivity and sensitivity compared to the other types of receivers [1]. However with the increase of operating frequency the stray radiation of heterodyne (local RF

oscillator) is increased. This parasitic radiation becomes a source of interference for neighbouring radio devices. Moreover, by means of this parasitic radiation it is possible to locate the covert radio receiver and recover its operating frequency. All this brings to the necessity of serious revision of receivers' construction.

There are several ways to reduce this stray radiation. One of them is the use of a stop-band filter on the way to antenna. The presence of this filter brings to the lessening of parasitic radiation, but does not eliminate it. Parasitic radiation may be reduced also at the expense of the schematic complication of the superheterodyne receiver [2]. While this method increases the size of the receiver and its power consumption it can't completely get rid of parasitic radiation. As the proper way out, the transition to the intermediate optical range has been suggested [3-7]. The essence of this approach is in application of optical local oscillator (laser) instead of RF one. In this type of receivers an incoming MW carrier, loaded by low frequency signal, after passage through resonant RF input circuit is converted to the optical domain where advantages of optical signal processing (OSP) can be exploited [8-10]. In the end a photo-detector retrieves the comparatively low-frequency signal. In this kind of combined receiver, so called microwave-photonic (MW-photonic) receiver, besides getting rid of parasitic radiation in RF range, it is possible to attain the requiring sensitivity, selectivity and bandwidth, while having immunity to the external electromagnetic stray radiation, small size, weight and power consumption [3-7,11-14].

The block-diagram of a MW-photonic receiver is presented in Fig. 1. The key element of MW-photonic receiver is an electro-optical modulator (EOM). To ensure high sensitivity of this type combined receivers it is necessary to have high Q-factor in RF input part and high efficiency in electro-optical transformation. While high Q-factor MW input circuitry is well-established, the problem was to find the proper EOM which would ensure strong interaction between electrical and optical waves. This is possible only in optical resonant structure that permits to prolong electrical field interaction with optical wave confined within the resonator.

The suitable optical resonators are: Fabry-Perot type and disk (or ring) one. The confinement of optical wave in Fabry-Perot resonator depends on the reflectance of mirrors serving also for light input and output from the resonator. Higher is the mirrors' reflectance, stronger is the confinement of light within the resonator. The light survive time within the resonator is proportional to the light confinement.

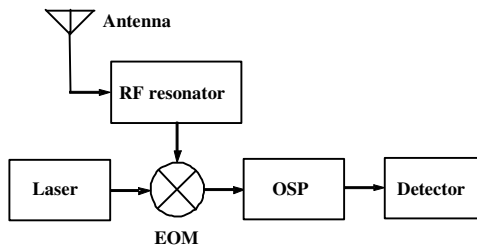


Figure 1. Schematic of microwave-photonic superheterodyne receiver.
EOM - electro-optical modulator, OSP - optical signal processing.

In disk (or ring) resonators an input/output of light from resonator is performed through coupling via evanescent waves between disk's edge (or ring) and prism. The distance between the disk's edge and prism determines the coupling rate between the confined light wave inside the disk and light outside of the resonator. The disk-prism gap should be wavelength-scale where coupling rate between waves in and outside is possible to tune by changing the width of the gap. Thus the gap serves as a some kind of mirror for this type of resonators.

An initial realisation of MW-photonic receiver is relied on EOM based on high Q - factor microdisk resonator [3-7]. Recently the structure of EOM based on high Q-factor Fabry-Perot microresonator has been suggested [15]. Below the brief description of both types of modulators is presented.

2. Electro-optical modulators of microwave-photonic receivers

2.1. Microdisk electro-optical modulators

The first circular optical modulator was demonstrated in 2001 where a LiNbO_3 microdisk cavity was used [3-7]. EOM is based on the z-cut LiNbO_3 disk resonator with optically polished curved side-walls

(Fig.2a). Evanescent prism-coupling is used to couple laser light into and out of a resonant TE-polarised high Q-factor optical whispering-gallery mode (WGM) which exists at the periphery of the disk. A metal electrode structure fed by an RF signal is designed to overlap with the optical field. The resonator's high optical Q-factor increases the effective interaction length of photons with an applied RF microwave field. Combined with a simultaneously resonant microwave structure a highly sensitive receiver at microwave frequencies is achieved [11-14]. Schematic showing the receiver proposed for mm-wave RF detection is presented in Fig.2b [5].

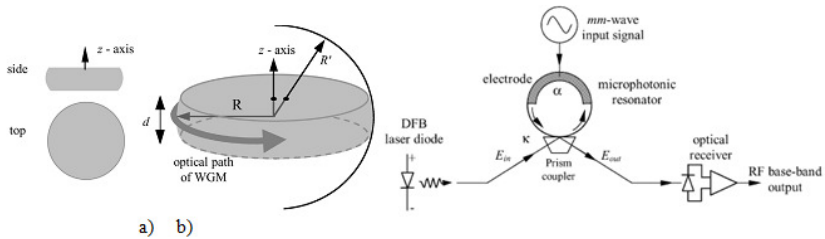


Figure 2. a) Geometry of a microdisk: R is the disk radius, d is the disk thickness and curved side-walls with radius of curvature R' ; b) The receiver proposed for mm-wave RF detection.

An electromagnetic wave received by a RF antenna feeds electrodes of the micro-photonic modulator. The modulator directly converts the RF signal to an optical carrier via the electro-optic effect. The phase-modulated optical signal is internally converted to amplitude modulation through interference with previous optical round trips [5].

The typical radius of LiNbO_3 microdisk is $R = 3.18$ mm and the thickness is $d \leq 1$ mm for operation at 7.67 GHz. The sidewall of the disk is optically polished with a radius of curvature R' , which typically is equal to the radius of the disk. For operation as MW resonator the gold electrodes are located on the top and bottom of the microdisk. RF signal from the microstrip line is applied to the metallic electrodes. For optical part operation a single-mode laser initiates optical whispering gallery modes inside the microdisk (at $\lambda_0 \approx 1550$ nm). The trapezoidal prism is used to input and output an optical radiation from the

microdisk by means of evanescent waves. For this the air gap between the microresonator and the prism should be about the optical wavelength to fit the optimal connection between the prism and the micro-disk. The quality factor Q for the considered microdisk is 4.1×10^6 , and the free space range (FSR) of optical spectrum is 7.67 GHz [5]. The resonant interaction of MW radiation with optical wave in the microresonator takes place when the MW wavelength frequency is a multiple of the FSR of the microresonator. The frequency of the microwave carrier f_{MW} should be an integral multiple m of the optical FSR of resonator such that $f_{FSR} = 1/\tau_{disk} = (2\pi R n_{opt}/c)^{-1}$ where τ_{disk} is the optical round-trip time of the disk, R is the radius of disk and n_{opt} is the refraction index of LiNbO₃ in the corresponding optical range.

The optical whispering gallery mode is confined around the microdisk equator (due to the side-wall curvature) and the modulating E field is confined between the top electrode and the ground around the microdisk so the electro-optical overlap is relatively large. Due to the RF resonance in input electric circuitry the voltage across the microdisk larger than that of the input. The small thickness of the microdisk translates the applied voltage to a large modulating electric field. Such modulators allow efficiently implementing MW-optical conversion and assuring required sensitivity and selectivity of contemporary MW-phonic receivers. However challenges in fabrication of LiNbO₃ microdisks and demands in precision tuning of a micropism's position hinder their wide application. The last is confirmed also by the modest list of publications in this area.

To get rid of above-mentioned complications it was suggested to replace LiNbO₃ microdisk with a high-Q planar Fabry-Perot (F-P) microresonator based on LiNbO₃ operating element [15,16]. Planar configuration of a microresonator has advantages in realization and the usage of a micropism stands no longer.

2.2 Electro-optical modulator based on high Q-factor Fabry-Perot microresonator

It is known, that F-P and microdisk resonators are identical in optical characteristics and mathematical description, and therefore the

choice of resonator's type depends on its feasibility [17]. These two types of microresonators (F-P and microdisk) differ by the round-trip: in the microdisk one the round-trip L_{RT} is equal to its circumference $L_{RT} = 2\pi R$, where R is the microdisk's radius, while in the F-P microresonator $L_{RT} = 2L_{FP}$, where L_{FP} is the distance between the mirrors of the microresonator. The prototype of an EOM based on F-P microresonator is shown in Fig. 3 [15].

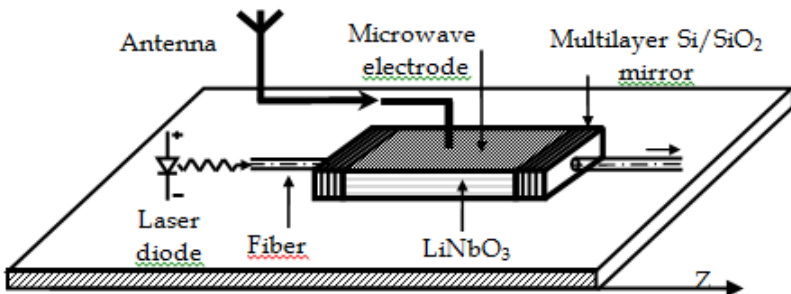


Figure.3. Electro-optical modulator based on F-P microresonator with the multilayer mirrors.

The operating part of the microresonator is based on the z-cut LiNbO_3 wafer. Multilayer mirrors at the transversal facets are alternating quarter-wavelength Si/SiO_2 layers providing high Q-factor of optical microresonator. The lateral facets are covered by layers of metal (top and bottom) for supply of microwave field.

Numerical simulation of optical characteristics of EOM based on F-P microresonator for application in MW-phonic receivers is performed by the method of single expression [18-20]. The microresonator structure consisting of LiNbO_3 layer sandwiched between mirrors consisting of three pairs of quarter-wavelength layers of Si/SiO_2 allows attaining optical spectral characteristics identical to that of the microdisk optical microresonator [15]. The results of numerical simulations permit to assert that the proposed EOM based on F-P microresonator can be offered as an optically identical to microdisk resonator and can be considered as advantageous alternative to the microdisk one.

Numerical simulation of electro-optical characteristics of a modulator based on F-P microresonator for application in MW-phonic

receivers is also performed [16]. To analyse an influence of MW field on propagating optical wave within the F-P microresonator the value of influence degree has been computed as the sum of inputs of decelerated and accelerated velocities of light in the course of a round-trip of the microresonator. The obtained periodicities for the frequencies of zero and maximal interaction of waves are in a good agreement with the corresponding data for microdisk resonators [3-6]. The effect of increase of operating frequencies of F-P electro-optical modulator with the decrease of the length of the top metallic electrode is also obtained. The steps of operation frequencies are also in a good agreement with the data for microdisk resonators [3-6].

3. CONCLUSION

The idea to transfer an intermediate frequency of MW superhetrodyne receiver in optical range is considered now as productive and prospective. By this operation it is not only possible completely to get rid of parasitic radiation in RF range, but also attain the high sensitivity, selectivity and bandwidth, while having immunity to the external electromagnetic stray radiation, small size, weight and power consumption [3-7,11-14].

Realisation of this idea relies strongly on the construction of electro-optical modulator providing effective interaction between MW electrical signals with optical wave. The key element of MW-photonics receiver up to now is LiNbO₃ microdisk (or ring) resonator, which needs precise microdisk preparation and optical wavelength-scale positioning of micropism for optimal light input/output from the microresonator.

An application of high Q-factor F-P microresonator instead of microdisk has been suggested recently [15,16], that is prospective due to elimination of micropism from construction and application of planar structure instead of circular one.

4. REFERENCES

- [1] U.L. Rohde, J.C. Whitaker, Communications Receivers: DSP, Software Radios, and Design (McGraw-Hill, New York, 2001).

- [2] R.A. Sowadski, US Patent 5179728 -Spurious product and local oscillator radiation suppression system for use in superheterodyne radio receivers. No. 767736, filled on 1991.
- [3] D. A Cohen, M. Hossein-Zadeh, A. F. J. Levi, Microphotonic modulator for microwave receiver, *Electron. Lett.* **37** (2001) 300.
- [4] D. A. Cohen, A.F.J. Levi, Microphotonic components for a mm-wave receiver, *Solid-State Electronics* **45** (2001) 495.
- [5] D.A. Cohen, M. Hossein-Zadeh, A.F.J. Levi, High-Q microphotonic electro-optic modulator, *Solids-State Electronics* **45** (2001) 1577.
- [6] D. A. Cohen, A. F. J. Levi, Microphotonic millimetre-wave receiver architecture, *Electronic Lett.*, **37** (2001) 37.
- [7] V. S. Ilchenko, A. A. Savchenkov, A. B. Matsko, L. Maleki, Submicrowatt photonic microwave receiver, *IEEE Photonics Technol. Lett.* **14** (2002) 1602.
- [8] R.A. Minasian, Photonic signal processing of microwave signals, *IEEE Transactions on Microwave Theory and Techniques*, Special Issue on Microwave Photonics, Invited Paper **54** (2006) 832.
- [9] L.Ng. Binh, *Photonic Signal Processing: Techniques and Applications* (Taylor&Francis Group, 2008).
- [10] *Microwave Photonics: Devices and Applications*, Edit. S. Iezekiel (John Wiley & Sons, Ltd., 2009).
- [11] V. S. Ilchenko, A. B. Matsko, L. Solomantine, A. A. Savchenkov, D. Seidel, L. Maleki, Ka-Band all- resonant photonic microwave receiver, *IEEE Photonics Technol. Lett.* **20** (2008) 1600.
- [12] M. Hossein-Zadeh, and A. F. J. Levi, Ring resonator-based photonic microwave receiver modulator with picowatt sensitivity, *IET Optoelectronics*, **5** (2011) 36.
- [13] M. Hossein-Zadeh, and A. F. J. Levi, Self-homodyne photonic microwave receiver architecture based on optical modulation and filtering, *Microwave and Optical Technol. Lett.* **50** (2008) 345.
- [14] A. A. Savchenkov, A. B. Matsko, W. Liang, V. S. Ilchenko, D. Seidel, L. Maleki, Single-side band electro-opticalmodulator and tunable microwave photonic receiver, *IEEE Trans. on Microwave Theory and Tech.*, **58** (2010) 3167.
- [15] H.V. Baghdasaryan, T.M. Knyazyan, A.K. Aharonyan, M. Marciniak, Numerical Modelling of Fabry-Perot Microresonator Based Electro-Optical Modulator for Microwave- Photonic Receiver, *ICTON 2011*,

- Conference Proceedings, Stockholm, Sweden, June 26- 30 (2011), We.C5.7, 4pages.
- [16] H.V. Baghdasaryan, T.M. Knyazyan, A.K. Aharonyan, M. Marciniak, Numerical Analysis of Operation of Metallic Electrodes in Fabry-Perot Electro-Optical Modulator of Microwave-Photonic Receiver, ICTON 2012, Conference Proceedings, Coventry, UK, July 2-5 (2012), Tu.D4.4, 4pages.
 - [17] I.-L. Gheorma, R.M. Osgood, Jr., Fundamental limitations of optical resonator based high-speed EO Modulators, IEEE Photonics Technology Letters, **14** (2002) 795.
 - [18] H.V.Baghdasaryan, “Method of backward calculation”, in the book: Photonic Devices for Telecommunications: how to model and measure/ Edit. G.Guekos, Springer (1999) 56-65.
 - [19] H.V. Baghdasaryan, T.M. Knyazyan, Problem of Plane EM Wave Self-action in Multilayer Structure: an Exact Solution, Optical and Quantum Electronics **31** (1999) 1059.
 - [20] H.V. Baghdasaryan, T.M. Knyazyan, T.H. Baghdasaryan, B. Witzigmann, and F. Roemer, Absorption loss influence on optical characteristics of multilayer distributed Bragg reflector: wavelength-scale analysis by the method of single expression, Opto–Electronics Review **18** (2010) 438.

S-Band 800 Watt Gallium Nitride Based Pallet Amplifier For Air Traffic Control Radars

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Abstract — This article presents 800 Watt Gallium Nitride (GaN) based power amplifier operating in 2.7-2.9GHz frequency band with Class AB bias. The RF performance of the amplifier under pulse conditions with 300us pulse width and 10% duty cycle is characterized. The amplifier is designed for S-band Air Traffic Control Radar applications and can operate both under short pulse/low duty cycle and medium pulse/duty cycle conditions. The amplifier is based on 400 Watt, internally pre-matched, GaN high electron mobility transistor (HEMT).

Index Terms—Power amplifier, GaN, power transistors, power gain.

1. Introduction

Demand for high power S-band amplifiers for radar applications, such as Air Traffic Control, is strong and as always, high power output and high efficiency are key requirements for these amplifiers. GaN transistors provide excellent power density and efficiency and as such are excellent choice for S-band high power applications. High efficient power amplifiers (PA) make the system level thermal aspects more manageable, reduce operational cost, and provide an opportunity to reduce the overall transmitter size by reducing the cooling element size.

A high power S-band Class AB power amplifier (PA) is designed, implemented and characterized based on Gallium Nitride (GaN) high electron-mobility transistor (HEMT) operating over the 2.7-2.9GHz frequency range. A power gain of 11.0dB and drain efficiency of 53.9% was achieved at 800 Watt power output level across the operating 2.7-2.9GHz band. Conventional 2-way high power Wilkinson divider/combiner [1] is developed that can handle amplifier operation even when one of the transistors fails during the operation. Integra Technologies IGN2729M400 [2] transistor is used for amplifier design. Figure 1 below shows a picture of the amplifier. Rogers RO4350 circuit board

material with a dielectric constant of 3.48 is utilized in order to minimize the size and loss from the transmission line matching networks while providing sufficient power handling capability for 800 Watt operation.

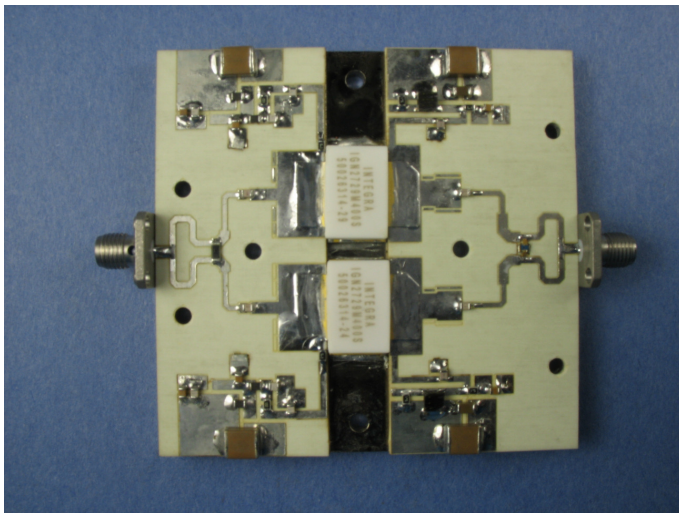


Figure 1. 2.7-2.9GHz, 800 Watt pallet photo. Dimensions: 72.1mm x 69.6mm

To the best of our knowledge this is the highest power pallet amplifier based on two single ended transistors operating in the specified frequency range.

2. Pallet Amplifier Design

2.1. IGN2729M400 transistor

IGN2729M400 is an internally pre-matched, GaN HEMT. It is designed for S-band radar applications operating over 2.7-2.9GHz instantaneous frequency band. Under 300us pulse width and 10% duty cycle conditions it supplies a minimum of 400 Watts of peak output power with 10.7dB of minimum gain. Transistor operates with 50V drain bias with 100mA bias current. It is specified for Class AB

operation. Rated minimum drain efficiency is 53%, which is calculated as

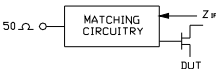
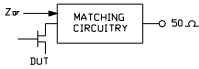
$$N_d = P_{out} / (V_D \times I_{Dpk})$$

where N_d is drain efficiency, P_{out} is peak output power, V_D is drain bias voltage and I_{Dpk} is peak drain current.

Thermal impedance is rated at 0.19°C/W at $P_{out} = 476\text{W}$ and $T_{case} = 57^\circ\text{C}$.

Device input and output impedances are given in the table below:

Table 1. IGN2729M400 IMPEDANCE CHARACTERISTICS

Frequency (GHz)	$Z_{IF} (\Omega)$	$Z_{OF} (\Omega)$
2.70	$3.1 - j1.6$	$3.3 - j1.0$
2.80	$3.1 - j1.0$	$3.0 - j0.3$
2.90	$3.3 - j0.3$	$3.2 + j0.6$
Impedance Definition		

2.2. 2-way Wilkinson Power Divider and Combiner

Wilkinson power divider and combiner have been designed, individually fabricated and tested. For each network, divider and combiner, isolation resistor was selected and evaluated. Both resistors were selected with sufficient power rating to handle the worst case scenario operation and provide safe operation in case one of the power transistors were to fail during filed operation.

Table 2 below summarizes measured and simulated S-parameters for both the power divider and combiner, including port to port isolation:

Table 2. S-parameters for 2-Way Power Divider

	F (GHz)	S11(dB)	S12(dB)	S13(dB)	S23(dB)
Simulation Results	2.70	-27.78	-3.11	-3.11	-25.61
	2.80	-32.76	-3.11	-3.11	-29.49
	2.90	-39.49	-3.11	-3.11	-36.66
Power Divider	2.70	-19.63	-3.13	-3.14	-19.35
	2.80	-20.92	-3.13	-3.14	-20.77
	2.90	-22.52	-3.16	-3.16	-22.30
Power Combiner	2.70	-20.30	-3.14	-3.18	-16.88
	2.80	-22.10	-3.15	-3.18	-17.96
	2.90	-24.35	-3.18	-3.18	-19.16

2.3. 800 Watt Amplifier

Two IGN2729M400 transistors were combined in parallel to achieve 800 Watt operating power across the 2.7-2.9GHz operating frequency range. Amplifier is specified with 68 Watt input drive level and minimum gain of 10.70dB [3]. Recorded worst case efficiency was 53.9% and power gain was 11.02dB. Pulse droop, which was measured from 30us to 270us interval was -0.39dB recorded at 2.8GHz frequency. Recorded worst case Return Loss was 16.0dB across the band. Overall, amplifier demonstrates excellent stability against the load mismatch and it is rugged to 3:1VSWR. Amplifier's power transfer curves, Pin-Pout characteristics, are given in the figure below.

As can be seen from the graphs, at 68 Watt input drive level amplifier produces more than 850 Watt output power at 2.7 and 2.9GHz frequencies whereas it reaches more than 1050 Watt at 2.8GHz. This behavior can be explained by transistor characteristics.

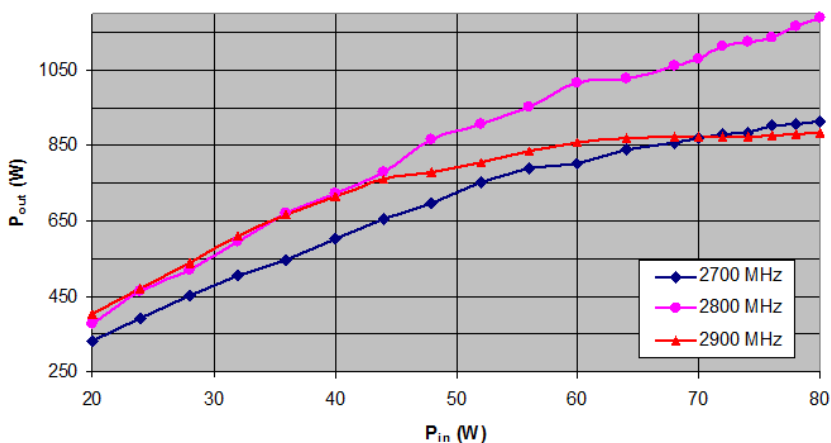


Fig.2: Measured Power Output vs. Power Input Characteristics

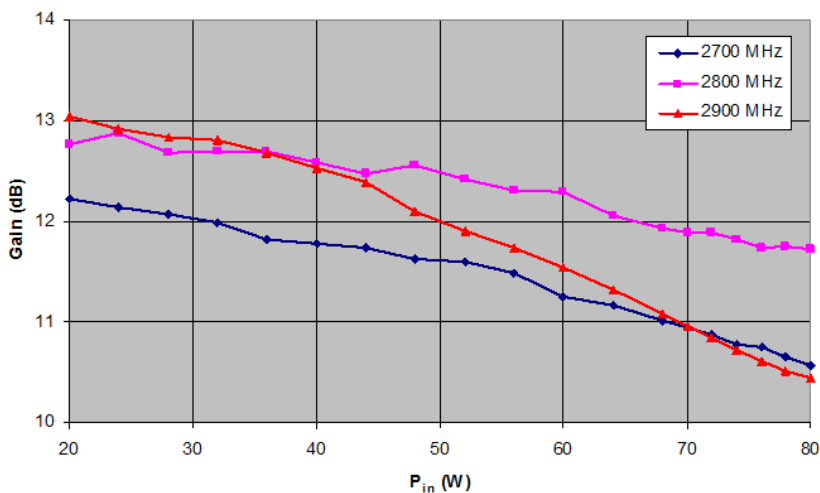


Fig.3: Power Gain vs. Power Input

Figure 3 above shows amplifier gain performance versus input drive level and as expected, for the reason explained above, gain is higher at 2.8GHz (at more than 50 Watt drive levels). Figure 4 below shows power gain versus frequency at P_{in}=68 W. As can be seen from the graph, gain flatness is better than 1dB in 2.7-2.9GHz range.

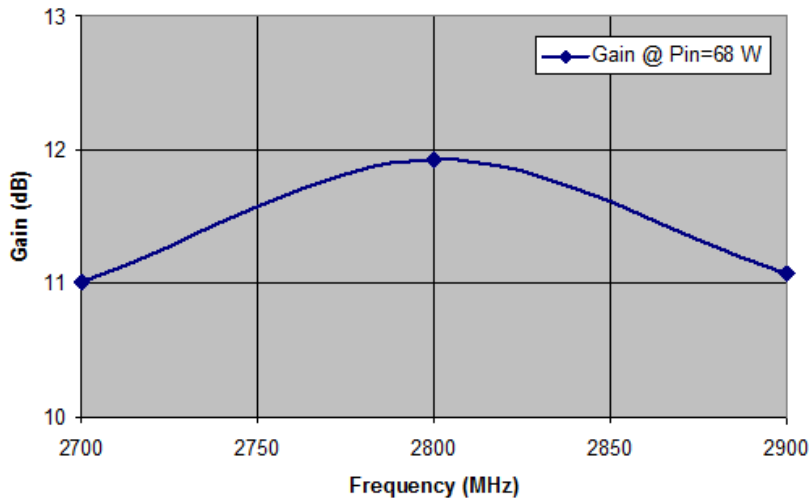


Fig.4: Power Gain vs. Operating Frequency at Pin=68W

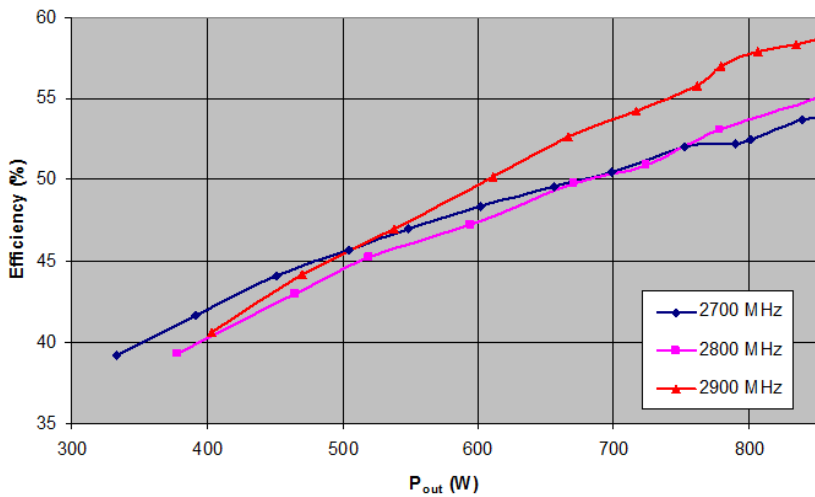


Fig.5: Drain Efficiency vs. Power Output

Figure 5 above shows amplifier drain efficiency versus Power Output and as expected it reaches maximum efficiency at saturated

power levels. About 53.5% minimum drain efficiency was recorded across the entire frequency band at 800 Watt power output level.

Conclusions

S-band GaN based 800 Watt pallet amplifier was designed and implemented. Minimum of 800 Watt operating power was obtained with 11dB power gain and 53% drain efficiency. Recorded worst case pulse droop was -0.39dB at 2.8GHz at power output level of 1059 Watt. At 2.8GHz and 2.9GHz frequencies recorded drain efficiency was 59.0% and 59.5%, respectively. Amplifier can operate with medium pulse/duty cycle as well as short pulse/low duty cycle pulse conditions and demonstrates excellent pulse fidelity under both pulse conditions.

3. References

- [1] Robert E. Collin, Foundations for Microwave Engineering, 2nd Edition, John Wiley & Sons
- [2] Datasheet IGN2729M400 www.integratech.com
- [3] Datasheet IGNP2729M800 www.integratech.com

Investigation of Carbon-PEEK piezoelectric properties by using a near-field scanning microwave microprobe

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We have used a near-field scanning microwave microprobe (NSMM) to investigate the electric properties of the Carbone Poly-Ether-Ether-Ketone (Carbone-PEEK) bulk sample. Sensitive characterization of Carbone-PEEK and high resolution mapping of Carbone-PEEK lines were realized by measuring the change of microwave reflection coefficient (S_{11}) at about 4 GHz operating frequency. The conducting channel in a Carbone-PEEK structure under external electric field was visualized by using a NSMM. The increase in conductivity due to the piezoelectric effect was caused by generation and accumulation of excess holes in the Carbone-PEEK. As the electric field intensity increased, the conductivity of Carbon increased due to the generation of carriers through the sample. The current-voltage characteristics correlate well with the NSMM measurements.

The visualization of conductivity-voltage behavior provides a detailed configuration of the carrier injection, the mobile charge distribution in the Carbone-PEEK structure. Thus, the present combination of NSMM images with the current-voltage characteristics is very effective for evaluating the carrier distribution in the Carbone-PEEK. The NSMM provides a unique approach to investigate the piezoelectric transport mechanism along the conducting channel in Carbone-PEEK structure.

Introduction

The properties of carbon fiber reinforced composite materials are

well recognized [1-3]. The applications for carbon composite materials are possible as promising materials for mechanical and electrical applications. One property of considerable interest in the case of micro-reinforced composites containing conductive fibers like carbon is the electrical conductivity. These composites with their superior mechanical properties may provide electrically conductive medium.

Although considerable efforts have been made at investigating the mechanical properties of these high performance composites, there is a real lack of information as far as physical properties, in particular electrical conductivity and electrical anisotropic properties are concerned. The study of conductivity under the external electromagnetic current is important for understanding current transport, charge accumulation, charging, and trapping in carbon composite materials. Therefore, measurement of small variations in electrical conductivity is important.

PEEK is an abbreviation for Poly-Ether-Ether-Ketone, a high performance engineering thermoplastic. PEEK grades offer chemical and water resistance similar to PPS (Poly-Phenylene-Sulfide) but can operate at higher temperatures. PEEK can be used continuously to 250 °C and in hot water or steam without permanent loss in physical properties. For hostile environments, PEEK is a high strength alternative to fluoropolymers. PEEK carries a zero flammability rating and exhibits very low smoke and toxic gas emission when exposed to flame. And unfilled PEEK is approved by the FDA for food contact applications (since 1998).

We demonstrate a direct probing of the conductivity distribution in a carbon fiber reinforced composite material by using a near-field scanning microwave microscope (NSMM). To show the application of NSMM for the direct visualization of the conductivity image of the sample surface, we directly visualized the conductivity of the surface of carbon fiber reinforced composite materials. This NSMM approach is based on nondestructive probing of a local electromagnetic near-field interaction between the probe tip and the composite materials, and can be used in exploring materials at the nano-scale.

Experiment

Carbon composite is based on continuous carbon fibers embedded in a PEEK matrix with vertical and horizontal arrangement as shown in Fig. 1 (a). The carbon fibers with each prepared layer are about $7\text{ }\mu\text{m}$ in diameter and typically occupy about 60 % of the volume. Figure 1 (b) shows the scanning area on the Carbon channel with about $200\text{ }\mu\text{m}$ width.

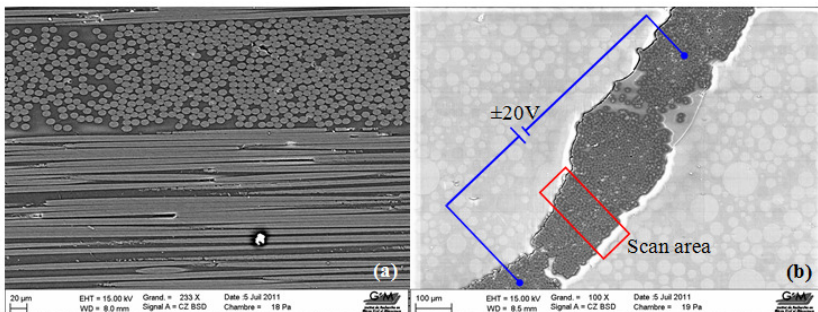


Fig. 1. (a) Vertical and horizontal arrangement of Carbon-PEEK structure. (b) SEM image of carbon fibers reinforced in PEEK matrix with localized scanning area and applied voltage points.

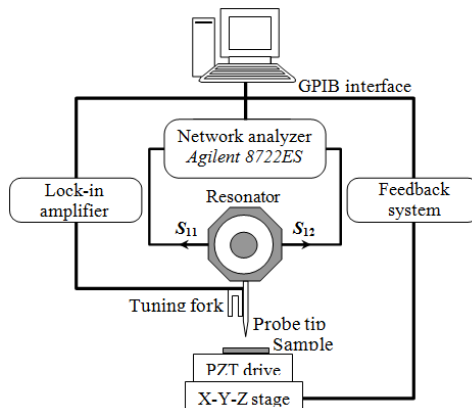


Fig. 2. Schematic of experimental setup for NSMM

We use a NSMM) to investigate Carbon-PEEK conducting properties for different fiber arrangements and under various external electric

fields. The basic experimental setup of our NSMM consists of a metal probe tip coupled to a dielectric resonator as shown in Fig. 2. The probe tip-sample distance was fixed by using a tuning fork feedback control system at about 20 nm above the surface and the operating frequency was $f = 4.1$ GHz. A tungsten probe tip with a diameter of 50 μm was glued onto one of the prongs of a tuning fork, which attached to the resonator [4]. The microwave reflection coefficient S_{11} is determined by the intrinsic impedance of the sample. The resonance frequency of a given mode was TE_{01} and the unloaded Q -factor was 24,000. A network analyzer (Agilent 8753ES) was used in measuring the reflection coefficient S_{11} .

In order to study the anisotropic transport properties of Carbon-PEEK structure, we measured the I - V by 4-probe method and compared with the near-field microwave characteristics.

The surface morphology of the Carbon-PEEK structure was characterized by a scanning electron microscope (SEM).

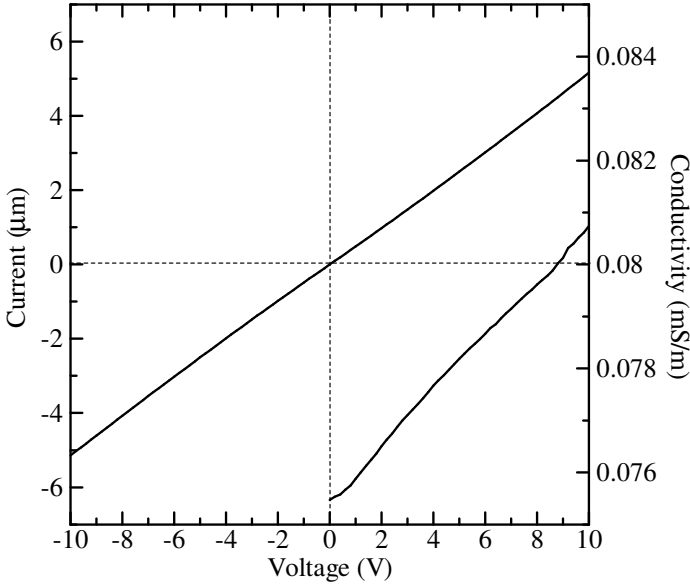


Fig. 3. I - V characteristics obtained from Carbon-PEEK structure.

Results and discussion

To characterize the conductivity of metals we measured the reflection coefficient S_{11} at the resonance frequency. The resonant frequency is sensitive to the near field interaction of the probe tip with the sample. The cavity and resonator store electromagnetic energy with the resonant frequency depending on the conductivity. By measuring the reflection coefficient S_{11} of the microwaves input into the resonant cavity it is possible to determine the conductivity of the metal samples. A formula showing how the reflection coefficient S_{11} depends on conductivity of the metal sheets can be derived by using standard transmission line theory and is given by assuming impedance matching between the probe tip and the microwave source [5]

$$S_{11} = 20 \log \left| \frac{|Z_{in}| - Z_0}{|Z_{in}| + Z_0} \right|, \quad (1)$$

where S_{11} is the microwave reflection coefficient, Z_0 is the impedance of probe ($Z_0 = 50 \Omega$), and Z_{in} is the complex impedance of the Carbon-PEEK/Al system and defined as

$$Z_{in} = Z_c \frac{Z_s + jZ_c \tan(k_c t_c)}{Z_c + jZ_s \tan(k_c t_c)} \cong \frac{2}{Z_a k_a t_p \delta_c^2 \sigma_c^2} - j \frac{1}{\sigma_c t_c}. \quad (2)$$

Here, Z_c , k_c , t_c ($t_c = 10$ mm), σ_c ($\sigma_c = 400$ S/m), and δ_c ($\delta_c = 15$ cm) is the characteristic impedance, wave number, thickness, conductivity, and the skin depth of Carbon layer, respectively. Z_s is the complex impedance of matched PEEK/Al system, Z_a , and k_a , is the characteristic impedance and wave number of air ($Z_p = 233 \Omega$ and $k_a = 86 \text{ m}^{-1}$), and t_p is the thickness for PEEK layer ($t_p = 20$ mm).

Figure 3 shows the I - V characteristics of Carbon-PEEK channel. The trans-channel conductivity linearly increased as applied voltage increased due to the injection of the additional carriers in the Carbon as shown in Fig. 3.

In order to further understand electrostatic phenomena at the Carbon-PEEK structure we take advantage of the noncontact and non-

destructive evaluation capabilities of a NSMM. The S_{11} behavior depends on applied voltages. The behavior is due to injected carrier trapping in the Carbon channel.

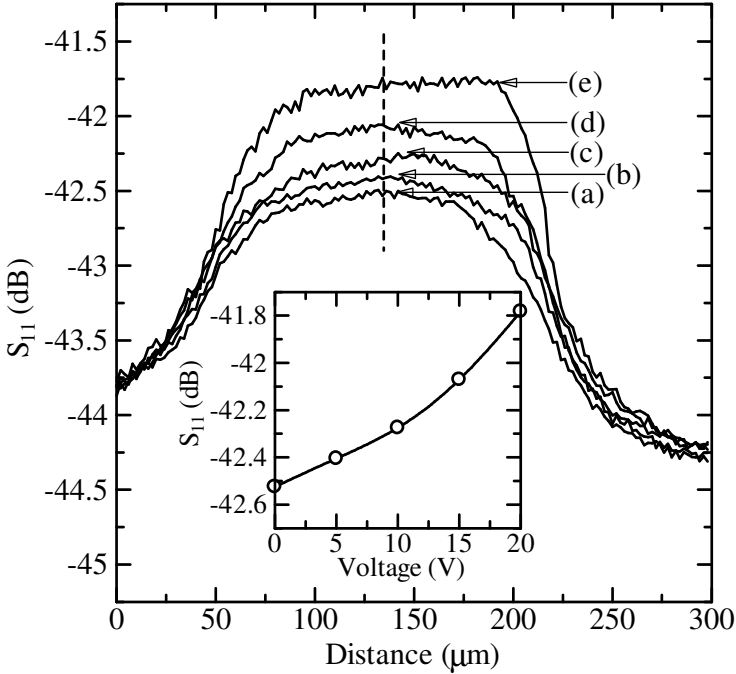


Fig. 4. NSMM line scans obtained from Carbon-channel structure for applied external voltage of (a) 0 V, (b) 5 V, (c) 10 V, (d) 15 V, and (e) 20 V. The inset shows microwave reflection coefficient S_{11} vs. applied voltage at position along dashed line.

Figure 4 shows the NSMM line scans along the channel of a Carbon-PEEK (marked by red rectangle in Dig. 1 (a)) at various applied voltages ranged 0 - 20 V. The change of applied voltages affected the I - V and microwave reflection coefficient characteristics of the structure due to the changes of the channel's conductivity. The microwave reflection coefficient S_{11} measurements for a Carbon-PEEK structure also reveal the changes of conductivity distribution in the conducting channel. That is, the electric conductivity distribution probed by

NSMM depends on applied voltages and well confirmed by I - V characteristics.

Conclusions

We demonstrated the measurement of piezoelectric properties of Carbon-PEEK sample by using a NSMM. As the conductivity of sample increased, the intensity of the reflection coefficient S_{11} increased due to the increment of conducting properties of the Carbon channel. The piezoelectric properties of Carbon-PEEK was estimated by measuring the microwave reflection coefficient S_{11} and compared with the I - V characteristics. We developed a calculation model how the reflection coefficient S_{11} depended on the conductivity of the Carbon-PEEK sample by using a standard transmission line theory. These results clearly show the sensitivity and usefulness of this device for investigating piezoelectric properties of composite materials.

Acknowledgements

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References

- [1] A. Lagarkov, S. Matytsin, K. Rozanov, A. Sarychev, J. Appl. Phys. **84** (1998) 3806.
- [2] M. Y. Koledintseva, J. Drewniak, R. Dubro, Prog. Electromagn. Res. B **15** (2009) 197.
- [3] L. Liu, L. B. Kong, W. Y. Yin, Y. Chen, S. Matitsine, Microwave Dielectric Properties of Carbon Nanotube Composites, Carbon Nanotubes, Jose Mauricio Marulanda (Ed.), ISBN: 978-953-307-054-4 (2010).
- [4] J. Kim, A. Babajanyan, A. Hovsepyan, K. Lee, B. Friedman, Rev. Sci. Instrum. **79** (2008) 086107.

[5] D. Pozar, Microwave Engineering (Anderson-Wesley, New York, 1990).

Design of miniature high power waveguide ceramic filter with improving characteristics

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By using ceramic materials with dielectric constant 37, the miniature waveguide ceramic band-pass filters were implemented. The proposed waveguide ceramic band-pass filters were developed based on the mode matching method and the simulation process. The frequency characteristics were improved compared with the commercial dielectric ceramic filters which have been widely used in market. In order to overcome the main disadvantage of high power application, we used the SMA connectors for in/output ports. The results of power tests were obtained about 30W. For the future applications, miniature waveguide ceramic band-pass filters would be replaced by the metal cavity filters and widely used as the parts of base station for 4th generation wireless communication.

1. Introduction

The demand for wireless communication systems has been increased with popularization of wireless mobile communication services rapidly. The excellent frequency selectivity and the need for smaller and lighter communication system have been emphasized fully. Especially, to meet the characteristics and the minimization of volume, size and weight, the research has been progressed continuously. Nowadays, mainly the band-pass filters have been used by the metal cavity filters for the base station for wireless communication [1,2]. Furthermore, the demands for the dielectric ceramic filters have been increased because of the limitation of metal cavity, such as the volume, cost, size, and weight. Commercial dielectric TEM mode ceramic filters which have

been widely used in commercial market cannot withstand the high power applications as well as the low quality Q factor. Note that, the maximum power of commercial dielectric TEM mode ceramic filter was recorded less than about 1W.

In this paper, the miniature waveguide ceramic band-pass filters were implemented by using a ceramic material with dielectric constant $\epsilon_r = 37$. The size and weight of proposed waveguide ceramic band-pass filters were decreased smaller than that of metal cavity filters significantly. In order to increase the quality factor Q value we used the waveguide TE mode. To solve the power problem of ceramic filters, we used the SMA connector at *in/output* ports of filter. The designed miniature waveguide ceramic filters were obtained about 30W.

2. Design and analysis

The size of waveguide ceramic band-pass filters which was fully filled with dielectric constant is proportional to $1/\sqrt{\epsilon_r}$ due to focusing of the most part of the electromagnetic field in the dielectric material [3]. Also, because of use the TE mode, it has a merit that can increase quality factor Q value than commercial dielectric ceramic filter which used TEM mode. We analyzed H -plane discontinuity in rectangular waveguide which is filled with material with the large dielectric constant proposed by Konish and applied to mode matching techniques [4]. The S parameter S_{11} and phase value is induced by mode matching techniques as below

$$|S_{11j-1,j}| = \frac{k_{j-1,j}^2 - 1}{k_{j-1,j+1}^2 + 1} \quad (1)$$

$$\Phi = \angle(S_{11}) \quad (2)$$

The size of waveguide ceramic band-pass filter was determined by the HFSS simulation software (Ansoft) and optimized through the design process as shown in Fig.1. The central frequency was 2.3 GHz for long term evaluation (LTE) and Wibro-band frequencies. Specifications of waveguide ceramic filter were shown in Table 1. For the high power application, it made a hole at input and output ports. The diameter, depth, and location of hole depended on the S parameter and

phase value at the ports. We could obtain the resonance frequency after impedance matching through the optimization of each elements and hole.

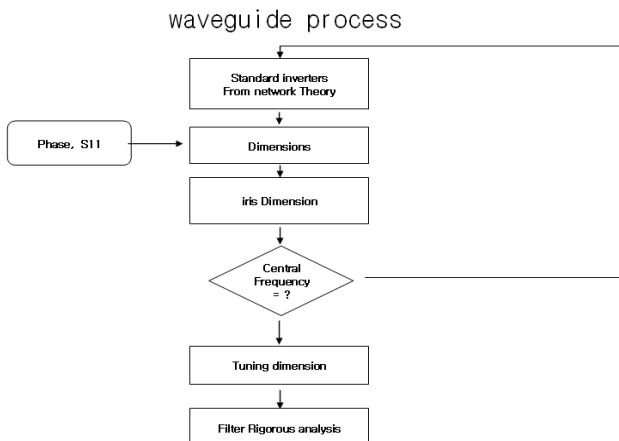


Figure 1. The design process of waveguide ceramic filter

Table 1. Specifications of waveguide ceramic filter

Item	Specifications
Center frequency [GHz]	2.3
Bandwidth frequency [MHz]	100
Return loss [dB]	-15
Insertion loss [dB]	-1
Power [W]	30

3. Results and Discussion

The optimized sizes of waveguide ceramic filter of width, height, and length were about $18.0\text{ mm} \times 9.0\text{ mm} \times 67.5\text{ mm}$, respectively, with 5 elements as shown in Fig. 2. The Frequency characteristics of the waveguide ceramic filter were shown in Fig. 3. The maximum out-

put power test level was recorded about 45 dBm (30W) by error vector magnitude (EVM) and spurious emission tests as shown in Table 2.

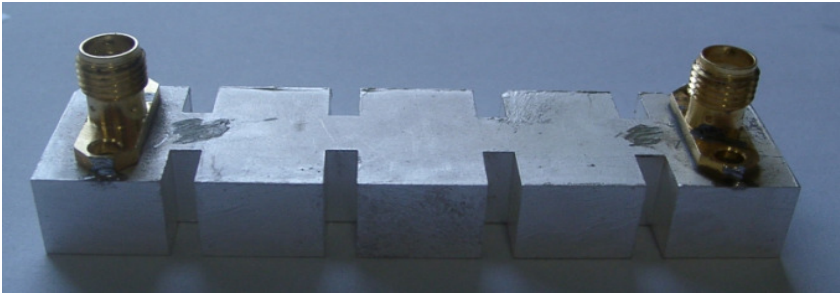


Figure 2. The proposed waveguide ceramic band-pass filter

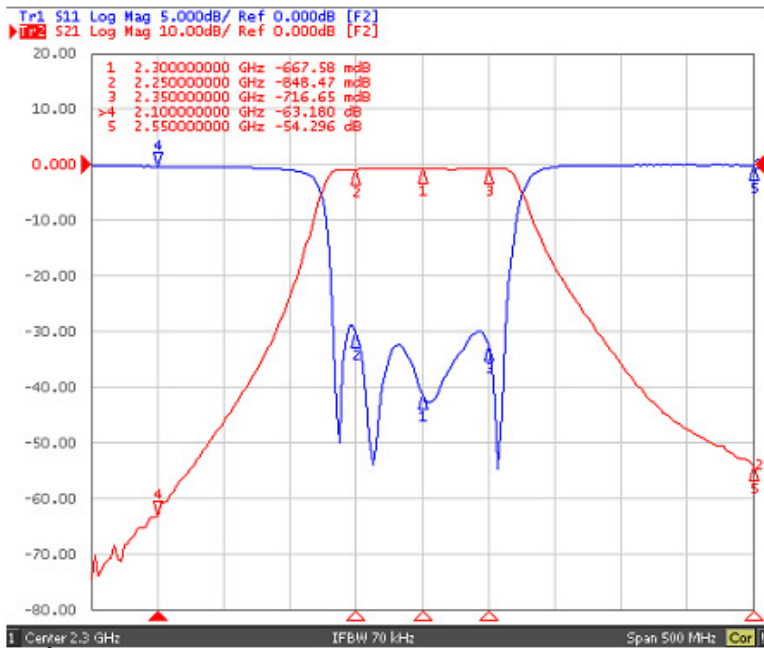


Figure 3. Frequency characteristics of the waveguide ceramic filter.

Table 2. Results of high power test.

Output level	EVM ¹⁾ test	Spurious emission values ²⁾
20dBm	1.72%	-53.02dB @4.77MHz ³⁾
25dBm	1.55%	-53.88dB @4.77MHz
30dBm	1.47%	-51.64dB @4.77MHz
35dBm	1.49%	-48.15dB @4.77MHz
40dBm	1.47%	-45.45dB @4.77MHz
45dBm	1.47%	-40.81dB @4.77MHz

Note: ¹⁾ EVM of common metal cavity filter – 1.47% @45dBm.

²⁾ Spurious emission values of common metal cavity filter -53.92dB @45dBm.

³⁾ Point of far 4.77MHz from the center frequency of 1st FA.

4. Conclusion

Using dielectric ceramic materials with dielectric constant 37, we could make the miniature waveguide ceramic band-pass filters. To solve the problem of impedance matching which is happened from surface discontinuity between air and dielectric substance, we found the S_{11} parameter and phase value through the mode matching techniques. The size of waveguide ceramic filter was optimized with HFSS simulation tool. To overcome the high power of 30W, the *in/output* ports of filter was used the SMA connector directly. This study was implemented in part of system for LTE and Wibro band. Furthermore, to meet the specification of more than 5GHz in the future, it was emphasized the design process.

Acknowledgements

This work was supported by Parts and Materials International Collaborative R&D Program (funded by the Ministry of Knowledge Economy (MKE, Korea(N0000481)).

5. References

- [1] J. Uher, J. Bornemann, U. Rosenberg, Waveguide components for antenna feed systems: Theory and CAD (Artech House, MA, 1993).

- [2] G. L. Matthaei, L. Young, E. M. T. Jones, Microwave filters, impedance-matching networks, and coupling structures (Artech House, MA, 1980).
- [3] D. M. Pozar, Microwave Engineering 3rd ed. (Wiley & Sons, New York, 2005).
- [4] K. M. Luk, K.W. Leung, (eds.) Dielectric resonator antennas (Research Studies Press, Hertfordshire, 2003).

Reduction of the Clutter in Non-Coherent LFM CW Radars

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Abstract

Algorithm of clutter reduction in LFM CW radars is presented. Similarities in sequential measurements are formalized via introduced identification function. Ensemble averaging allows significantly simplify moving target identification.

1.Introduction

Recently, portable CW radars find wide application in military and civil areas taking into account a number of requirements, such as, small dimensions, competitive price and so on. Among the main areas of application for the mentioned radars are surveillance security systems, vehicle collision avoidance systems and etc. [1], [2]. Recently developed applications imposed specific requirements on radars parameters; hence, the parameters of such systems should be improved. The mentioned systems are considered as “intellectual” radio devices and need serious software support for effective operation. For such purpose, development of efficient algorithms is vital. As it is known, one of the most important parameters of radar systems is its dynamic range. The total capability of the system can be improved, by dynamic range boosting, although it is not always sufficient. Sometimes indirect improvement of system parameters can be more efficient. These include: suppression of the reflected signal from the local objects (clutter), which will have the same result as the dynamic range boosting. This problem, in traditional coherent-pulse radar systems, is solved by periodical comparison of received signals of more than two neighboring pulses, and detection of their spectral differences [6]. In CW radar the situation substantially differs, since in this case the spectra are not fully identical and there is only likeness between them. Thus, the problem is summarized in detecting similarities of signal

spectra received from neighboring sequential periods.

2. System and measurements

One of the most effective ways of distance measurement with CW radars is using of a linear frequency modulation (LFM). The system transmits LFM signal with the periodically rising and falling frequency slopes. Such measurements during slopes will allow us to identify the distance and the radial speed of the target. The signal reflected from the target contains the frequency shift, simultaneously due to the signal delay and Doppler Effect. The received signal is down converting and the result is the pulsing signal shown in Fig. 1.

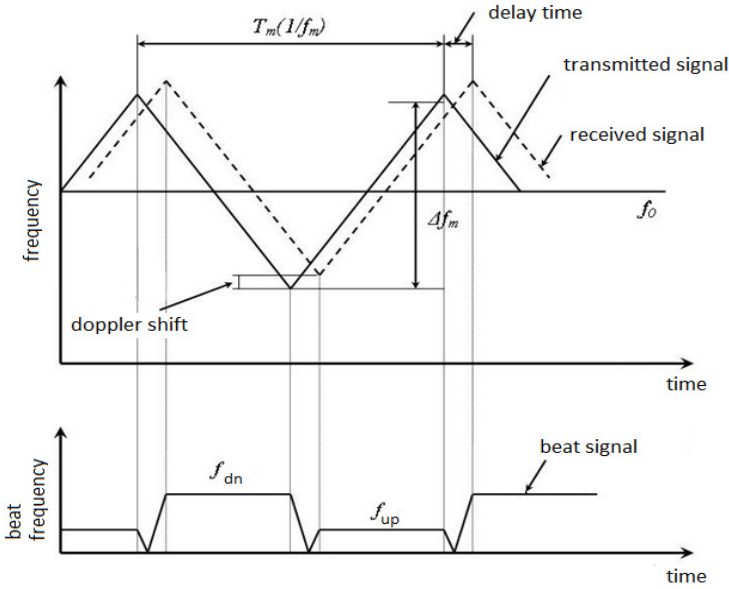


Fig. 1

Achieved signal is the sequence of the signals with f_{up} and f_{down} frequencies corresponding to the LFM up- and down-periods. Depending on the target movement direction and the period of the transmitted signal, the possible values of resulting frequency are:

$$f_1 = |f_{delay} + f_{doppler}| \quad (1)$$

$$f_2 = |f_{delay} - f_{doppler}| \quad (2)$$

Thus it is obvious, that the detected frequencies of the signal reflected from local objects will be same for the rising and falling slopes. Meanwhile for the case of moving object they will differ by the value of twice of the Doppler frequency [3], [5]. We have carried out a huge amount of spectral measurements including clutter, human and vehicle targets.

The spectra of down converted signals for up- and down-periods are shown in Fig. 2 (clutter), meanwhile the same for the moving targets is shown in Fig. 3. Although, the spectra of the signal reflected from clutter are not strict identical, but there is qualitative resemblance between them.

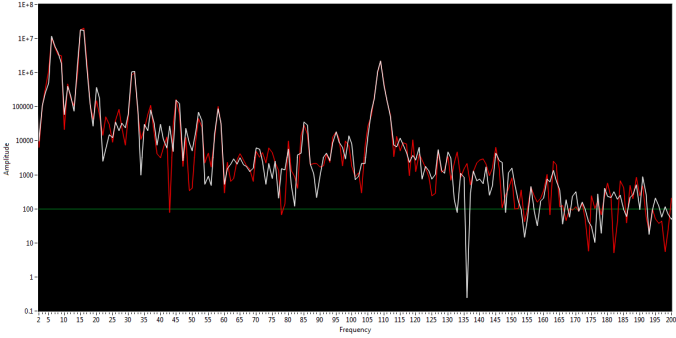


Fig.2 Spectra of clutter

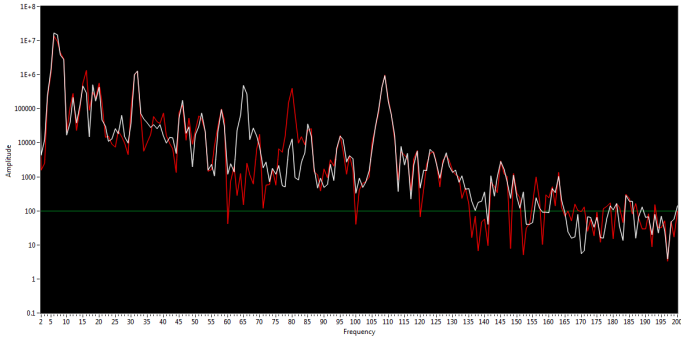


Fig.3 Spectra of clutter and moving targets

Taking into account aforementioned considerations, we can conclude that for non-coherent radars it is also possible to suppress the clutter, which will improve the potential of the system if the respective processing algorithm will be applied. The obtained results are presented and discussed in [4]. The essence of the tested method is in comparison and suppression of the spectral peaks with the same frequencies. Actually, the method is simple enough and realizable, but there is not enough resemblance between spectral peaks, caused by the system non-coherence and noise existence, that's why it is hard to get desirable level of clutter suppression. Based on considerations, that spectral differences are caused by the system non-coherence, it can be concluded that they will have random character. The method of ensemble averaging of the random processes can be successfully used in this case as well. We already have done some preliminary estimation on that topic, and achieved results prove the reliability of mentioned considerations. In order of proper description, we introduce *identification function* as the spectrum of point-by-point ratio of two sequentially measured periods. Such identification function of spectral components vs. frequency is shown in Fig. 4.

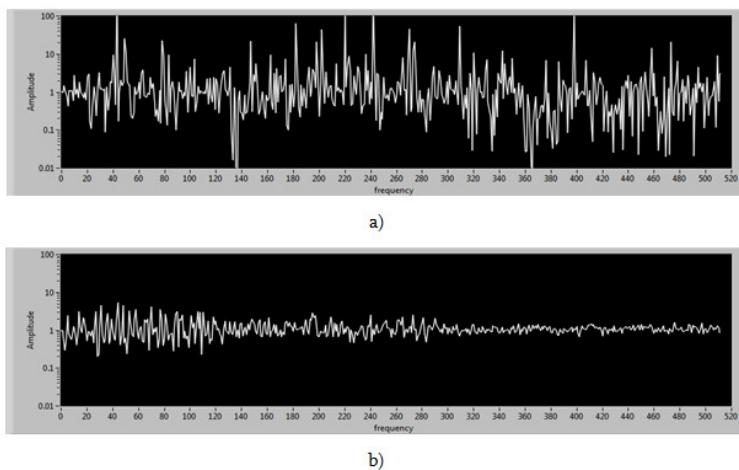


Fig.4. Identification function vs frequency (a - without averaging, b - 10 times ensemble averaging)

3.Weather Conditions and Conclusion

Resent observations shown, that there is a dependency between reliability of detection and the weather. Weather conditions have substantial influence on the quality of clutter suppression.

It is well known that, ideally, the spectrum of received radar signal must be invariant during LFM sawtooth periods, if there is no moving object in coverage, but really, they differ a little due to fading. FFT spectrums and the respective difference have been counted during two adjacent periods. The weather dependent results are shown in Fig. 5 and Fig. 6.

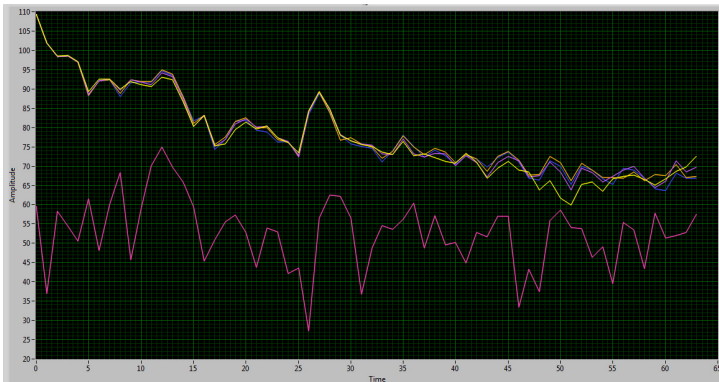


Fig. 5 - spectra of land clutter.

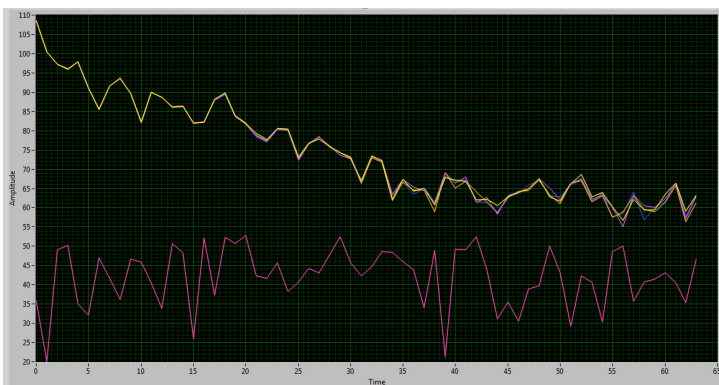


Fig. 6 - spectra of land clutter with a layer of 10 cm snow.

It is obvious that snow coverage significantly improves the situation and the difference of spectra in two adjacent periods is small enough. Probably, it is caused by the reduction of small-scale fading [6].

4. References

1. Komarov, S. Smolskiy - Fundamentals of Short-Range Fm Radar, Artech House Radar Library, 289pp. (2003)
2. Wang W., Cai J., and Yang Y. A Novel Method to Identify Multi-target by Transformable Periods LFM Waveform // Proc. of International Conference on Communications. Circuits and Systems. 2005. V. 2. P. 744-747.
3. A. Muzhikyan, A. Hakhoumian, S. Martirosyan, V. Nikoghosyan, N. Poghosyan, T. Poghosyan, K. Rustamyan, and T. Zakaryan, "Short-Range Ku-Band Hybrid-Mode CW-LFM Radar", Proc. of 11-th INTERNATIONAL RADAR SYMPOSIUM IRS-2010, June 16-18, 2010, Vilnius, Lithuania, pp.1-3.
4. A. Hakhoumian, S. Martirosyan, A. Muzhikyan, V. Nikoghosyan, N. Poghosyan, T. Poghosyan, K. Rustamyan and T. Zakaryan, "Light-Weight Short-Range Ku-Band CW-LFM Radar," Proc. of International Conference 'The Technique of Microwave and THz Waves and its Application in Biomedical and Radar Technologies and in Remote Sensing' (IRPhE'2010), Ashtarak-Aghveran, Armenia, 23-25 September, 2010, pp.87-90
5. C.C. Duarte, B.P. Dorta Naranjo, A.A. Lopez, A.B. del Campo, "CWLFM Radar for Ship Detection and Identification," IEEE Aerospace and Electronic Systems Magazine, vol.22, no.2, Feb. 2007, pp.22-26
6. B. Sklar, "Rayleigh Fading Channels in Mobile Digital Communication Systems. Part I: Characterization" IEEE Communications Magazine, vol. 35, no. 9, Sept 1997, pp.136- 146

Increasing of the signal to noise ratio by using threshold devices

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The using of threshold nonlinear elements-limiters with the optimal linear filter in the received signal processing systems to improve the signal-to-noise ratio (SNR) at the output of the receiver is proposed in this paper. The influence of threshold devices on the output SNR, when the input signal is an additive mixture of information signal (rectangular or bipolar pulses) and Gaussian white noise was investigated. Estimates on the basis of computer simulations show that increase of SNR depends strongly on the shape of the useful signal and the noise power. The proposed system is most effective at low SNR at the input of the system. SNR increases $\approx 1,6$ times compared to the optimal linear filter in the case of bipolar pulses.

The fundamental problem of communication systems is the separation of the information signal from the background noise.

There are numerous studies on signal processing using both linear and nonlinear systems, which are aimed to solve this problem (see [1-9]).

A total transition of telecommunication systems to digital synthesis methods, transmission, reception and processing of information signals was realized in the last decade.

Digital receivers have a threshold nature in which the presence of noise can induce a new and more regularly regime, which will lead to an increase of the coherence degree and cause an increase of the signal to noise ratio (SNR), etc. One of the most striking demonstrations of this type of behavior of nonlinear systems under the influence of noise is the effect of stochastic resonance (SR). SR effect defines a group of phenomena in which the response of a nonlinear system to a weak external signal may significantly increase due to increasing noise intensity in the system [6-13]. The integral output characteristics of the

system, such as the gain and SNR, have a significant maximum at a certain optimal noise level.

Nonlinear threshold elements are usually applied after a linear filter. However, with the presence of noise, using nonlinear devices, it is possible to realize such transformations of the information signal, which are impossible without the noise [3-9, 14]. Therefore, it is possible to get an additional gain of SNR, in the case of nonlinear processing of the signal-noise mixture, before linear filtering, taking into account the nature of the noise and shape of information signal.

In this paper, it is also proposed to use threshold nonlinear elements-limiters with the optimal linear filter in the received signal processing systems to improve the output SNR of the receiver.

The influence of threshold devices on the output SNR, when the input signal is an additive mixture of information signal and Gaussian white noise, was investigated.

Frequently used signals in telecommunication systems – rectangular and bipolar pulses, are considered as useful signal.

Passing of the noise through the limiters

As a non-linear system two voltage limiters (top and bottom) with a zero threshold voltage are proposed, which input-output characteristics are defined as follows:

$$U_{out1} = \begin{cases} U_{in}, & U_{in} \geq 0 \\ 0, & U_{in} < 0 \end{cases}, U_{out2} = \begin{cases} 0, & U_{in} \geq 0 \\ U_{in}, & U_{in} < 0 \end{cases} \quad (1)$$

In this case, if the same noise simultaneously is applied to the inputs of both limiters, the resulting output signals of the limiters will be completely random, but they are well correlated with each other. The summation of these signals exactly recreates the input noise, without any changes (see Fig. 1).

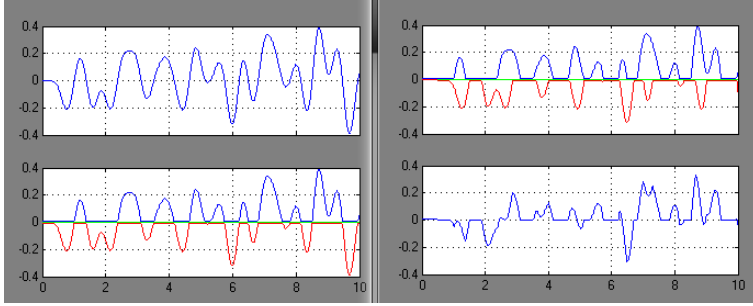


Fig. 1

However, the variance (the power) of the noise, obtained by adding of the output signals of limiters, after any time shift of the one of them, is smaller than the variance of the input noise. Indeed, when U_{out1} is not equal to 0, then $U_{out2} = 0$, and vice versa (when $U_{out1} = 0$) U_{out2} is only negative (see Fig. 1). Consequently, for any relative time shifts of these signals, the time intervals where the values U_{out1} and U_{out2} simultaneously nonzero will be unavoidable. Since one of them takes only positive and the other negative values only, so the instantaneous values (hence power) of the signal, received at summing U_{out1} and U_{out2} , are less than the instantaneous values of the input noise.

To estimate the suppression of the noise depending on the time delay between U_{out1} and U_{out2} , it is need to know the function of cross-correlation of output signals of limiters.

$$R_{+-}(\tau) = \overline{U_{out1}(t) \cdot U_{out2}(t - \tau)} - \overline{U_{out1}(t)} \cdot \overline{U_{out2}(t - \tau)} \quad (2)$$

Because input noise is a normal process with zero mean, it is easy to see that $\overline{U_{out1}(t)} = -\overline{U_{out2}(t - \tau)}$. Taking also into account the fact that the input noise has a Gaussian distribution with a two-dimensional probability density [14]:

$$p[U_{out1}(t), U_{out2}(t - \tau), \tau] = \frac{1}{2\pi\sigma^2\sqrt{1-r^2(\tau)}} \exp\left(\frac{U_{out1}^2(t) + U_{out2}^2(t - \tau) + 2r \cdot U_{out1}(t) \cdot U_{out2}(t - \tau)}{2\sigma^2(1-r^2(\tau))}\right) \quad (3)$$

(σ^2 - is the variance of the input noise, r - the correlation coefficient), it is easy to obtain an expression for the distribution of output signals of limiters:

$$R_{+-}(\tau) = \overline{U_{out1}(t) \cdot U_{out2}(t-\tau)} + \overline{U_{out1}(t)}^2 = \frac{\sigma^2}{2\pi} [-\sqrt{1-r^2} + r \arccos(r+1)] \quad (4)$$

Substituting the value of the correlation coefficient for band limited white noise [14] $r(\tau) = \sin(\omega_0 \tau) / (\omega_0 \tau) = \text{sinc}(\omega_0 \tau)$, we obtain:

$$R_{+-}(\tau) = \frac{\sigma^2}{2\pi} [-\sqrt{1 - (\text{sinc}(\omega_0 \tau))^2} + \text{sinc}(\omega_0 \tau) \cdot \arccos(\text{sinc}(\omega_0 \tau)) + 1] \quad (5)$$

The cross-correlation function graph is shown in Fig. 2.

The interval of cross-correlation between the output noise of the limiters is $\tau \approx 2/\omega_0$, which is close to the interval of correlation of input noise. It is evident to assume that the noise power obtained by adding the limiters outputs U_{out1} and U_{out2} , after their relative time delay, should decrease with increasing the time delay within the cross-correlation interval. A further increase of the delay time (greater the interval of cross-correlation), will not bring to the noise additional reduction.

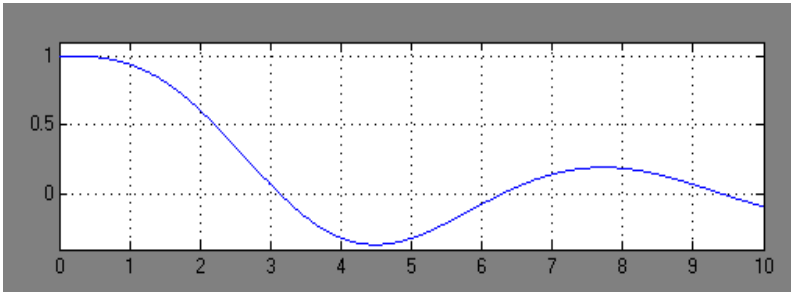


Fig. 2.

Passing of the signal-noise mixture through the limiters

We have considered various useful signals (rectangular monopolar and bipolar pulses) and noise mixture transition through system con-

taining nonlinear elements. Additive mixture of signal and noise are fed to the inputs of the described above limiters. The limiters output signals are adjusted on each other after a certain delay time one of them. The useful signal is filtered from the noise by using an optimal linear filter.

It should be noted that in this case, the optimal linear filter should not be matched with the input pulse, but must be matched to the useful signal in the mix of signal and noise formed in the output of the nonlinear system. Consequently for matched filter design it is necessary to find out the shape of the output information signal after passing of nonlinear system. The shape of the output information signal depends on the shape of input signal, noise level and relative time delay of the limiters. In the case of the high level of noise the output signals of both limiters contain components of information signal due to the stochastic resonance. Hence the sum of output signals can bring not only to the reduction of the noise level and also may lead to the decreasing the power of information signal. Thus to ensure the SNR improvement it should be realized the corresponding time delay for certain information signal.

Modeling of the system

The effectiveness of a nonlinear system described above have been modeled in MATLAB/Simulink environment. Rectangular monopolar and bipolar pulses have been considered as a useful signal (Fig.3).

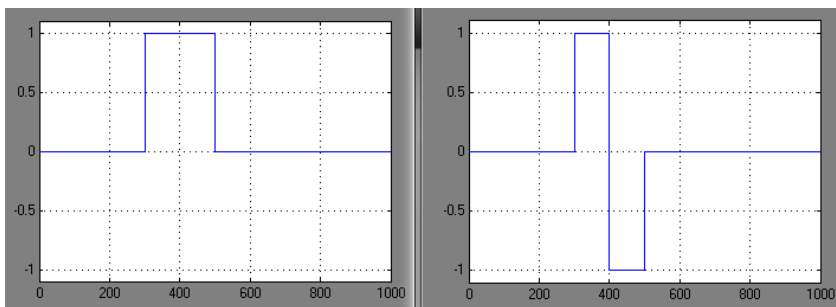


Fig.3.

To estimate the efficiency of the proposed nonlinear system, the SNR at its outputs compared with a SNR at the output of the optimal linear filter. Thus, with nonlinear system modeling for each information signal have been simultaneously realized corresponding optimal filter design. There as the optimality criterion have been chosen maximal value of SNR. The expression of the transfer function of optimal filter for rectangular monopole pulse with τ duration has a following view:

$$K_{\text{opt}}(f\omega) = (k/f\omega) [1 - \exp(-f\omega\tau)] \quad (6)$$

Here have been supposed that maximum value of the filter response is at the $t_0 = \tau$.

Block diagram of the simulated system for unipolar single pulse is shown in Fig. 4.

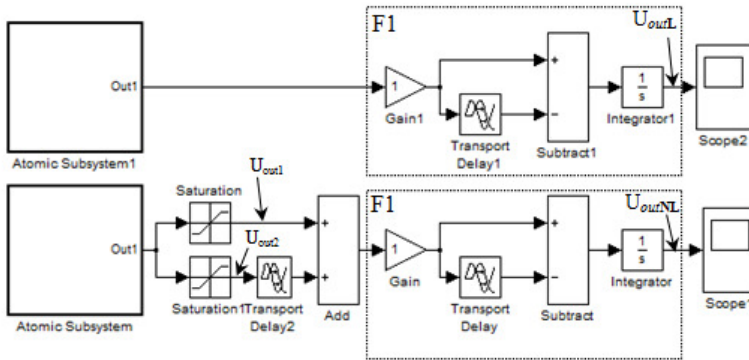


Fig. 4.

The signal-noise mixture is simultaneously supplied to the inputs of both limiters. The output signal of one of them (in our case it is $U_{\text{out}2}(t)$) passing through the delay systems added to the signal $U_{\text{out}1}(t)$. The useful signal is filtering from obtained new signal-noise mixture by using linear optimal filter F2. It should be noted, that optimal filter must be matched with a new useful signal, formed in the output of the nonlinear system.

The shape of useful signal depends on as the information signals as the relative Δt time delay of limiters output signals. Thus, in the case of

alone monopole pulse then $\Delta\tau \ll \tau$, the pulse isn't destroyed, so the F1 filter repeats the F2 filter. And for instance if $\Delta\tau = \tau$ pulse duration increases twice (see. Fig. 5a), consequently the time delay in the filter F1 should be also twiced. The results of simulations, in the case $\Delta\tau = \tau$, are presented in Fig. 5b.

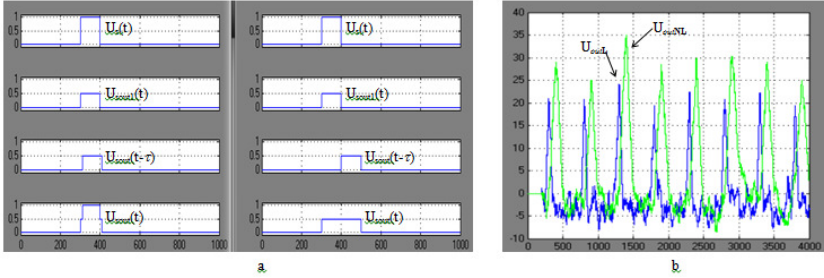


Fig.5.

Estimates show that pretreatment of mixture signal-noise using limiters leads to an increase $SNR \approx 1.2$ times, compared with the optimal linear filter.

The effectiveness of proposed system has been also tested in the case of high power of noise, when the information signal was a bipolar pulse with duration τ (see Fig.3b).

The corresponding matched filter has a transfer coefficient is given by:

$$K(j\omega) = (k/j\omega) \cdot [1 - 2 \exp(-j\omega\tau/2) + \exp(-j\omega\tau)] \quad (7)$$

To avoid the compensation of positive and negative parts of output information signal, after relative time delay and at addition of them, the amount of time delay should be at least equal to the pulse duration τ . Indeed, due to stochastic resonance the output signals of both limiters simultaneously contain as the positive, as the negative parts of the bipolar pulse. Consequently, if the time delay is less than τ the power of useful output signal obtained by adding of outputs of limiters dec-

reases. The shape of formed singnal in the out of nonlinear system with thetime delay $\Delta t = \tau$ presented in Fig.6.

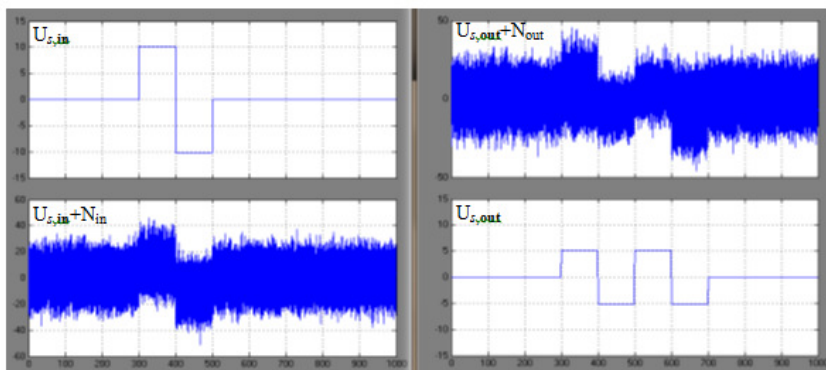


Fig.6.

Spectral densities of this pulse consist of four components.

$$S_{in}(\omega) = (U_0/f\omega)\{(1 - 2\exp(-j\omega\tau/2) + 2\exp(-j\omega\tau) - 2\exp(-3j\omega\tau/2) + \exp(-2j\omega\tau))\} \quad (7)$$

Consequently, for the transfer coefficient of corresponding optimal linear filter obtain:

$$K_m(j\omega) = (k/j\omega) \{1 - 2\exp(-j\omega\tau/2) + 2\exp(-j\omega\tau) - 2\exp(-3j\omega\tau/2) + \exp(-2j\omega\tau)\} \quad (8)$$

Joint modeling scheme of the optimal filter matched with bipolar pulse and the nonlinear filter with limiters is shown in Fig.7, where $T_1 = \tau$, $T = \tau/2$.

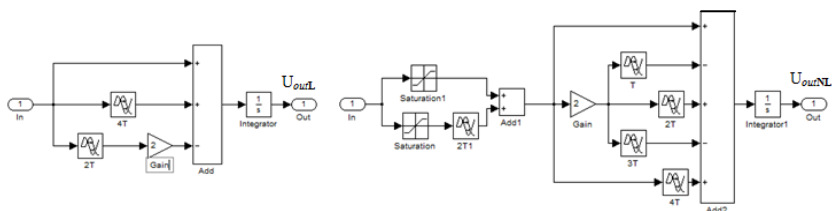


Fig.7.

The simulation results are presented in Fig.8.

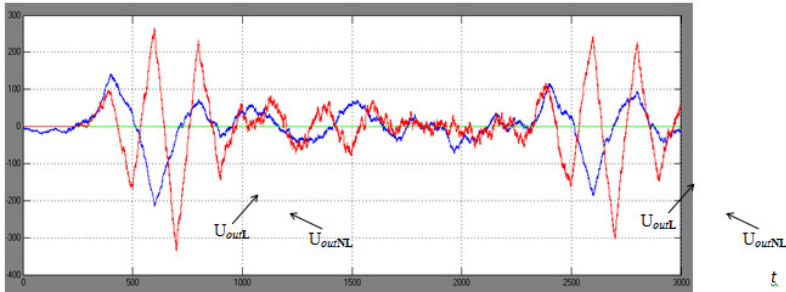


Fig.8.

The estimations based on the results of the simulation indicate that the use of limiters when filtering bipolar signal from the noise increases the signal-to-noise ratio in ≈ 1.6 times compared to the optimal linear filter.

Conclusions

In the proposed non-linear filters by splitting the total signal (the positive and negative parts) and their relative time delay is possible to significantly reduce noise, leaving the energy of the useful signal practically unchanged, thus ensuring the growth of the SNR compared to the optimal linear filter .

Estimates on the basis of computer simulations show that increase of SNR depends strongly on the shape of the useful signal and the noise power. The proposed system is most effective at low SNR at the input of the system. In bipolar useful pulses SNR compared to the optimal linear filter increases $\approx 1,6$ times.

When the more efficient use of nonlinear systems (correct choice of the form of the useful signal and accurate coordination of nonlinear filters to the useful signal) can be expected to further increase of SNR.

References

- [1] P. Jung, Stochastic resonance and optimal design of threshold detectors, Physics Letters A 207 (1995) 93-104.

- [2] F. Chapeau-Blondeau, D. Rousseau, Noise-aided SNR amplification by parallel arrays of sensors with saturation, *Physics Letters A* 351 (2006) 231–237.
- [3] D. Rousseau, G.V. Anand, F. Chapeau-Blondeau, Noise-enhanced nonlinear detector to improve signal detection in non-Gaussian noise, *Signal Processing* 86 (2006) 3456–3465.
- [4] F. Chapeau-Blondeau, D. Rousseau, Noise improvements in stochastic resonance: From signal amplification to optimal detection, *Fluctuation and Noise Letters*, Vol. 2, No. 3 (2002) L221-233.
- [5] F. Duan, D. Abbott, Binary modulated signal detection in a bistable receiver with stochastic resonance, *Physica A* 376 (2007) 173–190.
- [6] A.A. Saha, V.G. Guha, Detectors based on stochastic resonance, Part 2: Convergence analysis and perturbative corrections, *Signal Processing* 87 (2007) 134–147.
- [7] M. D. McDonnell, N.G. Stocks, E.M. Pearce, D. Abbott, Optimal information transmission in nonlinear arrays through suprathreshold stochastic resonance, *Physics Letters A* 352 (2006) 183–189.
- [8] F. Duan, D. Abbott, Signal detection for frequency-shift keying via short-time stochastic resonance, *Physics Letters A* 344 (2005) 401–410.
- [9] S. Peltonen, M. Gabbouj and J. Astola, Nonlinear Filter Design: Methodologies and Challenges, 2nd Int. Symp. on Image SPand Analysis (ISPA01), Pula, Croatia, June 19-21, 2001.
- [10] K. Wiesenfeld, F. Moss, Stochastic resonance and the benefits of noise: from ice ages to crayfish and SQUIDS, *NATURE*, Vol. 373, (1995) 33–36.
- [11] В.С. Анищенко, А.Б. Нейман, Ф. Мосс, Л. Шиманский-Гейер, Стохастический резонанс как индуцированный шумом эффект увеличения степени порядка, *Успехи физических наук*, т. 169, N 1, (1999) 8-38.
- [12] J.C. Comte, S. Morfu, Stochastic resonance: another way to retrieve subthreshold digital data, *Physics Letters A* 309 (2003) 39–43.
- [13] D. Wu, S. Zhu, Stochastic resonance in a bistable system with time-delayed feedback and non-Gaussian noise, *Physics Letters A*, (2006), doi:10.1016/j.physleta.2006.11.006.
- [14] С.И.Баскаков, "Радио/технические цепи и сигналы", Москва, Высшая школа, 1988.

Effect of annealing temperature on Copper-Phthalocyanine thin films investigated by a near-field microwave microscope

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In this paper, the dependence of the crystals and band gap structures of Copper-Phthalocyanine (CuPc) thin films on annealing temperatures was studied. The phase changes from α - to β -phase of CuPc thin films due to different annealing temperatures were observed by using a X-ray diffractions, absorption spectra, scanning electron microscopy, and a near-field microwave microprobe technique. Electrical conductivity measurements of thin films were carried out at different temperatures in range 25-160 °C and were estimated by using near-field microwave microprobe method.

1. Introduction

Copper-Phthalocyanine (CuPc) is one of the promising organic compounds due to its low cost, high thermal and chemical stability, and its excellent field effect response [1,2]. Intrinsic properties of CuPc thin film such as crystal structure, microstructure of grains, band gap and thickness influence the characteristics of organic semiconductor devices. The charge transport properties in organic semiconductors are quite different from those in inorganic materials and thus this issue has attracted a great deal interest in CuPc.

Near-field microwave microprobe (NFMM) techniques with high sensitivity have been developed for the microwave- and millimeter-wave ranges [3-8]. An important ability of the NFMM is contactless and nondestructive and characterization of thin films, in particular, the characterization of electrical properties of films. Contactless and non-destructive characterization techniques are very useful for these applications. NFMM technique, which directly measures the physical properties such as surface resistance of thin films, shows practical promise.

In this paper, we report on the crystal and band gap structure of CuPc films and measure the dependence of sheet resistance on annealing temperatures using a NFMM. The change of absorption spectra of CuPc films was measured as a function of annealing temperatures. The surface roughness and sheet resistance of CdS films with different microstructures and morphologies were investigated by using X-ray diffraction (XRD), scanning electron microscopy (SEM), and NFMM. We used a NFMM coupled to a high-quality dielectric resonator with a distance regulation system at an operating frequency $f = 4.3$ GHz. The changes in the sheet resistance of the CuPc thin films due to different annealing temperatures were investigated using a NSMM by measuring the microwave reflection coefficient S_{11} .

2. Experiment

CuPc powder was purchased from Aldrich Chemical Co. and used without further purification. Slide glass used as substrate. The substrate was cleaned with acetone, ethyl alcohol and distilled water. The CuPc thin films are fabricated by the standard vacuum evaporation technique. CuPc source, contained in a ceramic crucible (bowl), was resistively heated in high vacuum chamber at 10^{-7} Torr. The deposition rate was controlled at 0.02-0.05 nm/sec. The thickness of CuPc films was about 100 nm. The deposition rate and thickness were monitored by a thickness monitor. During deposition substrate temperature was 25 °C (room temperature). In order to study the effect of annealing on the film characteristics of CuPc films, deposited thin films were annealed at 200 °C and 350 °C for 1 hour. The optical absorption of the CuPc thin films was measured in the range 450-850 nm using a SCINCO UV-Vis spectrometer. The changing of the crystal structure of CuPc thin films was inspected by X-ray diffraction. The surface morphology of the thin films was characterized by a SEM. The temperature dependence of DC conductivity of thin films was measured by source meter (Keithley 2400). The AC conductivity of thin films was estimated by NFMM. The experimental setup of our NSMM was described in detail in Ref. [5].

We designed a NSMM system with a tuning fork distance control system to keep a constant distance between the sample and the tip. The probe tip was made of gold wire with a diameter of 50 μm with tapered end size of 1 μm . The probe tip was oriented perpendicular to the sample surface and the other end of the tip was directly connected to a coupling loop in the dielectric resonator. The reflection coefficient S_{11} was measured by network analyzer (Agilent 8722ES). To drive the tuning fork, an AC voltage was applied to one contact on the tuning fork at its resonance frequency using the oscillator of a lock-in amplifier. The resulting current from the other contact was measured by using the current input of the same lock-in amplifier. The output from the lock-in amplifier was fed into the feedback system to control the tip-sample distance using a piezo electric tube (PZT) that supports the sample stage. The probe tip to sample distance was kept at about 10 nm. The sample was mounted onto an x-y-z-translation stage for coarse adjustment which was driven by a computer-controlled microstepping motor with a resolution of 0.01 μm , whereas fine movement of the sample was controlled by a PZT tube. All NFMM measurements were made at the same sample-tip distance.

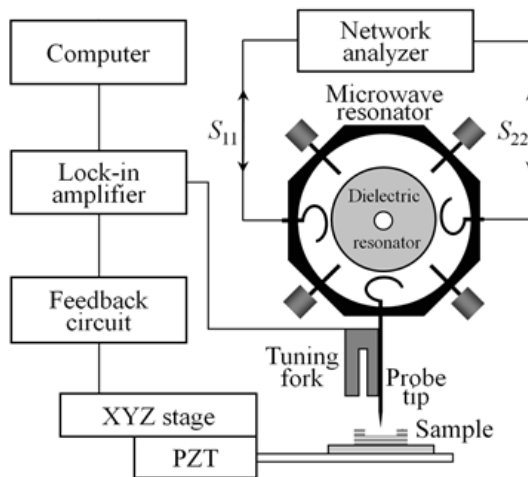


Fig. 1. Experimental setup of NFMM.

3. Results and discussion

As known, the absorption spectra of copper-phthalocyanine thin film depend from crystal structure of thin film [7]. The α -phase of CuPc thin films shows two absorption maxima located at wavelength of 625 and 694 nm while β -phase shows two maxima located at wavelength of 645 and 720 nm.

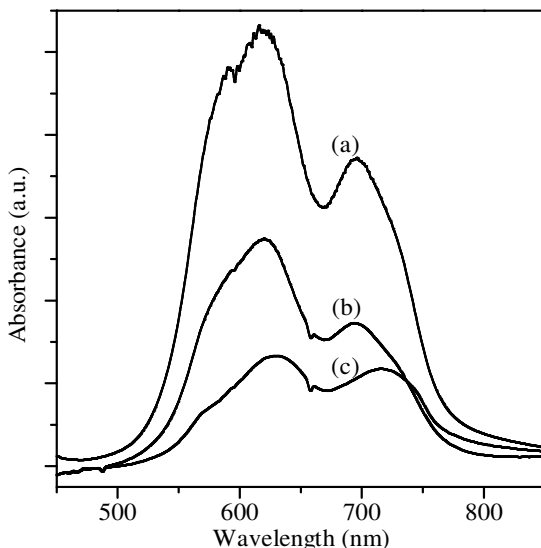


Fig. 2. Optical absorption spectra of CuPc thin film for (a) as-grown sample (deposited at RT), annealed at (b) 200 °C and (c) 350 °C temperature.

Figure 2 shows the optical absorption spectra of CuPc thin films deposited at room temperature and annealed at different temperatures for 1 hour. As shown in Fig. 2, the pattern of absorption spectra depends on annealing temperature of thin film. The higher energy maximum peak on absorption spectra of deposited at room temperature (RT) film larger than the second peak. Similar behavior noticed for thin film deposited on RT and annealed at 200 °C. This behavior represents the typical features of the α -phase of CuPc.

Tab. 1. Positions of absorption peak and energy gap of CuPc deposited on glass and annealed at different temperatures.

Annealing temperature (°C)	Position of absorption peaks (nm)		Direct energy gap (eV)	
	α - phase	β - phase	α - phase	β - phase
RT	611.4	695.8	2.03	1.78
200	619.8	693.5	2.00	1.79
350	629.3	715.4	1.97	1.73

The positions of absorption peaks and related values of the direct energy gap for the different sample shown in Table 1. The positions of peaks (619.8 nm and 693.5 nm) on spectra of thin film annealed at 200 °C approximately converge with positions of maximum peaks on absorption spectra of α -phase CuPc. The positions of peaks on spectra of film annealed at 350 °C are shifted from above mentioned positions on spectra of film annealed at 200 °C. It is allows to suppose, that CuPc thin film deposited on RT and annealed at 350 °C has structure of β -phase. In spite of decreasing, the higher energy maximum peak on absorption spectra of thin film annealed at 350°C still higher than low energy peak. Meanwhile on absorption spectra of β -phase CuPc thin film deposited on ITO glass substrate low energy peak higher than high energy peak [7,8].

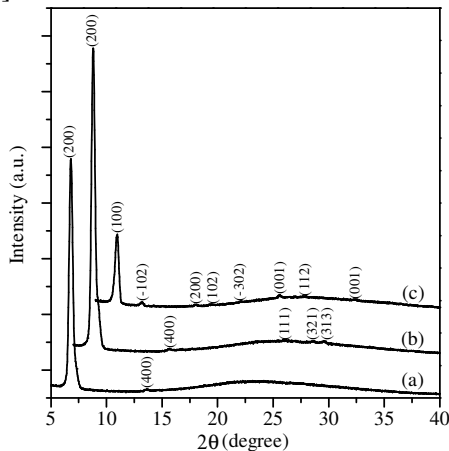


Fig. 3. X-ray diffraction spectra of CuPc thin films for (a) as-grown sample (deposited at room temperature) and annealed at (b) 200 °C and 350 °C temperatures.

For detailing of crystal structure of annealed CuPc thin films they were investigated by X-ray diffraction. Figure 2 shows X-ray diffraction patterns for CuPc thin films annealed at different temperatures. The observed diffraction peaks were identified according [9]. On diffraction spectra of thin film deposited at RT observed two peaks corresponding to the (002) and (400) lattice plane of α -phase CuPc. When annealing temperature is 200 °C observed five peaks corresponding to crystal structure of α -phase CuPc. The X-ray diffraction pattern of the film annealed at 350 °C shows eight peaks corresponding to β -phase. With using of angle positions of these peaks were calculated crystal cell parameters: $a = 1.464$ nm, $b = 0.470$ nm, $c = 1.732$ nm, $\alpha = 90.00^\circ$, $\beta = 105.49^\circ$, $\gamma = 90.00^\circ$. The space group of crystal structure is monoclinic $P2_1/c$ and this is agreed with [9].

SEM images of CuPc thin films deposited at room temperature, annealed at 200 °C and 350 °C is shown in Fig. 4. In the scanning pattern of thin film deposited on RT was observed randomly located curved nanorods with are branching. The cross sections of nanorods are round or oval. As result of the annealing at 200 °C the morphology is changing and nanorods standing up perpendicularly to the substrate surface. The morphology of thin film annealed temperature of 350°C is not homogenous and consists of different regions with various orientations of densely packed nanotubes (Fig. 4 (c)). The whiskers on different parts are differing by orientation as shown on Fig. 4 (d). The nanorods have lamellar like shape and the ends some of them like whiskers as shown on Fig. 4 (e). In some regions elongated whiskers fell down and have horizontal position. These whiskers have different shapes and maximum length of them is about 1 μ m. It is noticeable, that on dependence of annealing conditions the morphology of CuPc thin films deposited on glass substrate is different. SEM image of our thin film deposited at room temperature corresponds to SEM image of thin film deposited at 150 °C on the glass substrate [10].

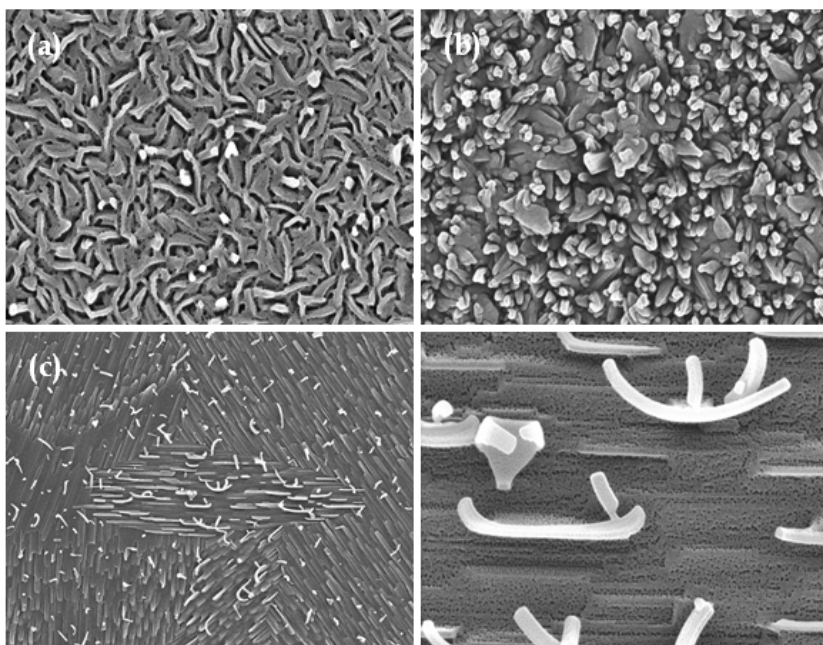


Fig. 4. SEM images of CuPc thin films for (a) as-grown sample (deposited at RT), annealed at (b) 200 °C and (c) 350 °C temperatures. (d) and (e) shows the magnified view of different region of 350 °C annealing sample.

The transport properties of Cu-Pc thin films depend from crystal structure, energy band structure and film morphology [7,10]. The conductivity of CuPc thin film at room temperature is low and order of magnitude is 10^{-10} S/cm [6,10]. Practically the conductivity and mobility measurement of CuPc thin films is carried out by method of field effect transistor (FET). In this paper conductivity properties of thin film were investigated by thermoelectric measurement. On the surface of CuPc thin films were deposited bottom gold contacts and channel length was 50 μm . Voltage and current dependencies were measured at temperature range 25-160 °C. Figure 5 show the conductivity of CuPc thin films annealed at different temperatures as function of heating temperature. The conductivity of thin film annealed at 350 °C (β -phase)

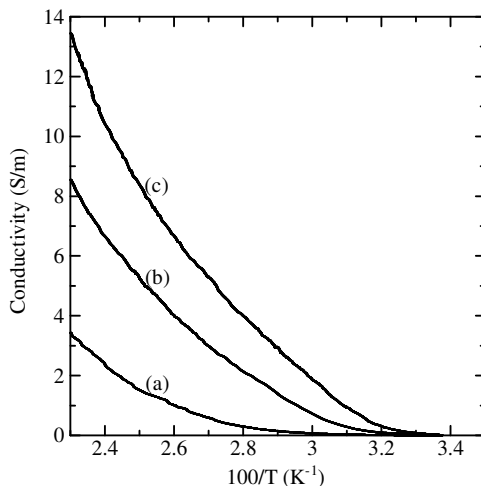


Fig. 5. The conductivity of CuPc thin films as a function of measurement temperature for (a) as-grown sample (deposited at room temperature) and annealed at (b) 200 °C and 350 °C temperatures.

higher than that of annealed at 200 °C (α -phase). Also the conductivity of thin film annealed at 200 °C higher than that deposited at room temperature. The annealing procedure is increasing the conductivity of CuPc thin film deposited on the glass substrate. In case of CuPc thin film deposited on glass shown that in result of phase transformation from the α - to the β - phase conductivity of thin film is changing and the conductivity depends from grain orientation of thin film [5]. From temperature-conductivity dependencies in range of 25-50 °C using Arrhenius plot were estimated thermal activation energy of CuPc thin films deposited at room temperature, annealed at 200 °C and 350 °C. The values of thermal activation energy of thin films were 1.88 eV, 1.84 eV, and 1.80 eV, respectively. These approximate to values of direct band gap energy calculated from absorption spectra (Table 1). The thermal activation energy value of CuPc thin film deposited ITO which was measured by field effect method (1.49 eV [5]) was close to band gap energy (1.59 eV) defined by photoelectron emission and by optical absorption [12]. For study of transport properties of CuPc thin films deposited on ITO glass and annealed at different temperatures

was used ITO bottom contact and Au electrode evaporated on top of sample [7]. Due to heating of the thin film substrate from room temperature to 350 °C the energy gap was changed from 1.8052 to 1.7844 eV [7]. These values are close to our values of thermal activation energy of CuPc thin films deposited on the glass substrate. In our case the difference of the values of thermal activation energy of thin film evaporated on RT and thin film annealed at 350 °C is 0.08 eV.

The conductivity of CuPc thin films deposited on glass substrate at RT and after annealed at different temperature was estimated by NFMM method. Figure 6 shows the measured microwave reflection coefficient S_{11} profile for (a) air and CuPc films for (b) as-growth, (c) 200 °C, and (d) 350 °C annealing temperatures at 1 hour. The matched resonant curve of the (a) air has a minimum level of -52.09 dB, which is the reference level of S_{11} of our measurements. As the annealing temperatures increased up to 350 °C, the minimum reflection coefficient S_{11} increased from -42.09 dB to -39.24 dB as shown in the inset to Fig. 6 (a).

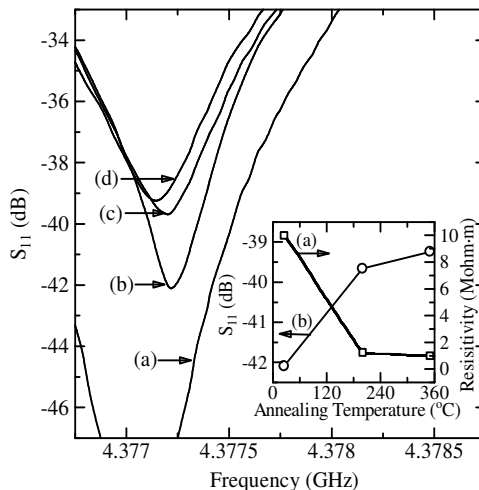


Fig. 6. The microwave reflection coefficient S_{11} profile for (a) air and CuPc thin films for (a) as-grown sample (deposited at RT), annealed at (b) 200 °C and 350 °C temperatures. The inset shows the microwave reflection coefficient (left axis) and resistivity (right axis) vs. annealing temperatures for 100 nm CuPc thin film.

The amount of change in the reflection coefficient depends on the resistivity. An expression how the reflection coefficient S_{11} depended on the electric properties of the CuPc film can be derived by using standard transmission line theory and is given by assuming impedance matching between the microwave probe and the microwave source [8]

$$S_{11} = 20 \log \left| \frac{Z_{in}^R - Z_0}{Z_{in}^R + Z_0} \right|, \quad (1)$$

where Z_0 is the impedance of probe ($Z_0=50 \Omega$), Z_{in}^R is the real part of the complex impedance of the CuPc/glass system and it can be calculated as

$$Z_{in}^R = Z_g \frac{1 + Z_g / R_c}{1 + 2Z_g / R_c}, \quad (2)$$

where Z_g is the impedance of glass ($Z_g=169 \Omega$) and R_c is the sheet resistance of CuPc film ($R_c=1/(\sigma_c t_c)$, where σ_c is the conductivity and t_c is the thickness). Note, that the sheet resistivity of the CuPc films for as-growth condition is about 100 Mohm·m (the electrical conductivity of 10^{-10} S/m [9] and the film thickness is 100 nm). We found that the reflection coefficient S_{11} decreased as the annealing temperature increased up to up to 350 °C. The changes of reflection coefficient S_{11} depends on variations of the sheet resistivity of CuPc film as we can see from Eq. (1) and (2). Note that, CuPc film sheet resistivity decreased about 80 times at the 200 °C and 100 times at the 350 °C annealing temperature. As the annealing temperatures increased up to 350 °C, the sheet resistivity decreased up to 1.0 Mohm·m (inset of Fig. 6).

4. Summary

The annealing procedure at different temperatures changes the crystal phase structure and grain microstructure of CuPc thin films. By annealing at 350 °C the β -phase CuPc thin film is obtained on the glass substrate. Annealing procedure at 350 °C increases conductivity of CuPc thin film compired with them for sample evaporated at RT up to 100 times and decreases thermal activation energy by 0.08 eV. The NFMM method is useful for the measurement of the electromagnetic properties of organic thin films.

Acknowledgements

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3. References

- [1] J. Simon, J. Andre, *Molecular Semiconductors* (Springer, Berlin, 1985).
- [2] C. Leznoff, A. Lever, *Phthlocyanines: Properties and Applications* (VCH, New York, 1993).
- [3] S. Dutta, C. Vlahacos, D. Steinhauer, A. Thanawalla, B. Feenstra, F. Wellstood, S. Anlage, H. Newman, *Appl. Phys. Lett.* **74** (1999) 156.
- [4] B. Knoll, F. Keilmann, A. Kramer, R. Guckenberger, *Appl. Phys. Lett.* **70** (1997) 2667.
- [5] W. Park, J. Kim, K. Lee, *Appl. Phys. Lett.* **79** (2001) 2642.
- [6] B. Friedman, M. Gaspar, S. Kalachikov, K. Lee, R. Levisky, G. Shen, H. Yoo, *J. Am. Chem. Soc.* **127** (2005) 9666.
- [7] B. Friedman, B. Oetiker, K. Lee, *J. Kor. Phys. Soc.* **52** (2008) 588.
- [8] D. Pozar, *Microwave Engineering* (Anderson-Wesley, New York, 1990).
- [9] O. Berger, W. Fisher, B. Adolph, S. Tierbach, V. Melev, J. Sheirer, *J. Mat. Sci.: Mat. Electron.* **11** (2000) 331.
- [10] Yu. Lee, W. Tsai, J. Maa, *Appl. Surf. Sci.* **173** (2001) 352.
- [11] A. Sussman, *J. Appl. Phys.* **7** (1967) 2748.
- [12] A. Ikushima, T. Kanno, Sh. Yoshida, A. Maeda, *Thin Solid Films* **273** (1996) 35.

Preliminary results of C-, Ku-, and Ka-band multi-frequency radiometric measurements of clear air and clouds brightness (antenna) temperatures

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In this paper the results of multi-frequency (at 5.6GHz, 15GHz and 37GHz) and polarization measurements of clear air and clouds brightness (antenna) temperatures are presented, measured under various observation angles. The results have been obtained during the measurements carried out in Armenia from the measuring complex built under the framework of ISTC Projects A-872 and A-1524. The measurements were carried out at cross (vertical and horizontal) polarizations, under various angles of sensing by radiometers of ArtAr-5.6, ArtAr-15, and ArtAr-37 C-, Ku-, and Ka-band combined scatterometric-radiometric systems, developed and built under the frameworks of the above Projects by ECOSERV Remote Observation Centre Co.Ltd., in collaboration with IRPhE of ANAS.

1. Introduction

Radio physical methods and means of remote sensing, such as radars, radiometers and combined radar-radiometers have wide application in meteorology and agriculture, for soil and atmospheric remote sensing. To achieve high accuracy and unambiguity in retrieval soil and snow moistures, soil temperature and snow melting time, precipitation quantity (clouds water content), snowfall and rainfall parameters, a synergy data of various independent and differing techniques and measurements at various frequencies and polarizations is necessary.

Hail and shower cause great and severe damage to agriculture and human properties. To reduce material damage in size it is necessary to have many stations of anti-hail protection equipped by hail clouds

detector-classifiers. Usually, for hail detection powerful Weather Doppler radar is used, operating at short centimeter or millimeter band of waves. These radars cost several hundred thousand USD, have serious disadvantages and cannot solve the problem totally. Therefore a reason is appeared to develop and to produce new kind of detector-identifiers, which will cost cheaper and have additional advantages.

Clouds brightness temperature is a function of many parameters, in which air and particles temperature, fraction type (water or ice) and particles size are the principal variables. The changes of clouds radio brightness temperatures, related with the changes of dielectric properties of particles and their temperatures, depend on frequencies and polarizations of observation. Therefore, by synergetic application of data of multi-frequency and multi-polarization microwave radiometric observations it is possible to detect and to recognize type of the clouds, its water content and the stage of transformation of water vapour and drops of water to hail (to ice). Radiometric observation may not miss the stage of transformation of water vapour and drops of water to hail, because water and ice dielectric constants are very differ and such formations' brightness temperatures will sufficiently vary one from the other. So, for precise and high probable detection and classification of hail clouds, for real time scale recording of hailing time, for upcoming hail storm's start time prediction, for retrieval of hail-stones' probable sizes, for assessment of expected quantity of hail precipitation it is necessary to develop multi-frequency and multi-polarization microwave radiometric system to carry out clear sky and clouds sustainable monitoring. Before that, it is necessary to specify appropriate frequencies and polarization for solution of hail clouds detection, classification and precipitation parameters and quantities assessment by multi-frequency and multi-polarization microwave radiometric system.

In this paper the results of multi-frequency (at 5.6GHz, 15GHz and 37GHZ) and polarization measurements of clear air and clouds brightness (antenna) temperatures are presented, measured under various observation angles. The results have been obtained during the measurements carried out in Armenia from the measuring complex built under the framework of ISTC Projects A-872 and A-1524. The

measurements were carried out at cross (vertical and horizontal) polarizations, under various angles of sensing by radiometers of ArtAr-5.6, ArtAr-15, and ArtAr-37 C-, Ku-, and Ka-band combined scatterometric-radiometric systems, developed and built in collaboration with the Institute of Radiophysics and Electronics of Armenian National Academy of Sciences by ECOSERV Remote Observation Centre Co.Ltd., under the frameworks of the above Projects.

2. Measuring facilities and microwave devices

The measurements were carried out in ECOSERV Remote Observation Centre's experimental site, equipped by indoor and outdoor measuring platforms, scanners, an indoor calibration room and facilities [1-5]. The calibration facilities of the indoor calibration room are used for microwave devices external calibration purposes, by sky and by indoor ABB layer. Besides of calibration needs this indoor measuring complex is used for researches of clouds and precipitations microwave features.

For these measurements C-, Ku-, and Ka-band radiometers of ArtAr-5.6, ArtAr-15, and ArtAr-37 combined scatterometric-radiometric systems (CSRS) were used [6-10]. Detail descriptions of utilized CSRS and the whole experimental site and facilities are possible to find as well in <http://www.ecoservroc.com>. The principal advantages of these unique measuring complex are the capability to perform, multi-frequency, spatio-temporally combined angular and polarization measurements of soil, snow, ice, water surface, clear air, clouds and precipitation microwave, active-passive characteristics, under controlled and far field conditions of sensing. Except of the above mentioned external calibration facilities radiometric channels of all utilized CSRS have internal calibration modules, comprising thermo stabilized noise input generators and thermo stabilized, controlled microwave input keys. These keys in their switched off operational mode are used as internal calibration levels (starting points) for estimation of measured data, such as water surface, external calibration ABB layers and sky brightness temperatures. The noise generators feed the systems' inputs by specified calibration noise signals of 18K of level, for instance,

which are necessary for estimation of observed surfaces and sky brightness temperatures fluctuations.

The measurements of clear air (sky) and clouds brightness temperatures were carried out by two ways. The first way includes direct measurements of sky brightness temperatures from indoor calibration room or from outside located (outdoor) platform. Outdoor platform allowed carry out measurements of sky brightness temperature under various azimuth angles as well.

The second way includes measurements of changes of smooth water surface brightness temperature due to appearance of clouds and precipitation. Scatterometric measurements by a scatterometers of utilized ArtAr-5.6, ArtAr-15, and ArtAr-37 CSRS were used to estimate perturbation level of pool water surface, for correction clouds and precipitation contribution in water surface brightness temperature.

3. The results of measurements

Before and after all series of measurements of sky brightness temperature (antenna temperatures or more exact apparent temperatures) from indoor platform the measurements of indoor ABB layer's brightness temperature (antenna temperatures or more exact apparent temperatures) were carried out. Measurements of sky brightness temperature from indoor platform were carried out consequently under elevation angles 10^0 and 30^0 from nadir at both vertical and horizontal (cross) polarizations of observation. Before and after all series of measurements of sky brightness temperature from outdoor platform the measurements of noise temperatures of ArtAr-5.6, ArtAr-15, and ArtAr-37 CSRS radiometric channels' controlled microwave input keys at their switched off operational modes were carried out, to get starting points for estimation the absolute values of sky brightness temperatures at various frequencies and polarizations.

Before all series of measurements of pool water surface microwave reflective and emissive characteristics, preliminary measurements of indoor ABB layer's and clear sky brightness temperatures at observation angles 10^0 and 30^0 from nadir were performed. After that, the CSRSs were set on the mobile buggy and measurements of smooth pool

water surfaces were carried out at each angle of incidence from 80^0 to 0^0 , by a step of 10^0 . The measurements were carried out at “v” and “h” polarizations for radiometric observations and at “vv” and “vh” or “hh” and “hv” polarizations for scatterometric observations, under various conditions of water t_w and air t_a temperatures. During each series of measurement, at the beginning and at the end of the series, an internal calibration noise signals were switched on and were used for calibration of data obtained by radiometric channels of observation. For scatterometric channels calibration internal calibration signals of a level of $\sim 10^{-11} \text{W}$ were used, estimated to the receivers inputs. These calibration signals allowed approximately estimate absolute values of water surface radar backscattering coefficients and brightness temperatures. Remote control of each CSRS was performed by its personal computer set in the work laboratory built just near the platforms. During measurements the output signals of scatterometric and radiometric receivers were recorded by personal computers as a file. After each series of measurements the saved files have been reproduced on the computer monitor as chart records and were used for further processing and estimation of the observed surface radar backscattering coefficients and brightness (antenna) temperatures at various polarizations and frequencies.

The absolute values of water surface or sky brightness temperatures were estimated from the following equation:

$$T_{Bi} = T_i^K - \frac{U_i^K - U_i^S}{\Delta U_i^{cal}},$$

where, $T_{Bi} = \chi \cdot (273 + t_K)$ is a brightness temperature of the matched load, t_K is a physical temperature of the matched load in centigrade, the coefficient of emission of the matched load χ was estimated and was taken equal to 0.99. U_i^K , U_i^S and ΔU_i^{cal} are outputs of the radiometer receiver, corresponding to the matched load, water surface or sky and the increment of the radiometric output's due to internal calibration noise signal's switching, respectively. The accuracy of estimation of the absolute value of water surface and sky brightness (antenna) tempe-

ratures is about 8-15K, in accordance with the frequency. The accuracy of relative measurements of angular changes of water surface and sky brightness (antenna) temperatures is better than 0.1-0.5K, for C-Ka-band, respectively.

In Fig.1, Fig.2 and Fig.3 measured values of clear sky, lightly clouded and cloudy sky brightness temperatures are presented, measured at 15GHz from indoor calibration room. The results of such measurements at 37GHz are presented in Fig.4.

Comparison of data of Fig.1 - Fig.4 shows that at 37GHz radiothermal contrasts between clear air and cloudy sky brightness temperatures may reach up to 30-40K. At 15GHz radiothermal contrasts are smaller and may reach up to 15-20K.

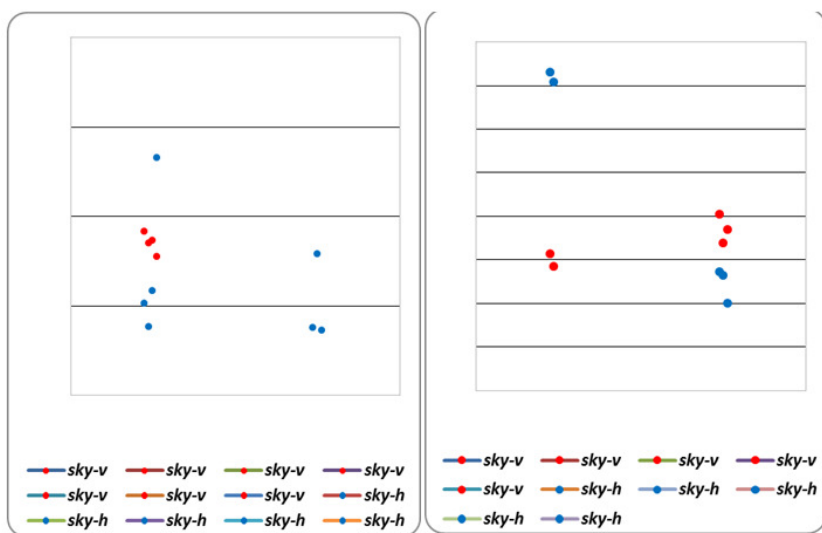


Fig.1 14.03.2011, 15GHz, clear sky and sunny, $t_{\text{air}}=12^{\circ}\text{C}$

Fig.2 20.03.2011, 15GHz, light cloudiness, $t_{\text{air}}=18^{\circ}\text{C}$
Red markers are Pol. “v” and blue markers are Pol. “h”.

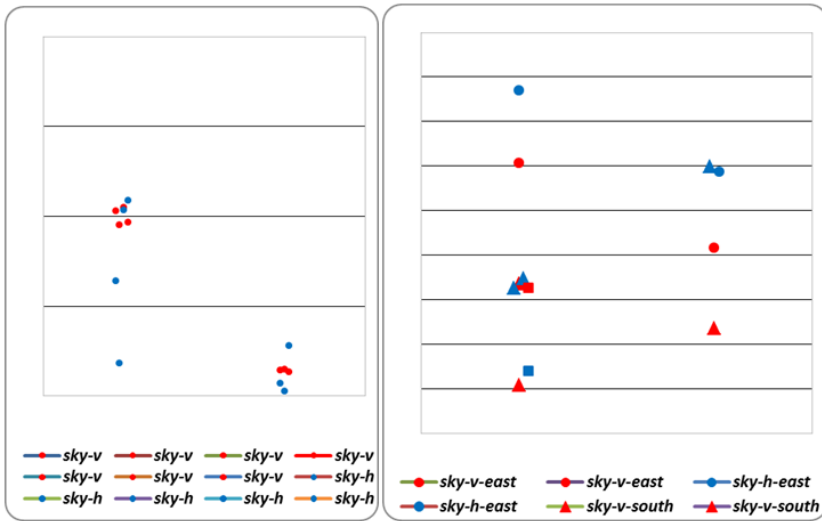


Fig.3 07.03.2011, 15GHz, cloudy to rain, $t_{\text{air}}=3^{\circ}\text{C}$
 Fig.4 37GHz, clear sky, lightly clouded and cloudy conditions
 Red markers are Pol. “v” and blue markers are Pol. “h”.

An interesting situation took place during 06.05.2009 series of measurements of pool water surface microwave emission and reflection. A very fast change in atmospheric condition took place during this experiment. At the beginning of the experiment the sky was clear. During several minutes, when the buggy with ArtAr-5.6, ArtAr-15, and ArtAr-37 CSRSs has stopped at a position corresponding to the angle of incidence 30° , a huge rain-cloud (cumulonimbus) has appeared on the scene, from a mountain, and very fast passed the experimental site. Therefore, the cloud's effect was recorded at the angles of incidence 30° and 20° , only. The measurements were not interrupted, and when the buggy stopped at a position corresponding to the angle of incidence 20° , a cloudburst has begun and continued about 10minutes. Rain-drops and hail have perturbed the pool water surface. When the rain stopped and the water surface has been smoothed, the measurements of water surface reflective and emissive characteristics were continued for angles of incidence 20° , 10° and 0° .

Rain and clouds influences on water surface microwave emissive characteristics in the form of corresponding radio thermal contrasts are presented in Fig.5. Radio thermal contrasts between rain perturbed (RP) and smooth (S) water surfaces were defined by the following equation:

$$\Delta T_{B_{V,h}}(RP-S)[K] = T_{B_{V,h}}(RP) - T_{B_{V,h}}(S) \cdot$$

The results of Fig.5 show that the absolute values of radiothermal contrasts due to nimbus at frequencies 5.6GHz, 15GHz and 37GHz may reach the values 15K, 40K and 60K, respectively. Such a contribution in water surface brightness temperatures may be explained by specular reflections of cloud's emission through water surface. Preliminary processing of obtained data and comparison of the results represented in Figs1-5 shows that cloud's emission may have a high contribution in water surface brightness temperature at both polarizations.

4. Conclusion

Thus, preliminary results of simultaneous and spatially coincident measurements of sky microwave emission at 5.6GHz, 15GHz and 37GHz under clear air, cloudy and rain conditions have shown that radiometer is an actual and significant tool for clouds classification and for estimation of clouds water content. Radiothermal contrasts due to nimbus at frequencies 5.6GHz, 15GHz and 37GHz may reach the values 15K, 40K and 60K, and more, respectively.

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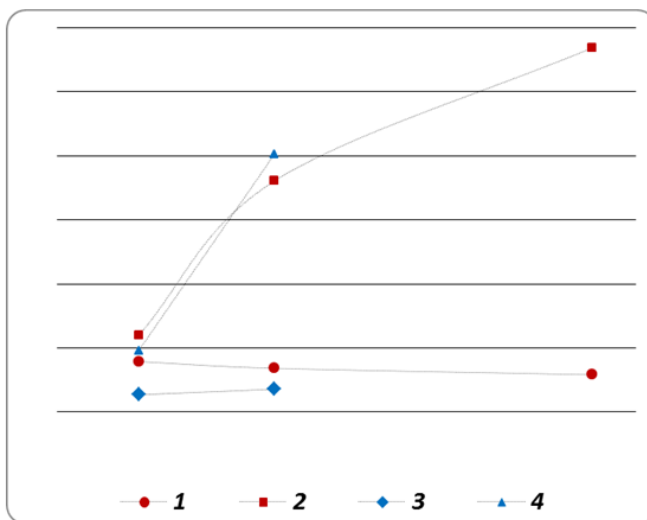


Fig.5 Frequency dependences of water surface radio thermal contrasts due to rain perturbation (RP) and appearance of a “nimbus” (N) at “v” (red markers) and at “h” (blue markers) polarizations

6. References

- [1] A.K.Arakelyan, A.A.Arakelyan, A.K.Hambaryan, M.L.Grigoryan, V.V.Karyan and G.G.Hovhannisyan, *Electromagnetic Waves and Electronic Systems* **1** (2012) 36-46.
- [2] A.A.Arakelyan, A.K.Arakelyan, S.A.Darbinyan, M.L.Grigoryan, I.K.Hakobyan, A.K.Hambaryan, V.K.Hambaryan, V.V.Karyan, M.R.Manukyan and G.G.Hovhannisyan, *IGARSS'08 Conf.* (2008) 4 CD-Rom.
- [3] A.K.Arakelyan, A.K.Hambaryan, V.K.Hambaryan, V.V.Karyan, M.R.Manukyan, M.L.Grigoryan, G.G.Hovhannisyan, A.A.Arakelyan and S.A.Darbinyan, *SPIE Conf. Ocean Sensing and Monitoring II* **7678** (2010) 76780C-1 – 76780C-8.
- [4] A.Arakelyan, A.Hambaryan, V.Hambaryan, V.Karyan, M.Manukyan, M.Grigoryan, G.Hovhannisyan, A.Arakelyan, M.Simonyan and M.Krdyan, *IGARSS Conf.* (2011) CD-Rom.

- [5] M.L.Grigoryan, A.K.Arakelyan, A.K.Hambaryan and A.A.Arakelyan, SPIE Conf. Remote Sensing for Agriculture, Ecosystems and Hydrology XIII **8174** (2011) 81742C-1 - 81742C-9.
- [6] A.K.Arakelyan, A.K.Hambaryan and A.A.Arakelyan, SPIE Conf. Sensing for Agriculture and Food Quality and Safety IV **8369** (2012) 836903-1 – 836903-9.
- [7] A.Arakelyan, I.Hakobyan, A.Arakelyan, A.Hambaryan, M.Grigoryan, E.Vardanyan, V.Karyan, M.R.Manukyan and G.G.Hovhannisyan, Progress in Contemporary Radio Electronics **6** (2006) 44-49.
- [8] A.Arakelyan, I.Hakobyan, A.Arakelyan, A.Hambaryan, M.Grigoryan, G.Hovhannisyan, M.Manukyan, V.Karyan, N.Poghosyan and S.Clifford, Progress in Contemporary Radio Electronics **2** (2007) 41-50.
- [9] A.Arakelyan, M.Grigoryan, A.Hambaryan V.Karyan, G.Hovhannisyan, A.Arakelyan and G.Muradyan, SPIE Conf. Radar Sensor Technology XIII, **7308** (2009) 730813-1-730813-12.

Propagation of a harmonic in time and an arbitrary polarized plane electromagnetic wave through a non-regular one-dimensional media

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The problem of a wave field description for a harmonic in time plane wave propagating through an arbitrary one dimensional absorbing media is considered. The consideration is performed in the framework of the so-called method of counterpropagating waves developed for a case of the normal incidence, so that the results given in this work can be considered in many respects as a generalization on this method. We also apply the suggested approach for calculations of a space distribution of a wave field both inside and outside of different layered constructions.

Introduction

It is well known that the description of a space distribution of a harmonic in time wave field for a one dimensional linear media can be done by mean of many methods such us the classic methods for solution of ordinary differential equations, integral equations method, transfer matrix and scattering matrix methods, Green function approach, imbedding method, phase function method, semiclassical method and so on [1-15]. All these methods were applied for calculations of difference physical situations. Each of them has its advantages and disadvantages. Choose of this or that solution method is dictated generally by the problem statement, namely which aspect of the problem is more interesting: its calculation part, its physical aspect or the possibility to get ideas for consideration of more complex cases. Although these methods aimed to solve the same problem exist parallel for a long time the connection between them almost was not discussed. This question was explored more complicity in the resent work [16], where one the base of the so-called method of counterpropagating

waves the connections between the difference methods were more transparently and completely shown.

In this work we generalize the method developed in the work [17, 18] for the case of a harmonic in time electromagnetic wave oblique incident on an arbitrary one dimensional media. As it is known for that case the space shape of a wave field depends on a wave polarization. Let us consider the harmonic in time electromagnetic wave ($\vec{E}(\vec{r})\exp\{-i\omega t\}$ - electric component, $\vec{H}(\vec{r})\exp\{-i\omega t\}$ - magnetic component) oblique incident on a one dimensional layer with optical properties characterizing by a space dependence and complex dielectric constant:

$$\varepsilon(x) = \begin{cases} 1, & x < x_1, \\ \varepsilon'(x) + i\varepsilon''(x), & x_1 < x < x_2, \\ 1, & x > x_2. \end{cases} \quad (9)$$

Supposing that the wave vector of the incident wave lies in the (x, y) plane for a space part of a wave field outside of the layer region one can write:

$$\vec{E}(\vec{r}) = \begin{cases} \vec{E}_0 \exp\{i\vec{k}\vec{r}\} + \vec{E}_r \exp\{i\vec{k}_r\vec{r}\}, & x < x_1, \\ \vec{E}_t \exp\{i\vec{k}_t\vec{r}\}, & x > x_2, \end{cases} \quad (10)$$

where $\vec{k}$, $\vec{k}_t$, $\vec{k}_r$ are the wave vectors of the incident, transmitted and reflected waves which can be written:

$$\vec{k} = \vec{k}_t = k_{0x} \vec{e}_1 + k_{0y} \vec{e}_2, \quad (11)$$

$$\vec{k}_r = -k_{0x} \vec{e}_1 + k_{0y} \vec{e}_2, \quad (12)$$

where

$$k_{0x} = k_0 \cos \alpha, \quad k_{0y} = k_0 \sin \alpha, \quad k_0 = \omega / c$$

and ω is the field frequency, c is the light velocity, α is the incident angle.

Here the vectors $\vec{e}_1$, $\vec{e}_2$, $\vec{e}_3$ are the unit dimensionless vectors in the x, y, z directions. Note the asymptotic behavior to be analog of Eq.

(10) takes place for the space component of the magnetic field $\vec{H}(\vec{r})$ as well.

From the theoretic and practice points of view the more interesting for consideration are the so-called *TE* and *TM* wave fields. The main property of these waves is that its do not change polarization in the scattering process and any polarized field can be presented by means of superposition of these fields.

For the *TE* waves (which is also called the *s*-polarized waves) all the amplitudes $\vec{E}_0$, $\vec{E}_t$, $\vec{E}_r$ are perpendicular to the incident plane (x, y) , i.e. they are directed along the *z* axis;

$$\vec{E}_0^s = E_0^s \vec{e}_3, \vec{E}_t^s = E_t^s \vec{e}_3, \vec{E}_r^s = E_r^s \vec{e}_3, \quad (13)$$

where E_0^s , E_t^s , E_r^s are the scalar quantities, which define the transmission and reflection amplitudes for the *s* wave;

$$t^s = \frac{E_t^s}{E_0^s}, r^s = \frac{E_r^s}{E_0^s}. \quad (14)$$

For the *TM* waves (which is also called the *p*-polarized waves) all the amplitudes of the field magnetic components $\vec{H}_0$, $\vec{H}_t$, $\vec{H}_r$ are perpendicular to the incident plane (x, y) , which is same that they have over the *z* axis direction;

$$\vec{H}_0^s = H_0^s \vec{e}_3, \vec{H}_t^s = H_t^s \vec{e}_3, \vec{H}_r^s = H_r^s \vec{e}_3, \quad (15)$$

where H_0^s , H_t^s , H_r^s define the amplitudes of the incident, transmitted and reflected waves, so that for the transmission and reflection amplitudes one can write:

$$t^p = \frac{H_t^p}{H_0^p}, r^p = \frac{H_r^p}{H_0^p}. \quad (16)$$

It is well known that when the optical properties of the layer depend on a one coordinate only (see Eq. (9)), then the space dependence of the wave field with asymptotic behavior Eq. (2) is defined by

wave equations containing a one space coordinate. In the regions outside of the layer volume the field has components in the form of a plane wave (see Eq.(2)), so that the electric and magnetic components of s and p wave fields are written.

$$\vec{E}^s(\vec{r}) = \exp\{ik_{0y}y\} E^s(x) \vec{e}_3, \quad (17)$$

$$\vec{H}^p(\vec{r}) = \exp\{ik_{0y}y\} H^p(x) \vec{e}_3. \quad (18)$$

By using the Maxwell equations one can write the following wave equations (see, for example, [18]) for the s -polarized field:

$$\frac{d^2 E^s(x)}{dx^2} + (k_{0x}^2 - U(x)) E^s(x) = 0 \quad (19)$$

where

$$U(x) = \omega^2 (1 - \varepsilon(x)) / c^2. \quad (20)$$

For the p -polarized field the Maxwell equations lead to the following wave equation:

$$\frac{d}{dx} \left(\frac{1}{\varepsilon(x)} \frac{dH^p(x)}{dx} \right) + (k_{0x}^2 - V(x)) H^p(x) = 0, \quad (21)$$

where

$$V(x) = \frac{\omega^2}{c^2} \frac{1 - \varepsilon(x)}{\varepsilon(x)} \sin^2 \alpha. \quad (22)$$

Note that in Eqs. (19), (21) k_{0x} is the x component of the wave vector $\vec{k}$: $k_{0x}^2 = \omega^2 \cos^2 \alpha / c^2$. As it follows from Eq. (9), when $x < x_1$ and $x > x_2$ the quantity $V(x)$ Eq. (22) does not depend on incident angle and $V(x) = 0$.

The standard conditions imposing on the solutions of Eqs. (19), (21) have the following form:

$$E^s(x+0) = E^s(x-0), \quad \left. \frac{dE^s(x)}{dx} \right|_{x+0} = \left. \frac{dE^s(x)}{dx} \right|_{x-0}, \quad (23)$$

$$H^p(x+0) = H^p(x-0), \frac{1}{\varepsilon(x+0)} \frac{dH^p}{dx} \Big|_{x+0} = \frac{1}{\varepsilon(x-0)} \frac{dH^p}{dx} \Big|_{x-0}. \quad (24)$$

These equalities should be performed for any value of x even for the values when the media dielectric constant changes discontinuously.

In this work we investigate Eq. (21) by the following the method developed in the works [17] for solving the one-dimensional Schrodinger equation, which is mathematically identical to Eq. (10). For clarity below in the first paragraph we will introduce some results of the method of counterpropagating waves, where a linear set of differential equations for the amplitudes of the counterpropagating waves are derived.

2. Brief presentation of the of the method of counterpropagating waves

The main idea of this method is a change of a wave equation on a set of two differential equations, so that for any space point the unknowns would be the amplitudes of the counterpropagating waves. Let us consider the couple of functions $c(x)$ and $d(x)$, which satisfy to the following set of linear differential equations:

$$\frac{dc(x)}{dx} = -\frac{iU(x)}{k_{0x}} c(x) - \frac{iU(x)}{k_{0x}} d(x) \exp\{-i2k_{0x}x\}, \quad (25)$$

$$\frac{d d(x)}{dx} = \frac{iU(x)}{k_{0x}} d(x) + \frac{iU(x)}{k_{0x}} c(x) \exp\{i2k_{0x}x\}. \quad (26)$$

As it was mentioned the method of counterpropagating waves was developed for consideration of the s wave equation (19). It is easy to check that the function $E^s(x)$, constructing with help of the functions $c(x)$, $d(x)$ in accordance the following formula

$$E^s(x) = c(x) \exp\{ik_{0x}x\} + d(x) \exp\{-ik_{0x}x\}, \quad (27)$$

satisfies to Eq. (19).

It is important to note that consideration of the wave equation (19) on the base of the set of Eqs. (25), (26) is mainly motivated by the fact

that the derivation of the wave function written by means of the functions $c(x)$ and $d(x)$ has the form of

$$\frac{dE^s(x)}{dx} = ik_{0x}(c(x)\exp\{ik_{0x}x\} - d(x)\exp\{-ik_{0x}x\}). \quad (28)$$

The set of Eqs. (25), (26) can be solved with different initial or boundary conditions, which define problem statement of the or the asymptotic behavior of the wave function. In a physical point of view the more interesting is a boundary type problem when the values of the functions $c(x)$ and $d(x)$ are given in the different regions left and right of the layer;

$$c(x_1) = c_0, \quad d(x_2) = d_0. \quad (29)$$

Note that the quantities c_0 and d_0 correspond to the magnitudes of the waves converging to the layer. The wave field determining in accordance to condition (29) is called as a converging solution. It is easy to check that in this case the space dependence of the wave field satisfies to the following integral equations

$$E^s(x) = c_0 \exp\{ik_{0x}x\} + d_0 \exp\{-ik_{0x}x\} + \int_{x_1}^{x_2} V(x') G_0^{(+)}(x, x') E^s(x') dx',$$

where the function

$$G_0^{(+)}(x, x') = -\frac{i}{2k_{0x}} \exp\{ik_{0x}|x - x'|\}$$

is the converging Green function for a free space.

3. The method of counterpropagating waves for the p -polarized waves

Below, we will investigate propagation features of p -polarized field. For convenience we will not use in the notations of unknown functions the index p : $H^p(x) \equiv H(x)$ and so on. Let us present the required wave field $H(x)$ with help of two functions $a(x)$, $b(x)$ by means of the following formula:

$$H(x) = a(x) \exp\{ik_{0x}x\} + b(x) \exp\{-ik_{0x}x\}. \quad (30)$$

The given formula implies that a one unknown function $H(x)$ is changed by two unknown functions $a(x)$, $b(x)$. This allows to get a certain degree of freedom in a choice of the functions $a(x)$, $b(x)$. Such freedom permits to impose a physically motivated connection between these unknown functions. So, the functions $a(x)$, $b(x)$ can be chosen so that the derivation of $H(x)$ will be written in the form of

$$\frac{dH(x)}{dx} = ik_{0x} \varepsilon(x) (a(x) \exp\{ik_{0x}x\} - b(x) \exp\{-ik_{0x}x\}). \quad (31)$$

As will be shown below when a certain condition takes place, then as it follows from Eqs. (30), (31), the functions $a(x)$, $b(x)$ can be interpreted as the amplitudes of the counterpropagating waves propagating in positive and negative directions. Using Eqs. (30), (31) one can write

$$a(x) = \frac{\exp\{-ik_{0x}x\}}{2} \left[H(x) - \frac{i}{k_{0x}} \frac{1}{\varepsilon(x)} \frac{dH(x)}{dx} \right], \quad (32)$$

$$b(x) = \frac{\exp\{ik_{0x}x\}}{2} \left[H(x) + \frac{i}{k_{0x}} \frac{1}{\varepsilon(x)} \frac{dH(x)}{dx} \right]. \quad (33)$$

Considering a first order derivation of these equalities and using the wave equation (21), one can show that the couple of functions $a(x)$, $b(x)$ satisfy to the following set of first order differential equations

$$\frac{da(x)}{dx} = -\frac{iQ_1(x)}{2k_{0x}} a(x) - \frac{iQ_2(x)}{2k_{0x}} b(x) \exp\{-i2k_{0x}x\}, \quad (34)$$

$$\frac{db(x)}{dx} = \frac{iQ_1(x)}{2k_{0x}} b(x) + \frac{iQ_2(x)}{2k_{0x}} a(x) \exp\{i2k_{0x}x\}, \quad (35)$$

where

$$Q_1(x) = V(x) + k_{0x}^2 (1 - \varepsilon(x)), \quad Q_2(x) = V(x) - k_{0x}^2 (1 - \varepsilon(x)). \quad (36)$$

By using Eqs. (34), (35) it is easy to check that the following equality takes place

$$\frac{da(x)}{dx} \exp\{ik_{0x}x\} + \frac{db(x)}{dx} \exp\{-ik_{0x}x\} = -ik_{0x}(1 - \varepsilon(x))(a(x) \exp\{ik_{0x}x\} - b(x) \exp\{-ik_{0x}x\}).$$

The set of Eqs. (34), (35) can be solved at difference initial or boundary conditions. Note, that in the case of an initial condition the known values of the functions $a(x)$, $b(x)$ are given in a same point or an asymptotic. For a boundary condition problem, the known values of the required functions are considered in a different asymptotic. So, for a boundary type problem, when

$$a(x_1) = 1, b(x_2) = 0, \quad (37)$$

the functions $a(x)$ and $b(x)$ are given in different points.

Let us denote the functions $a(x)$, $b(x)$ determined from Eqs. (34), (35) in accordance with condition (37) as

$$a(x) = a_l(x), b(x) = b_l(x). \quad (38)$$

Here, the use of the letter l in the notations of $a(x)$, $b(x)$ is motivated by that the function $H(x)$ (30) describes a scattering wave field incidents to a slab from its left side;

$$H_{left}(x) = a_l(x) \exp\{ik_{0x}x\} + b_l(x) \exp\{-ik_{0x}x\} \quad (39)$$

and

$$H_{left}(x) = \begin{cases} \exp\{ik_{0x}x\} + r(k_{0x}) \exp\{-ik_{0x}x\}, & x < x_1, \\ t(k_{0x}) \exp\{ik_{0x}x\}, & x > x_2, \end{cases} \quad (40)$$

where $t(k_{0x})$ and $r(k_{0x})$ are the transmission and reflection amplitudes of a scattering wave, when a wave filed is initially excited by means of a wave incident from a slab left side. Note, that the quantities $t(k_{0x})$, $r(k_{0x})$ are defined in an opposite with respect to the boundary condition (37) asymptotic;

$$a_l(x_2) = t(k_{0x}), b_l(x_1) = r(k_{0x}). \quad (41)$$

Note that in accordance to Eqs. (37), (38)

$$a_l(x_1)=1, b_l(x_2)=0. \quad (42)$$

It is also easy to formulate a boundary type problem, when the solution of Eqs. (34), (35) describes a wave field incident to a slab from its right side. For this case the unknown functions have the following behavior:

$$a(x_1)=0, b(x_2)=1. \quad (43)$$

For the functions $a(x)$, $b(x)$ satisfying to the condition (43) we will take the notations

$$a(x)=a_r(x), b(x)=b_r(x). \quad (44)$$

Here the use of the letter r in the indexes of unknown functions due to the corresponding wave field, i.e.

$$H_{right}(x)=a_r(x)\exp\{ik_{0x}x\}+b_r(x)\exp\{-ik_{0x}x\}. \quad (45)$$

Describes the right scattering problem, when the wave field has an asymptotic behavior in the form of

$$H_{right}(x)=\begin{cases} s(k_{0x})\exp\{-ik_{0x}x\}, & \text{when } x \rightarrow -\infty, \\ \exp\{-ik_{0x}x\}+p(k_{0x})\exp\{ik_{0x}x\}, & \text{when } x \rightarrow +\infty. \end{cases} \quad (46)$$

The functions $s(k_{0x})$ and $p(k_{0x})$ are the transmission and reflection amplitudes when a wave falls to a slab from its right side. These characteristics of a scattering process are presented as

$$a_r(x_2)=p(k_{0x}), b_r(x_1)=s(k_{0x}). \quad (47)$$

Note that in accordance to Eqs. (43), (44)

$$a_r(x_1)=0, b_r(x_2)=1. \quad (48)$$

So, we reduced the field determination problem for the p -polarized wave to the solution of some set of differential equations. It is shown that two known asymptotic behaviors of the wave field such as left and right scattering problems (see Eqs. (40) and (46)) correspond to boundary problems for Eqs. (34), (35). Despite to a transference connection between the problem statement and the boundary conditions the set of Eqs. (34), (35) does not make the solution way more easy then it was in the case of the wave equation (see Eq. (21)). This is due to the integration of Eqs. (34), (35) in accordance to a boundary condition should start from different points. The problem solution needs to another mathematical formulation, namely it should be formulated as an initial problem, when the values of unknown functions are given in a same point.

4. The scattering and transfer matrixes

As it was shown the formulation of a wave field determination in the form of a boundary problem for a set of differential equations is very advantageous in a physical point of view, but it meets difficulties in a performing of mathematical calculations.

Let us consider any solution of the wave equations by means of the wave field solutions corresponding to the left and right scattering problem. Any solution is defined up to two arbitrary constants, which can be chosen arbitrarily. As these constants we will choose the amplitudes of waves converging to a slab;

$$H(x) = \begin{cases} a_1 \exp\{ik_{0x}x\} + b_1(k_{0x}) \exp\{-ik_{0x}x\}, & x < x_1, \\ a_2(k_{0x}) \exp\{ik_{0x}x\} + b_2 \exp\{-ik_{0x}x\}, & x > x_2, \end{cases} \quad (49)$$

where a_1 , b_2 are the amplitudes of the converging waves, which are considered as given quantities, and b_1 , a_2 are the amplitudes of the diverging waves, which are considered as required quantities. The quantities b_1 , a_2 depend on the wave number k_{0x} and the given constants a_1 , b_2 . Note, that in Eq.(30), the unknown functions $a(x)$, $b(x)$ should satisfy to the following boundary condition:

$$a(x_1) = a_1, b(x_2) = b_2. \quad (50)$$

By using the boundary conditions for the wave fields of the left and right scattering problem Eq. (37), Eq. (43) it is easy to see that any solution defined by values of the amplitudes a_1 , b_2 can be written:

$$H(x) = a_1 H_{left}(x) + b_2 H_{right}(x). \quad (51)$$

In accordance to Eq. (38), Eq. (44) the functions $a(x)$ and $b(x)$, when the condition Eq. (50) is imposed, can be presented:

$$a(x) = a_1 a_l(x) + b_2 a_r(x), b(x) = a_1 b_l(x) + b_2 b_r(x). \quad (52)$$

Comparing Eq. (30) and Eq. (49) it is easy to see, that

$$a(x_2) = a_2(k_{0x}), b(x_1) = b_1(k_{0x}). \quad (53)$$

Taking into account, that (see Eq. (41) and Eq. (47))

$$a_l(x_2) = t(k_{0x}), b_l(x_1) = r(k_{0x}) \text{ and } a_r(x_2) = p(k_{0x}), b_r(x_1) = s(k_{0x})$$

and using Eq. (53), from Eq. (52) one can write

$$a_2(k_{0x}) = t(k_{0x}) a_1 + p(k_{0x}) b_2, b_1(k_{0x}) = r(k_{0x}) a_1 + s(k_{0x}) b_2. \quad (54)$$

These two equalities determine the well known form of the scattering matrix, which expresses the connection between the amplitudes of converging and diverging waves;

$$\begin{pmatrix} a_2(k_{0x}) \\ b_1(k_{0x}) \end{pmatrix} = \hat{S}(k_{0x}) \begin{pmatrix} a_1 \\ b_2 \end{pmatrix}, \hat{S}(k) = \begin{pmatrix} t(k_{0x}) & p(k_{0x}) \\ r(k_{0x}) & s(k_{0x}) \end{pmatrix}. \quad (55)$$

Using (55) it is easy to determine a transfer matrix for or a monodromy matrix of the considered problem:

$$\begin{pmatrix} a_2(k_{0x}) \\ b_2(k_{0x}) \end{pmatrix} = \hat{M}(k_{0x}) \begin{pmatrix} a_1 \\ b_1 \end{pmatrix}, \quad \hat{M}(k_{0x}) = \begin{pmatrix} t(k_{0x}) - \frac{p(k_{0x})r(k_{0x})}{t(k_{0x})} & \frac{p(k_{0x})}{s(k_{0x})} \\ \frac{-r(k_{0x})}{s(k_{0x})} & \frac{1}{t(k_{0x})} \end{pmatrix}, \quad (56)$$

which connects the amplitudes of waves propagating in opposite directions in regions left and right of a slab.

Let us consider two independent solutions of the set (34), (35): $a_1(x)$, $b_1(x)$ and $a_2(x)$, $b_2(x)$. It is easy to show that the following equality takes place:

$$a_1(x)b_2(x) - b_1(x)a_2(x) = \text{const}. \quad (57)$$

As it follows from (57)

$$a_1(x) \frac{db_2(x)}{dx} + b_2(x) \frac{da_1(x)}{dx} - b_1(x) \frac{da_2(x)}{dx} - a_2(x) \frac{db_1(x)}{dx} = 0. \quad (58)$$

Note that both couples of functions $a_1(x)$, $b_1(x)$ and $a_2(x)$, $b_2(x)$ satisfy two the same set of Eqs. (34), (35), which express connections between the unknown functions and their derivations.

If one choose as independent solutions of Eqs. (34), (35) the unknown functions corresponding to the left and right scattering problems (see Eq. (38) and Eq. (44))

$$a(x) = a_l(x), \quad b(x) = b_l(x) \quad \text{and} \quad a(x) = a_r(x), \quad b(x) = b_r(x), \quad (59)$$

when from Eq. (57) one can write:

$$a_l(x)b_r(x) - b_l(x)a_r(x) = \text{const}. \quad (60)$$

Since this equality should take place for any values of x , then one in right to write

$$a_l(x_1)b_r(x_1) - b_l(x_1)a_r(x_1) = a_l(x_2)b_r(x_2) - b_l(x_2)a_r(x_2).$$

Using Eqs. (41), (42) and Eqs. (47), (48) it is easy to see that

$$t(k_{0x}) = s(k_{0x}), \quad (61)$$

i.e. the transmission amplitudes of the left and right scattering problems equals to each other. As it follows from Eq. (61) the transfer matrix determinant equals (see Eq. (56)) to one: $\det[\hat{M}] = 1$.

5. Converging and diverging solutions of a wave equations

Below we will mention a dependence of the required functions $a(x)$, $b(x)$, determining the space form of a wave field (see Eq. (30)), on the wave number of an incident wave:

$$a(x) = a(k_{0x}, x), \quad b(x) = b(k_{0x}, x). \quad (62)$$

Let us consider the functions $a(-k_{0x}, x)$, $b(-k_{0x}, x)$, which are obtained from the functions $a(k_{0x}, x)$, $b(k_{0x}, x)$ as a result of a non-algebraic action such as a change of k_{0x} sign on opposite (see [16]). Let us suppose that the functions $a(k_{0x}, x)$, $b(k_{0x}, x)$ are defined in accordance to the boundary condition Eq. (50), where the amplitudes of the converging waves are given. Since the amplitudes a_1 , b_2 do not depend on k_{0x} , the functions $a(-k_{0x}, x)$, $b(-k_{0x}, x)$ will satisfy to the same condition as the function $a(k_{0x}, x)$, $b(k_{0x}, x)$;

$$a(k_{0x}, x_1) = a(-k_{0x}, x_1) = a_1, \quad b(k_{0x}, x_2) = b(-k_{0x}, x_2) = b_2. \quad (63)$$

In the function $H(x, k_{0x})$ (30) the change of k_{0x} sign leads to the function

$$H(-k_{0x}, x) = a(-k_{0x}, x) \exp\{-ik_{0x}x\} + b(-k_{0x}, x) \exp\{ik_{0x}x\}, \quad (64)$$

defined by the following asymptotic behavior (see Eq.(49));

$$H(x, -k_{0x}) = \begin{cases} a_1 \exp\{-ik_{0x}x\} + b_1(-k_{0x}) \exp\{ik_{0x}x\}, & x < x_1, \\ a_2(-k_{0x}) \exp\{-ik_{0x}x\} + b_2 \exp\{ik_{0x}x\}, & x > x_2, \end{cases} \quad (65)$$

where (see Eq. (53))

$$a(-k_{0x}, x_2) = a_2(-k_{0x}), \quad b(-k_{0x}, x_1) = b_1(-k_{0x}). \quad (66)$$

In accordance with Eqs. (34), (35), when k_{0x} is changed to $-k_{0x}$ for the functions $b(-k_{0x}, x)$, $a(-k_{0x}, x)$ one can write down:

$$\frac{da(-k_{0x}, x)}{dx} = \frac{iQ_1(x)}{2k_{0x}} a(-k_{0x}, x) + \frac{iQ_2(x)}{2k_{0x}} b(-k_{0x}, x) \exp\{i2k_{0x}x\}, \quad (67)$$

$$\frac{db(-k_{0x}, x)}{dx} = -\frac{iQ_1(x)}{2k_{0x}} b(-k_{0x}, x) - \frac{iQ_2(x)}{2k_{0x}} a(-k_{0x}, x) \exp\{-i2k_{0x}x\}, \quad (68)$$

It is easy to see that the functions $b(-k_{0x}, x)$, $a(-k_{0x}, x)$ satisfy to the same set of equations as the functions $a(k_{0x}, x)$, $b(k_{0x}, x)$. If one changes $b(-k_{0x}, x)$ on $a(k_{0x}, x)$ and $a(-k_{0x}, x)$ on $b(k_{0x}, x)$, when Eq. (67) passes to Eq. (35) and when Eq. (68) passes to Eq. (34).

As it follows from Eqs. (67), (68) the functions $H(-k_{0x}, x)$ satisfies to the wave equations (21) as well as the functions $H(k_{0x}, x)$;

$$\frac{d}{dx} \left(\frac{1}{\varepsilon(x)} \frac{dH(\pm k_{0x}, x)}{dx} \right) + (k_{0x}^2 - V(x)) H(\pm k_{0x}, x) = 0. \quad (69)$$

These solutions are linear independent from each other since the boundary conditions for them (see Eq. (50) and Eq. (63)) are differ. So, in accordance with Eq. (49), for the solution $H(k_{0x}, x)$ the quantities A_1 , B_2 are the amplitudes of the converging (incoming) waves, while, as it follows from Eq. (65), for $H(-k_{0x}, x)$ these quantities are the amplitudes of the diverging (outgoing) waves. We will call the solution $H(-k_{0x}, x)$ as a conjugate regarded to the solution $H(k_{0x}, x)$

If one apply the scattering matrix property Eq. (55) for presentation of a connection, which exists between the coefficients of the solution $H(-k_{0x}, x)$ Eq. (65), it is possibly write down;

$$\begin{pmatrix} b_2 \\ a_1 \end{pmatrix} = \hat{S}(k_{0x}) \begin{pmatrix} b_1(-k_{0x}) \\ a_2(-k_{0x}) \end{pmatrix} \text{ or } \begin{pmatrix} a_1 \\ b_2 \end{pmatrix} = \hat{S}^r(k_{0x}) \begin{pmatrix} a_2(-k_{0x}) \\ b_1(-k_{0x}) \end{pmatrix}, \quad (70)$$

where $\hat{S}^T(k_{0x})$ is a transpose of the matrix $\hat{S}(k_{0x})$ (55);

$$\hat{S}^T(k_{0x}) = \begin{pmatrix} t(k_{0x}) & r(k_{0x}) \\ p(k_{0x}) & s(k_{0x}) \end{pmatrix}.$$

Note that for $H(-k_{0x}, x)$ the quantities $a_2(-k_{0x})$, $b_1(-k_{0x})$ are the amplitudes of the outgoing waves. By changing in the transpose matrix $\hat{S}^T(k_{0x})$ the quantity k_{0x} on $-k_{0x}$, one will get the Hermitian conjugate of the scattering matrix:

$$\hat{S}^\dagger(k_{0x}) = \hat{S}^T(-k_{0x}) = \begin{pmatrix} t(-k_{0x}) & r(-k_{0x}) \\ p(-k_{0x}) & s(-k_{0x}) \end{pmatrix}. \quad (71)$$

Using the matrix $\hat{S}^\dagger(k_{0x})$ the second equation of (70) one can written:

$$\hat{S}^\dagger(k_{0x}) \begin{pmatrix} a_2(k_{0x}) \\ b_1(k_{0x}) \end{pmatrix} = \begin{pmatrix} a_1 \\ b_2 \end{pmatrix}. \quad (72)$$

Multiplying the both sides of this equation on $\hat{S}(k_{0x})$, we will get:

$$\hat{S}(k_{0x}) \hat{S}^\dagger(k_{0x}) \begin{pmatrix} a_2(k_{0x}) \\ b_1(k_{0x}) \end{pmatrix} = \hat{S}(k_{0x}) \begin{pmatrix} a_1 \\ b_2 \end{pmatrix}. \quad (73)$$

A simple comparison of Eq. (73) and Eq. (55) shows that

$$\hat{S}(k_{0x}) \hat{S}^\dagger(k_{0x}) = \hat{I} \text{ or } \hat{S}^\dagger(k_{0x}) = \hat{S}^{-1}(k_{0x}). \quad (74)$$

This result shows that the scattering matrix is a unitary matrix. Using the first equation of (74) and Eqs. (55), (71) one can write:

$$\begin{pmatrix} t(k_{0x}) & p(k_{0x}) \\ r(k_{0x}) & s(k_{0x}) \end{pmatrix} \cdot \begin{pmatrix} t(-k_{0x}) & r(-k_{0x}) \\ p(-k_{0x}) & s(-k_{0x}) \end{pmatrix} = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix}. \quad (75)$$

This matrix equation is equivalent to three algebraic equations

$$t(k_{0x})t(-k_{0x}) + r(k_{0x})r(-k_{0x}) = 1, \quad (76)$$

$$s(k_{0x})s(-k_{0x}) + p(k_{0x})p(-k_{0x}) = 1, \quad (77)$$

$$t(k_{0x})r(-k_{0x}) + s(-k_{0x})p(k_{0x}) = 0. \quad (78)$$

Note that here we took into account the equality of the scattering amplitudes of the left and right scattering problems: $t(k_{0x}) = s(k_{0x})$ (see Eq. (61)).

Using Eqs. (76)-(78), except the form of Eq. (56), the transfer matrix $\hat{M}(k_{0x})$ can be presented in the following forms as well:

$$\hat{M}(k_{0x}) = \begin{pmatrix} \frac{1}{t(-k_{0x})} & \frac{-r(-k_{0x})}{t(-k_{0x})} \\ \frac{-r(k_{0x})}{t(k_{0x})} & \frac{1}{t(k_{0x})} \end{pmatrix} = \begin{pmatrix} \frac{1}{s(-k_{0x})} & \frac{p(k_{0x})}{s(k_{0x})} \\ \frac{p(-k_{0x})}{s(-k_{0x})} & \frac{1}{s(k_{0x})} \end{pmatrix}. \quad (79)$$

This result corresponds to the general case, $\exp\{ik_{0x}x\}$ when the absorption properties of a media are taken account. Note, that as it follows from Eq. (79) the transmission amplitudes of the left and right scattering problems are equal to each other, while the reflection amplitudes are differ both in absolute value and phase.

It is important to note that when the dielectric susceptibility $\mathcal{E}(x)$ (it corresponds to the case of an absorption absent) is a real, then $Q_1(x)$, $Q_2(x)$ (36) are real quantities as well. As it follows from Eqs. (34), (35), in this case the change of k_{0x} sign is equivalent to the action of complex conjugate: $t(-k_{0x}) = t^*(k_{0x})$ and so on. So, for the absorption absent case the transfer matrix Eq. (79) takes the following well known form:

$$\hat{M}(k_{0x}) = \begin{pmatrix} \frac{1}{t^*(k_{0x})} & \frac{-r^*(k_{0x})}{t^*(k_{0x})} \\ \frac{-r(k_{0x})}{t(k_{0x})} & \frac{1}{t(k_{0x})} \end{pmatrix} = \begin{pmatrix} \frac{1}{s^*(k_{0x})} & \frac{p(k_{0x})}{s(k_{0x})} \\ \frac{p^*(k_{0x})}{s^*(k_{0x})} & \frac{1}{s(k_{0x})} \end{pmatrix}, \quad (80)$$

and Eqs. (76), (77) will present the law of conversation of a probability density flux:

$$t^*(k_{0x})t(k_{0x}) + r^*(k_{0x})r(k_{0x}) = s^*(k_{0x})s(k_{0x}) + p^*(k_{0x})p(k_{0x}) = 1.$$

As it follows from Eq. (80) for the absorption absent case the reflection amplitudes of the left and right scattering problems have a same module, however a different phase: $r(k_{0x}) = -p^*(k_{0x})t(k_{0x})/t^*(k_{0x})$.

5.A Cauchy problem for a wave propagation description

Let us consider a solution of the wave equation (21), which has the following asymptotic behavior:

$$\tilde{H}(k_{0x}, x) = \begin{cases} \exp\{-ik_{0x}x\}, & x < x_1, \\ \frac{1}{s(k_{0x})} \exp\{-ik_{0x}x\} + \frac{p(k_{0x})}{s(k_{0x})} \exp\{ik_{0x}x\}, & x > x_2. \end{cases} \quad (81)$$

It is easy to check that this solution is obtained from the solution of the right scattering problem (see Eq.(46)) by division on the transmission amplitude;

$$\tilde{H}(k_{0x}, x) = \frac{H_{right}(k_{0x}, x)}{s(k_{0x})}. \quad (82)$$

As it follows from Eq.(69) the function $\tilde{H}(-k_{0x}, x)$ is a solution of the wave equation as well. It accordance with Eq. (81) it has the following asymptotic behavior:

$$\tilde{H}(-k_{0x}, x) = \begin{cases} \exp\{ik_{0x}x\}, & x < x_1, \\ \frac{1}{s(-k_{0x})} \exp\{ik_{0x}x\} - \frac{r(k_{0x})}{t(k_{0x})} \exp\{-ik_{0x}x\}, & x > x_2. \end{cases} \quad (83)$$

where we used the equality (see Eqs. (78) and (84))

$$\frac{p(-k_{0x})}{s(-k_{0x})} = -\frac{r(k_{0x})}{s(k_{0x})} \text{ and } t(k_{0x}) = s(k_{0x}).$$

By using Eq. (40) which is the asymptotic form of the left solution and Eq. (81), Eq. (85) one can write

$$H_{left}(k_{0x}, x) = \tilde{H}(-k_{0x}, x) + r(k_{0x})\tilde{H}(k_{0x}, x), \quad (86)$$

which by means of Eq. (82) can be presented in the form of

$$H_{left}(k_{0x}, x) = \frac{1}{s(-k_{0x})} H_{right}(-k_{0x}, x) + \frac{r(k_{0x})}{s(k_{0x})} H_{right}(k_{0x}, x), \quad (87)$$

Let us denote the solutions $\tilde{H}(k_{0x}, x)$, $\tilde{H}(-k_{0x}, x)$ as

$$\tilde{H}(-k_{0x}, x) = H_1(x), \quad \tilde{H}(k_{0x}, x) = H_2(x),$$

and introduce them by means of counterpropagating waves (see Eq. (30));

$$H_1(x) = a_1(x) \exp\{ik_{0x}x\} + b_1(x) \exp\{-ik_{0x}x\}, \quad (88)$$

$$H_2(x) = a_2(x) \exp\{ik_{0x}x\} + b_2(x) \exp\{-ik_{0x}x\}, \quad (89)$$

where we made the following notations (see Eq.(62)):

$$a_1(x) = a(-k_{0x}, x), \quad b_1(x) = b(-k_{0x}, x) \quad \text{and} \quad a_2(x) = a(k_{0x}, x), \quad b_2(x) = b(k_{0x}, x).$$

The couples of the functions $a_1(x)$, $b_1(x)$ and $a_2(x)$, $b_2(x)$ will satisfy to the same set of Eqs. (34), (35), but the initial conditions for them are differ: $a_1(x_1) = 1$, $b_1(x_1) = 0$ and $a_2(x_1) = 0$, $b_2(x_1) = 1$. The magnitudes of the functions $a_1(x)$, $b_1(x)$ and $a_2(x)$, $b_2(x)$ in the point $x = x_2$ will define the scattering amplitudes for the considered fields. Indeed, considering into Eqs. (88), (89) $x = x_2$ and comparing them with asymptotic behaviors Eqs. (81), (83), one can write:

$$a_2(x_2) = \frac{p(k_{0x})}{s(k_{0x})}, \quad b_2(x_2) = \frac{1}{s(k_{0x})} \quad \text{and} \quad a_1(x_2) = \frac{1}{s(-k_{0x})}, \quad b_1(x_2) = -\frac{r(k_{0x})}{t(k_{0x})}. \quad (90)$$

Using the given four equalities one will get

$$t(k_{0x}) = s(k_{0x}) = \frac{1}{b_2(x_2)}, \quad p(k_{0x}) = \frac{a_2(x_2)}{b_2(x_2)}, \quad r(k_{0x}) = -\frac{b_1(x_2)}{b_2(x_2)}. \quad (91)$$

The obtained result allows to formulate the problem of a wave field determination in the form of a Cauchy type problem for a set differential equations;

$$H_{left}(k_{0x}, x) = \left[a_1(x) - \frac{b_1(x_2)}{b_2(x_2)} a_2(x) \right] \exp\{ik_{0x}x\} + \left[b_1(x) - \frac{b_1(x_2)}{b_2(x_2)} b_2(x) \right] \exp\{-ik_{0x}x\}, \quad (92)$$

where couples of the functions $a_1(x)$, $b_1(x)$ and $a_2(x)$, $b_2(x)$ are solutions of Eqs. (34), (35);

$$\frac{da_{1,2}(x)}{dx} = -\frac{iQ_1(x)}{2k_{0x}} a_{1,2}(x) - \frac{iQ_2(x)}{2k_{0x}} b_{1,2}(x) \exp\{-i2k_{0x}x\}, \quad (93)$$

$$\frac{db_{1,2}(x)}{dx} = \frac{iQ_1(x)}{2k_{0x}} b_{1,2}(x) + \frac{iQ_2(x)}{2k_{0x}} a_{1,2}(x) \exp\{i2k_{0x}x\}, \quad (94)$$

which should satisfy to the following initial conditions:

$$a_1(x_1)=1, \quad b_1(x_1)=0 \quad \text{and} \quad a_2(x_1)=0, \quad b_2(x_1)=1. \quad (95)$$

Note that in accordance with Eq. (57), which takes place for any couple of solutions, and Eq. (95)

$$a_1(x_1)b_2(x_1) - b_1(x_1)a_2(x_1) = a_1(x_2)b_2(x_2) - b_1(x_2)a_2(x_2) = 1. \quad (96)$$

It is interesting to check the magnitudes of expressions, which are concluded in the rectangular brackets of Eq. (92), in the border points of a slab $x = x_1$ and $x = x_2$. Comparing Eq. (92) with Eq.(40), one can prove that the following four equalities should take place:

$$a_1(x_1) - \frac{b_1(x_2)}{b_2(x_2)} a_2(x_1) = 1, \quad (97)$$

$$a_1(x_2) - \frac{b_1(x_2)}{b_2(x_2)} a_2(x_2) = t(k_{0x}), \quad (98)$$

$$b_1(x_1) - \frac{b_1(x_2)}{b_2(x_2)} b_2(x_1) = r(k_{0x}), \quad (99)$$

$$b_1(x_2) - \frac{b_1(x_2)}{b_2(x_2)} b_2(x_2) = 0. \quad (100)$$

Indeed, Eq. (97) is directly followed from the initial condition Eq. (95). By using Eq. (96) it is easy to check that Eq. (98) coincides with the first equality of Eq. (91). The initial condition Eq. (95) provides a transformation of Eq. (99) into the second equation of Eq. (91). The last equality Eq. (100) is automatically performed.

So, we have shown that the problem of a scattering field determination can be presented in a Cauchy type problem for a set of differential equations. For the case of the left scattering problem the wave field $H_{left}(k_{0x}, x)$ Eq. (92) is presented by means of the couples of functions $a_1(x)$, $b_1(x)$ and $a_2(x)$, $b_2(x)$, which are the solutions of the set of differential equations (93), (94) with the initial conditions Eq. (95). The given formulation of the scattering problem is very useful to make numerical formulation, but it is inconvenient for analytic consideration. Even a simple case of the uniform slab, which should be solved on the base of Eq.(92) - Eq.(95), requires many calculations. As it follows from the result presented in the last section, it will be very useful to have a scattering problem formulation, which will be convenient both for a numerical calculation and for an analytical consideration as well.

Conclusion

Above we generalized the method of counterpropagating wave for a problem of Propagation of a harmonic in time and an arbitrary polarized plane electromagnetic wave through a non-regular one-dimensional media. The main idea of consideration based on the wave field presentation in the form of

$$H(x) = a(x) \exp\{ik_{0x}x\} + b(x) \exp\{-ik_{0x}x\}, \quad (101)$$

so that the derivation of a unknown function can be written

$$\frac{dH(x)}{dx} = ik_{0x} \mathcal{E}(x) (a(x) \exp\{ik_{0x}x\} - b(x) \exp\{-ik_{0x}x\}). \quad (102)$$

However the wave field presentation can be taken in the form of

$$H(x) = A(x) \exp\{i\varepsilon(x)q(x)x\} + B(x) \exp\{-i\varepsilon(x)q(x)x\}, \quad (103)$$

$$\frac{dH(x)}{dx} = iq(x)\varepsilon(x) \{A(x) \exp\{i\varepsilon(x)q(x)x\} - B(x) \exp\{-i\varepsilon(x)q(x)x\}\}. \quad (104)$$

where

$$q^2(x) = k_{0x}^2 - V(x). \quad (105)$$

On the base of Eqs. (103)-(105) a set of differential equations for the functions $A(x)$, $B(x)$ can be written as well [4]. So we had show, that the problem of a wave field determination is reduced to a set of differential equations with initial condition.

References

1. P. Yeh, *Optical Waves in Layered Media* (Wiley,1988).
2. A. Yariv and P. Yeh, *Optical Waves in Crystals* (Wiley,1984).
3. R. Bellman and G. M. Wing, *An Introduction to Invariant Imbedding* (John Wiley & Sons, Inc. 1975).
4. S. Khorasani and B. Rashidian, "Modified transfer matrix method for conducting interfaces," J. Opt. A Pure Appl. Opt. **4**, 251–256 (2002).
5. S. Khorasani and K. Mehrany, "Differential transfermatrix method for solution of one-dimensional linear nonhomogeneous optical structures," J. Opt. Soc. Am. B **20**, 91–96 (2003).
6. F.R. Di Napoli, R.L. Deavenport, Theoretical and numerical Green's function solution in a plane layered medium. J. Acoust. Soc. Am. **67**, 92–105 (1980).
7. P.G. Bergmann, The wave equation in a medium with a variable index of refraction. J. Acoust. Soc. Am. **17**, 329–333 (1946).
8. G. Barton, *Elements of Green's Functions and Propagation*. Oxford,U.K.: Clarendon Press, 1989.
9. R. P. Feynman and A. R. Hibbs, *Quantum Mechanics and Path Integrals*. New York: McGraw Hill, 1965.
10. A. I. Baz', Ya. B. Zel'dovich, A. M. Perelomov, *Scattering Reactions and Decay in Nonrelativistic Quantum Mechanics*. Jerusalem: Israel Program for Scientific Translations, 1969.

11. R. G. Newton, *Scattering Theory of Waves and Particles*. New York: McGraw-Hill, 1965.
12. S. He, S. Strom, V. H. Weston, *Time domain wave splitting and inverse problems*. Oxford: Oxford University Press, 1998.
13. S. Menon, Q. Su, R. Grobe, Iterative approach to Maxwell equations for dielectric media of spatially varying refractive index. *Phys. Rev. E* **67**, 046619 1-9 (2003).
14. S. Chandrasekhar, *Radiative Transfer*. New York: Dover Publ., 1960.
15. M.Lax, Multiple scattering of waves. *Rev. Mod. Phys.* **23**, 287, 1951.
16. A.Zh. Khachatryan, Propagation of electromagnetic wave through an one-dimensional layer with an arbitrary complex refractive index. *Electromagnetic waves and electronic systems* **15**, 49-65, 2010.
17. D.M. Sedrakian, A.Z. Khachatryan, On the problem of an electron scattering in an arbitrary one-dimensional potential field. *Phys. Lett. A* 265, 294, 2000.
18. D.M. Sedrakian, A.H. Gevorgyan, A.Z. Khachatryan Reflection of a plane electromagnetic wave obliquely incident on a one-dimensional isotropic dielectric medium with an arbitrary refractive index. *Opt. Comm.* **192**, 135, 2001.

On the measurements of UHF antenna parameters by two antennas technique

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The problem of determination of meter waveband antenna parameters from surface measurements at the presence of reflection from the Earth is discussed. The traditional measurement scheme with two antennas is considered for ascertain the degree of the Earth influence on radiation pattern and gain of antenna under test. The distance between antennas is more than the minimum distance of its far-field zones. Taking account of presence of a wave reflected from the Earth an expression for antenna gain is obtained at ignoring the approximation, at which the heights of location of antennas are considerably less than distance between antennas. At determination of antenna parameters for free space from surface measurements the recommendations for reducing the influence of the wave reflected from the Earth are given.

A problem of assessment of the Earth influence on the measurements of antenna parameters was considered in many works, in particularly [1-5]. For this purpose we consider location configuration of the antennas shown in fig. 1. Electrical axes of antennas are aligned. In geometrical optics approach the direct beam {1} and the reflected from Earth beam {2} reach the receiving antenna A_2 from the radiating antenna A_1 . There are the following relations between the angles and the sizes marked in fig. 1.

$$\alpha = \arcsin \frac{h_2 - h_1}{r_0}; \quad d = r_0 \sin \alpha; \quad \psi = \arctg \frac{h_1 + h_2}{d}; \quad \varphi = \psi + \alpha; \quad \beta = \psi - \alpha \quad (1)$$

For further, let's introduce the notations and explanations:

$\dot{R} = |\dot{R}|e^{-i\theta}$ - complex reflection coefficient for the wave falling on the Earth; $G_1(\gamma)$, $G_2(\gamma)$ - gains of antennas A_1 and A_2 in a free space

respectively; γ - angle between antenna axis and the direction of the beam, which is radiated or received by the antenna; α - angle between the horizontal and the line, connecting the axes of the antennas; ψ - grazing angle of the reflected beam {2}; P_0 - input power of the radiating antenna A_1 ; r_0 - way of the direct beam {1}; $r = r_1 + r_2$ - way of reflected beam {2}; $\Delta r = r - r_0$ - difference of their ways.

The angle γ is accepted as positive, if its reading from the axis of the antenna is anti-clockwise and negative – otherwise. For the antenna A_1 in the configuration of fig. 1 direct beam {1} is radiated in the direction $\gamma = 0$ and the reflected beam {2} is radiated in the direction $\gamma = -\varphi$. For the antenna A_2 the direct beam {1} is received in the direction $\gamma = 0$, and the reflected beam {2} is received in the direction $\gamma = \beta$.

In mentioned works, at consideration of the problem, the two conditions have used

$$r_0 \geq r_{fz}^{\min} \quad (2)$$

$$\frac{h_1}{r_0} < 1 \quad (3)$$

where $r_{fz}^{\min}$ - the maximum value from of minimal distances of far-field zones of antennas A_1 , A_2 , and from (3) follows $\beta \rightarrow 0$.

In [5] the problem have solved by the following manner. To take into account of directivity of antenna A_2 and distinguish ability of its pattern at receiving of direct and reflected waves, the function of the antenna A_2 directional pattern is expanded into series by degrees of the small parameter $h_1/r_0 < 1$ and expansion terms above two are neglected. At the heights $h_1 \approx (2.....10)m$ of the location UHF antennas above ground (to obtain significant radiation field at real parameters of the soil) from (3) follows, that the minimum distance r_0 , at which measurements should be carried out, lies in the interval $r_0 \approx (20.....100)m$

However, there is a wide class of TV UHF antennas and antenna array radiators, largest size L of which is comparable with the wavelength λ . When $L \leq \lambda/2$ the minimal distance r_0 of the far-field zone of such antennas starts away from a distance λ , i.e. $r_0 \approx \lambda$, and

when $\lambda/2 \prec L \approx \lambda$ the values r_0 are $r_0 \approx 2L$ [6]. For the meter waveband antennas with operating wavelengths $\lambda = (1...6)m$ the minimal distance r_0 will be in the range $(2...12)m$. In general, with

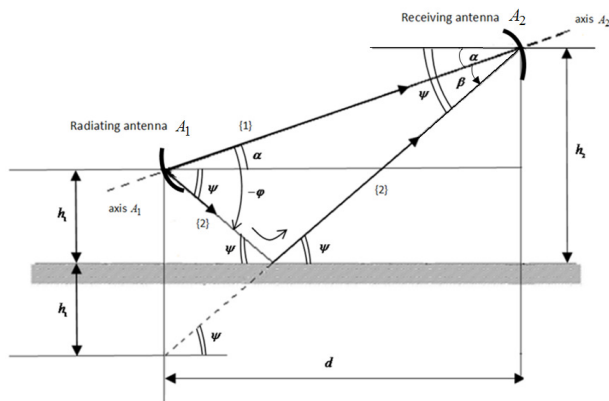


Fig. 1. Location configuration of antennas

viewpoints convenience of measurement carrying out, choices of measurement place, power and sensitivity of measurement devices more preferable to carry out the measurements on the small distances. But in the case of such measurements the approximation (3) is violated and the angle β between the direct and the reflected beams received by the antenna A_2 , will considerably differ from zero. In this case we will have a significant difference in the conditions of receiving of these beams by the directional antenna A_2 , and the approach, proposed in [5] for the solving of the problem, would be impossible.

The assigned task requires rigorous assessment of whatever value in the differences of receiving of the direct and reflected waves irrespective of ratio $h_1 / r_0 \ll 1$. This means that at correct solving of task we must keep the condition (2), but we would not limited by the condition (3). The following result of assigned task solution is obtained at meet of condition (2) only.

At the receiving antenna A_2 for the field intensity $\dot{E}_{dir}(\gamma)$ of the direct wave, excited by the radiating antenna A_1 in the direction $\gamma = 0$, we have according to [2] - [4]

$$\dot{E}_{dir}(0) \equiv \dot{E}_{dir}(0) | e^{i(\alpha x - k r_0)} = E^0(0) e^{i(\alpha x - k r_0 + \delta_{dir})}; \quad E^0(0) = \frac{1}{r_0} \sqrt{60 P_0 G_1(0)} \quad (4)$$

and for the field intensity $\dot{E}_{ref}(\gamma)$ of the reflected wave, excited by the radiating antenna A_1 in the direction $\gamma = -\varphi$, we have

$$\dot{E}_{ref}(-\varphi) \equiv \dot{E}_{ref}(-\varphi) | e^{i(\alpha x - k r - \vartheta)} = E^0(-\varphi) | \dot{R} | e^{i(\alpha x - k r - \vartheta + \delta_{ref})}; \quad E^0(-\varphi) = \frac{1}{r} \sqrt{60 P_0 G_1(-\varphi)} \quad (5)$$

The Pointing vectors corresponding to the direct and reflected waves are

$$\tilde{\Pi}_{dir} = \frac{|\dot{E}_{dir}(0)|^2}{240\pi} \quad (6)$$

$$\tilde{\Pi}_{ref} = \frac{|\dot{E}_{ref}(-\varphi)|^2}{240\pi} \quad (7)$$

To the output of the antenna A_2 these waves reach with a phase shift $\Delta\xi$, which is equal

$$\Delta\xi = \Delta\phi_{rad} + k\Delta r + \theta + \Delta\phi_{rec} \quad (8)$$

where $\Delta\phi_{rad}$ - phase shift, caused by the difference in the radiating directions of the direct and reflected waves by the antenna A_1 , $\Delta\phi_{rec}$ - phase shift, caused by the difference in the receiving directions of the direct and reflected waves by the antenna A_2 , shift $k\Delta r$ - due to the path difference Δr of these waves, and shift θ - caused by reflection from the Earth one of the waves. In consideration of the fact of coherent reception of the direct and reflected waves by the antenna A_2 , for the output power P_{A2} of the antenna A_2 we obtain following expression

$$P_{A2} = \left(\frac{\lambda}{4\pi r_0} \right)^2 P_0 G_2(0) G_1(0) \left[1 + 2 \frac{r_0}{r} \sqrt{K_1 K_2} |\dot{R}| \cos \Delta\xi + \frac{r_0^2}{r^2} K_1 K_2 |\dot{R}|^2 \right] \quad (9)$$

where

$$K_1 = \frac{G_1(-\varphi)}{G_1(0)}, \quad K_2 = \frac{G_2(\beta)}{G_2(0)} \quad (10)$$

Expression (9) satisfies to reciprocity principle, i.e. if the input

power of the antenna A_2 equals P_0 , then the output power of the antenna A_1 will $P_{A1} = P_{A2}$

Let's denote

$$G_1^{equiv} = G_1(0) \left[1 + 2 \frac{r_0}{r} \sqrt{K_1 K_2} |\dot{R}| \cos \Delta\xi + \frac{r_0^2}{r^2} K_1 K_2 |\dot{R}|^2 \right] \quad (11)$$

then

$$G_1^{equiv} = \left(\frac{4\pi r_0}{\lambda} \right)^2 \frac{P_{A2}}{P_0} \frac{1}{G_2(0)} \quad (12)$$

Value G_1^{equiv} is the gain of the antenna A_1 in direction of its axis ($\gamma=0$). It depends on the influence of the reflection from the Earth, directional properties of both antennas and the nature of the soil.

Let's assume that we change the height h_2 of antenna A_2 , but every time its axis is directed to the center of the antenna A_1 , which axis is changeless in the space. As a result of changing the height h_2 , a deviation of the direct beam radiated by the antenna A_1 occurs from an axis of antenna on some γ angle depending on value h_2 . In this case the values G_1 , K_1 , K_2 , $\dot{R}$ and $\Delta\phi_{rad}$, $\Delta\phi_{rec}$, $k\Delta r$, θ (i.e. and $\Delta\xi$) will change, because they all are functions of γ . The change of $\Delta\xi$ will lead to a periodic change of the second term in brackets (proportional to $\cos \Delta\xi$) of the expression (11). This periodic change reflects the known fact of antenna pattern petal shape depending on elevation angle γ and due to the influence of reflection from the Earth.

For great r_0 we have $h_1/r_0 \ll 1$ and $\beta \rightarrow 0$. The corresponding situation is shown in fig. 2. As far as $\beta \approx 0$, the direct and reflected beams are parallel and $r_0 \approx r$, $\Delta r \approx 2h_1 \sin \alpha$, $\varphi = 2\alpha = 2\psi$, $K_2 \approx 1$. Then, from (11) we obtain

$$G_1^{equiv}(0) = G_1(0) \times \left\{ 1 + 2\sqrt{K_1} |\dot{R}| \cos \Delta\xi + K_1 |\dot{R}|^2 \right\} \quad (13)$$

The analysis of expression (11) let us to determine the possible and suitable location configuration of antennas, at which the deviation of antenna gain G_1^{equiv} , measured in the presence of the reflected beam, is possibly minimized relative to the same antenna gain $G_1(0)$, but measured in free-space conditions.

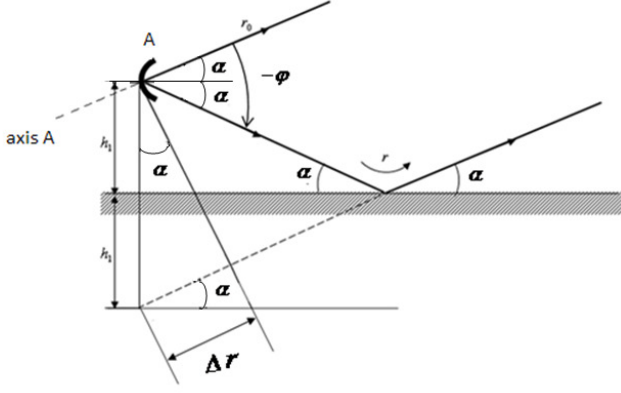


Fig.2. Situation for great r_0

With this goal we consider the measurement of antenna gain by the technique with two identical antennas, when $G_1 = G_2 = G$ [6],[7]. The expression (9) in this case has the view

$$P_A = \left(\frac{\lambda}{4\pi r_0} \right)^2 P_0 G^2(0) \left\{ 1 + 2 \frac{r_0}{r} \sqrt{K_1 K_2} |\dot{R}| \cos \Delta \xi + \frac{r_0^2}{r^2} K_1 K_2 |\dot{R}|^2 \right\} \quad (14)$$

Introducing the notation

$$D = \sqrt{1 + 2 \frac{r_0}{r} \sqrt{K_1 K_2} |\dot{R}| \cos \Delta \xi + \frac{r_0^2}{r^2} K_1 K_2 |\dot{R}|^2} \quad (15)$$

we have:

$$P_A = \left(\frac{\lambda}{4\pi r_0} \right)^2 P_0 G^2(0) D^2 = \left(\frac{\lambda}{4\pi r_0} \right)^2 P_0 (G^{equiv})^2 \quad (16)$$

where

$$G^{equiv} = \left(\frac{4\pi r_0}{\lambda} \right) \sqrt{\frac{P_A}{P_0}} \quad (17)$$

$$G^{equiv} = G(0) D \quad (18)$$

The calculated value G^{equiv} , defined by (17) with using measured ratio P_A / P_0 , differs from sought free-space gain $G(0)$, according to

(18), by a multiplier D . Note that the multiplier D contains the quantities K_1 and K_2 , which are equal $K_1 = G(-\varphi)/G(0)$, $K_2 = G(\beta)/G(0)$ according to (10). Thus, it is not necessary to know $G(0)$ to evaluate the difference $\Delta G(dB) = G^{equiv}(dB) - G(0)(dB) = D(dB)$ between calculated G^{equiv} and the sought $G(0)$, but it is necessary to know the ratios $G(-\varphi)/G(0)$ and $G(\beta)/G(0)$. It can be obtained on the basis of a priori approximate data from the results of theoretical and experimental characters or the results of antenna simulation.

As far as K_1 , K_2 and $|\dot{R}|$ are smaller than one, the main contribution, caused by the reflection in the difference $\Delta G(dB)$, gives the second term of the multiplier D . The simple method for reducing of value of multiplier D is the choosing of such location configuration of antennas at which $r = r_1 + r_2 \geq 2r_0$, and the product $K_1 K_2$ is small. For example, if $r_0/r \leq 0,5$ $K_1 + K_2 = 10dB$ ($K_1 K_2 = 0,1$) we have $\Delta G \leq 0,6dB$. For $r_0/r \leq 0,35$, $K_1 + K_2 = 15dB$ ($K_1 K_2 = 0,03$), $\Delta G < 0,25dB$ and in the case $K_1 + K_2 = 20dB$ ($K_1 K_2 = 0,01$), $\Delta G < 0,15dB$.

References

- [1] Р. Баумен, Калибровка образцовых антенн на частотах выше 1 ГГц, ТИИЭР, (1967), **55**, №6, 272.
- [2] Распространение радиоволн. Под ред. О.И. Яковлева (М., ООО ЛЕНАНД, 2009)
- [3] A.S. Saakian, Radio wave propagation fundamentals (Artech House, 2011)
- [4] Г.А. Ерохин и др., Антенно-фидерные устройства и распространение радиоволн (М., Радио и связь, 2007)
- [5] А.Р. Вольперт, Об измерении диаграмм антенн в условиях влияния отраженных от земной поверхности волн, Радиотехника, (1978), **33**, №6, 71.
- [6] А.З. Фрадин, Е.В. Рыжков, Измерение параметров антенно-фидерных устройств (М.,Связь, 1972)
- [7] Методы измерения характеристик антенн СВЧ. Под ред. Н.М. Цейтлина (М., Радио и связь, 1985)

The measurements of parameters of UHF antenna array radiators

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The technique for determination of UHF antenna pattern and gain in free space from near-surface measurements is proposed. It is based on measurement technique with two identical antennas. For measurements of mentioned parameters of UHF ring antenna the appropriate location configuration of antennas are chosen taking into account antenna far-field zone distance and results of its preliminary simulation. The chosen location configuration of antennas allow to reduce of influence of wave, which is reflected by the Earth and disturb measurements. The value of ring antenna gain is well correlated with result of antenna simulation, which performed by program HFSS.

The ring antenna, having a relatively small longitudinal dimension and a sufficient antenna gain is proposed as a possible radiator of UHF antenna array. The other parameters of the antenna are: the relative operating frequency bandwidth is 25%, VSWR < 2 in the whole frequency bandwidth, and polarization is horizontal. General view of antenna is shown in Fig.1.

For UHF antenna array modeling it is necessary to know parameters of radiators forming the array. The problem was set in determination of above mentioned parameters, which can be identified with the parameters of the radiator in free space.

With this goal the measurement procedure, described in [1], is used, on basis of which the configuration of two identical antennas location, as shown in Fig.2 was chosen.

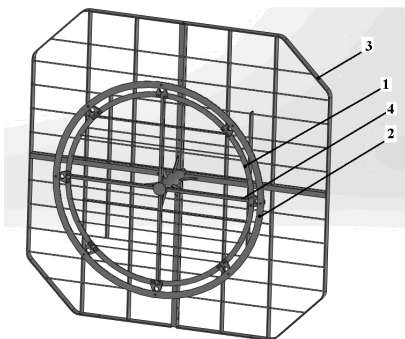


Fig.1. Ring antenna: 1 – metallic disk, 2 – ring radiator, 3 – reflector, 4 – coaxial feeder cable

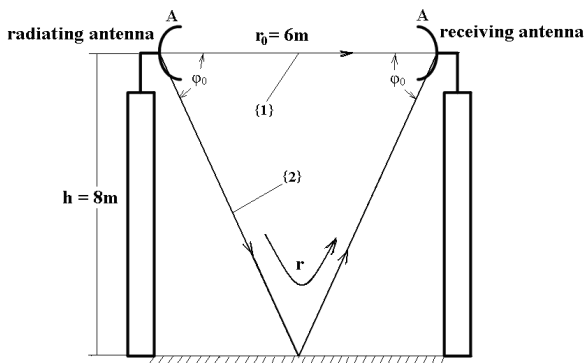


Fig.2. Location configuration of two identical antennas
{1} – direct beam, {2} – reflected beam

At perfect matching of the identical receiving antenna and radiating antenna, to which input is fed the power P_0 , the expression for the output power P^{rec} of receiving identical antenna will be [1].

$$P^{rec} = \left(\frac{\lambda}{4\pi r} \right)^2 P_0 G^2(0) \left[1 + 2 \frac{r_0}{r} K |\dot{R}| \cos \Delta \xi + \frac{r_0^2}{r^2} K^2 |\dot{R}|^2 \right] \quad (1)$$

where:

$G(0)$ – required gain of the antenna in free space;

$\dot{R} = |\dot{R}|e^{-i\theta}$ – reflectivity from the Earth; the phase θ equals 180° for horizontal polarization;

$\Delta\xi$ – phase shift between direct and reflected beams;

$$K = \frac{G(\varphi_0)}{G(0)} = \frac{G(70^\circ)}{G(0)} = -9dB, \quad (K = 0,126) \text{ – according to the results of}$$

simulation and preliminary assessed measurements of antenna directional pattern;

$\frac{r_0}{r}$ – ratio of the ways of direct and reflected beams.

In chosen configuration $\varphi_0 = 70^\circ$, $\frac{r_0}{r} \approx 0,35$, $\frac{r_0}{r_{ffz \min}} \approx 2,5$, and

$r_{ffz \min}$ – minimum distance of the antenna far-field zone.

Let's denote

$$G_{equiv} = G(0)A_{\max} \quad (2)$$

where

$$A_{\max} = \sqrt{1 + 2\frac{r_0}{r}K + \frac{r_0^2}{r^2}K^2} = 1 + \frac{r_0}{r}K \quad (3)$$

Expression of $A_{\max}$ have determined as the maximum value of the factor in square brackets of (1), when $\cos \Delta\xi = 1$ and $|\dot{R}| = 1$

Thus, the value $G_{equiv}(0)$ of the gain factor (obtained from the measurements by means of the proposed antenna location configuration) will be differ from the required $G(0)$ in the worst case by the $A_{\max}$ factor.

For the chosen configuration we have $\frac{r_0}{r} \approx 0,35$, $K = -9dB$ ($K = 0,126$) and

$$A_{\max} = 1,044 \text{ or } A_{\max} = 0,187dB < 0,2dB.$$

Antenna pattern and gain were measured by the block diagram shown in Fig.3a and 3b [2].

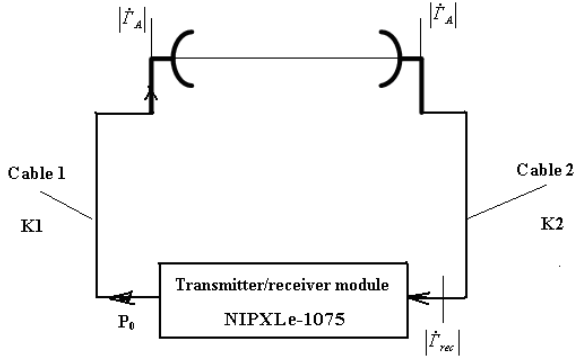


Fig.3a. Block diagram of gain measurement

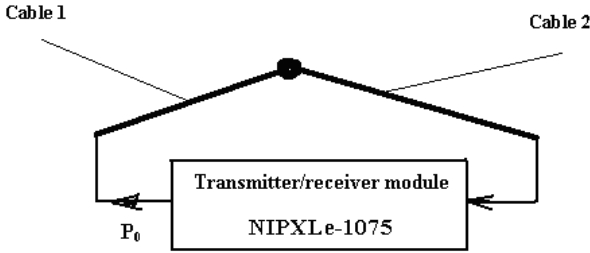


Fig.3b. Block diagram of received signal calibration

At reflection coefficients of the antennas $|\dot{\Gamma}_A|$, receiver $|\dot{\Gamma}_{rec}|$, transmission coefficients of cables 1,2 and receiver accordingly K_1 , K_2 and K_{rec} , the expression for the P_a^{rec} at the scheme 3a will be:

$$P_a^{rec} = \left(\frac{\lambda}{4\pi r_0} \right)^2 P_0 G_{equiv}^2(0) \left[1 - |\dot{\Gamma}_A|^2 \right]^2 \left[1 - |\dot{\Gamma}_{rec}|^2 \right] K_1 K_2 K_{rec} \quad (4)$$

At signal calibration by the scheme 3b, we have

$$P_b^{rec} = P_0 \left[1 - |\dot{\Gamma}_{rec}|^2 \right] K_1 K_2 K_{rec} \quad (5)$$

From the relations (4) and (5) we have

$$G_{eqv}(0) = \left(\frac{4\pi r_0}{\lambda} \right) \frac{1}{1 - |\dot{r}_A|^2} \sqrt{\frac{P_a^{rec}}{P_b^{rec}}} \quad (6)$$

The view of antenna location configuration at above mentioned measurements is shown in Fig.4.



Fig.4. View of location of antennas

The measured gain was $G_{eqv}(0) = 8,83dB$, and gain, obtained during the simulation by program HFSS, was $G = 9dB$. Note, that the error of $\Delta G(\Delta r_0)$ in gain due to an error of the distance between the antennas $\Delta r_0 = 20sm = 0,2m$ at $r_0 = 6m$ is $\Delta G(\Delta r_0) = 0,28dB$.

The covering of land area between antennas by thick metal mesh does not lead to the significant change of the received power level.

For the formation of reflected beam should be sufficient area of the earth surface about spot of the first Fresnel zone [3]. This condition is satisfied, since the radius of the first zone is about 3m for a given configuration.

The measured antenna pattern is shown in Fig.5. Measurement readouts were carried out through the step $0,5^\circ$ of antenna rotation.

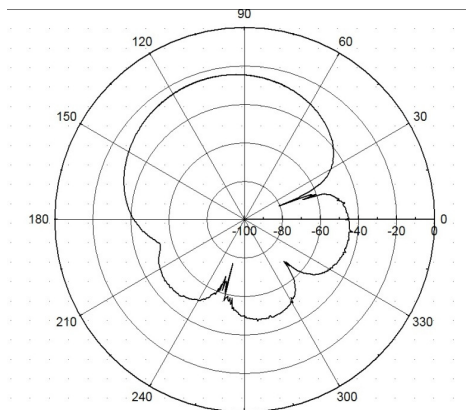


Fig.5. The measured directional pattern

Half-power width of pattern – 62^0 , back lobe level – (-20dB),
cross - polarization level – (-20dB).

References

- [1] V.H. Avetisyan, A.A. Sargsyan, On the measurements of UHF antenna parameters by two antenna technique (Present proceeding).
- [2] Методы измерения характеристик антенн СВЧ, Под ред. Н.М. Цейтлина (М. Радио и связь, 1985).
- [3] М.П. Долуханов, Распространение радиоволн (М.Связь, 1972).

The problem of automation of electromagnetic compatibility calculations for microwave-link equipment

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In this report we have discussed nowadays problem of electromagnetic compatibility (EMC) associated with modern trends of saturation all sectors of life with different radio electronic equipments (REE). In this context we determine the usefulness of automation of process identifying EMC. We have also proposed the stages of automation of identification EMC for radio relay stations (RRS). In the process of automation “Google Earth” program is used which is good support for creation more affordable final product. All process is dividing into five stages and based in recommendations of International Telecommunication Union (ITU) for different accessions. Also the nowadays stage of solving the problem of automation of EMC for RRS is submitted.

The stages of the automation problem solving for calculation of electromagnetic compatibility of radio electronic equipments (to be more exact radio relay lines of communication) has been identified. The process of automation and calculation algorithms involves the “Google Earth” program use as the source of the original data by geographic location of relay stations.

The continuous growth of human wants in technology and communication facilities, which would meet the demanding requirements of present level of development, lead to an increase of REE number per unit area and accordingly the question of its simultaneous operation in a single electromagnetic environment (EME) with the required quality, is raised. For it satisfaction is necessary the EMC of REE operating in a given EME. For determination of EMC fact it is necessary its calculation. Modern trends of the REE number increasing results to possible interference growth for the considered lines of communication, and therefore, the volume of EMC calculation is increased.

For the EMC calculation time and cost reducing, the task of developing and realizing of project of the given calculations automation RRS by the personal computer (PC) is assigned.

Automation task is divided into five stages.

The first stage

- Creating and input of REE database in PC, on the basis of a given EME. The database should contain the geographic location coordinates of REE (latitude, longitude, altitude above the ocean level), radio equipment type, parameters of transceiver antennas; data transfer rate or the operating frequency band, receiving and transmitting frequencies, radiation power.

- Selection of interacting REE from the created database by the criteria of the transmitting and receiving REE of radio relay stations operating frequency band overlapping and its mutual removal.

The second stage

- Automatic acquisition of path profiles between the selected interacting REE points, taking into account the roundness of the Earth and saving the images of these profiles.

- Defining the path gaps and its classification according to the given types: paths with open gaps, semi open gaps or closed gaps [1-3].

- Creation (for each receiving REE a database as a table) of interacting REEs, correspondingly with open, semi open and closed paths gaps.

- The input power level calculation (for each receiving REE) of the interfering signal from the transmitting REEs from the database of interacting REEs which have open track gaps between the considered receiving REE[2,3].

The third stage

- obstructions joining on the paths, according to the mutual positions criterion with semi open path gap between the considered receiving REEs and interfering REEs from the database of interacting REEs with semi open gaps [1].

- Approximation of the receiving obstruction types (tapered, cylindrical, sphere) in semi open paths.

- obstructions joining on the paths, according to the mutual posi-

tions criterion with closed path gaps between the considered receiving REEs and interfering REEs from the database of interacting REEs with closed gaps [1].

- Approximation of the receiving obstruction types (tapered, cylindrical, sphere) in closed paths.

The fourth stage

- The input power level calculation (for each receiving REE) of the interfering signal from the transmitting REEs from the database of interacting REEs which have semi open path gaps between the considered receiving REE.

- The input power level calculation (for each receiving REE) of the interfering signal from the transmitting REEs from the database of interacting REEs which have closed path gaps between the considered receiving REE.

- The calculation of the total interfering signal power level on the input of considered receiving REE-the receptor of interference.

- Determination of the EMC fact of the considered transceiver REE of each radio relay line (RRL) on the basis of EMC criteria of useful and total interfering signal comparison.

The fifth stage

- The choice of path profile between the first and the second stations of considered RRL and its printing from the received path profiles archive.

- The choice of profiles between the first station of the considered RRL and interacting REE and its printing, from received path profiles archive.

- The choice of profiles between the second station of the considered RRL and interacting REE and its printing, from received path profiles archive.

- The map printing of the mutually positioned REE, given in EME.

- Creation of the data file from automated calculation for preparation of the EMC project calculation considered by RRL.

At the given task solving, the usage of different methods that were recommended by the ITU and in [1- 3] is specified. The methods consist of chapters, containing the obstruction type observation, REE

technical characteristics that influence to EMC; database of necessary antennas characteristics, which are demanded for estimating of the obstruction making for other services. The methods of patterning and analyzing of profiles as well as the methods of path classification are recommended. The method of calculating the interfering signal power for different mechanisms of propagation of radio waves (in conditions of direct vision and diffraction) is offered. At radio waves diffraction conditions, i.e. on semi closed and closed paths, the various criteria of approximations of few obstructions by one is recommended.

At present, the program which uses the ITU recommendations can be able operate with the path profiles relief data, deduced by the “Google Earth” program is created. The result will be the image of path relief profile taking into consideration the roundness of the Earth. At thus on the visualization line between antennas the calculated radius of Fresnel free zone, depending on distance between the considered coordinates of RRL stations is superimposed. As a result, the significant obstruction on the path, value of gap over the obstruction and path type are determined, taking into account the refraction in atmosphere.

The program for calculating the signal in open gap paths is developed too. The program is carried out calculation of the total signal attenuation, in the conditions of signal weakening in an open area, absorbing in gases, in water stems of atmosphere and in the rain, taking into consideration the data about climate zone in which REE is located. The Climate zone and its coordinates (probable humidity and probable rain rate) are determined due to the ITU recommendations. The information existence about main lobes directions, the side lobe levels of interacting REE antennas and the power of the emitted signal are specified. This information is input from the EME database for considered region.

References

- [1] Василенко Г.О., Подманков И.И., Антипенко В.А., Альтер Л.Ш., Мельничук В.И. (ЛЮНИИР, СПб, 2005).
- [2] Бородич С.В. (М.: Радио и связь, 1990).
- [3] Бузов А.Л., Быховский М.А., Васехо Н.В., Волкова Ю.В., Жильцов А.У., Иванова Т.В., Носов В.И., Севостьянов С.В., Сорокин А.С., Сорокин Г.И. (М.: Эко-Трендз, 2006).

Distant infrared spectral analysis of industrial hot gas ejection in atmosphere

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Passive Infrared (IR) Spectral-radiometry of gases in the atmosphere is extremely important today, when pollution of the environment by natural ejections and those produced by human activity is growing very high. Particularly, spectral analysis of hot gas ejections i.e. combustion products from industrial plants is an essential part of ecological monitoring of the atmosphere.

In this report we present the results of IR spectral analysis of hot gas ejections from industrial plants in the spectral range from 2.5 to 5.5 μ m, at a distance of 3000m. The obtained with a hydrocarbon gas group, SO₂, N₂O, CO and CO₂ gases, as well as H₂O vapor. Relative content of ejected gases (to CO-CO₂ group) per unit time was evaluated by means of an integral intensity ratio for each gas.

Distant IR Spectral analysis of hot gas ejections (both industrial firms, and various vehicles) have huge value, in particular at ecological monitoring of an environment.

Introduction

The study of gaseous components in the atmosphere plays a significant role in the sphere of ecological researches. One of main tasks of environmental control is the spectral study of chemical composition of atmosphere pollution, as well as analysis of gaseous outbursts of either industrial processes or ground transport. Important value has also distant measurements of radiation temperatures of point and extended sources of thermal radiation in an industry and in atmosphere. Therefore, the increasing applications IR Spectral Radiometry in many areas of researches have made necessary development and manufactures simples in operation, operating in application, cheap and sending IR Spectral Radiometer. Just such Spectral Radiometer was developed by our group, to which description and some researches results presentation is devoted this paper. Passive IR Spectrometry of gases in the atmosphere is extremely important today, when pollution of the

environment by human activity is growing very high. Particularly, spectral analysis of hot-gas ejections in an essential part of ecological monitoring of the atmosphere. No less significant role has the knowledge of atmospheric IR spectral transparency, which is giving information on "Optical weather" conditions during field tests of thermal-vision equipment and other apparatus, as well as for the study of atmospheric water vapor and carbonic gas content.

Research methods

Developed by us "Sipan-A" IR Radiometric System is functioned in two regime-in active and passive. In active regime it works as an IR Radiometer for measuring of the local gas concentrations in surrounding space, by means of measurements of integral absorption in chosen IR band (in wavelength region from 3 to 13 μm), and in passive regime it is functioned as a distant spectral analyzer of hot gas ejections in the atmosphere. The changing Spectral Radiometer's regime work carried out by substitution of the knots photo-sensors and Light filter, and including of the external Absolute Black Body Model. The passive regime IR.

Radiometer named as "USR-A" is detailed described in the work [1], there are analyzed principle of work and construction of the instrument. Structurally instrumentation "Sipan-A" consists three main units: Optic-Mechanical Unit (OMU), Electronic Control Unit (ECU) with the part of an automatized processing of the measurement results (on the personal computer), and knot of the external Absolute Black Body Model. The electrical link between them implements by means of cables. The optical scheme of the OMU is shown in a Figure 1.

The full working spectral range of the instrument (in passive regime of works) is covered with the help of two packages of changeable light filters and photo-sensors in sub-diapasons from 2.5 to 5.5 μm , and 7.9 -13.5 μm . The working spectral range of the "Sipan-A" Radiometer (in active regime of works) is covered with the help d the set narrow-band interference light filters in wavelength range from 3 to 12 μm . During exploitation OMU by means of cline directing placed on the rotary mechanism, which fastens to a horizontal platform specially

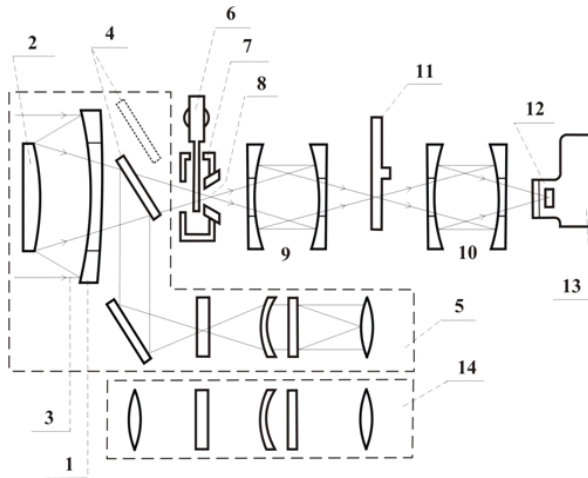


Figure 1. Optical scheme of OMU. 1-Primary mirror; 2-secondary mirror ; 3-radiation from the object; 4- removable flat mirror; 5- visier; 6-modulator; 7-basic cavity; 8- field diaphragm; 9,10- projecting objective; 11- disc with interference photo filters; 12- sensitive photodetector area;13- liquid nitrogen dewar; 14- visual tube.

made tripod. ECU structurally of desktop fulfillment. All organs of indication and handle are located on the ECU front panel. In laboratory conditions ECU is installed on desktop, and in field condition it can be installed in a body of auto-laboratory with the help of shock absorbers. The appearances of OMU and ECU "Sipan-A" Spectral Radiometer are shown in Fig.2 (A), (B). Briefly principle of "Sipan-A" operation consists of the following. In OMU the radiation flux from the researched object is going with the help of optical system (see Figure 1.) and is focused on a sensing site of the photo-sensor.

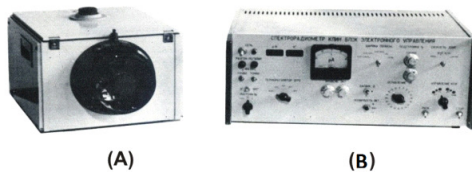


Figure2. OMA-(A) & ECU-(B) of "Sipan-A" Spectral Radiometer.

Further, the preamplifier strengthens an electrical signal and transmits to ECU. In ECU the electronic circuits strengthen, demodulated and filtered a signal from an output of the photo-sensor. In an outcome on the output of ECU, there is a signal, which amplitude is a measure of absolute radiation of the studying object.

Knowing the value of the assembled radiation power (with the help of a data preliminary carried out instrument's power calibration), spectral filtering properties of the system and degree of transform of output signal to absolute measurement of the object radiation characteristics. It is necessary to mention, that the knots of photo-sensors includes the InSb and CdHgTe could photo resistors, for the working in wavelength regions from 2.5 to 5.5 μ m and from 7.9 to 13.5 μ m respectively.

The main technical parameters of the "Sipan-A" systems are:

- Entry Objective (Cassegrainian system) Diameter - 108mm
- A focal length - 200mm
- The focused distances - from 5m to ∞
- Working spectral range -from 2.5 to 13.5 μ m
- Field of View -17ang.min.
- Distances, for spectral analysis of hot gas ejection in atmosphere, up to 5000m
- The volumetric concentration range of the measuring gas pollution in atmosphere from 0.25 LAV (Limited Admit Volume) to 10% (in volume)
- The IR Radiometric System "Sipan-A" can provide the operative ecological Monitoring of gaseous pollution in the atmosphere.

Results and discussions

The results of distant IR spectral analysis are given of hot gas ejections in the atmosphere from industrial plants. Universal Spectral Radiometer "Sipan-A" has been used for measurements in the spectral range 2.5 to 5.5 μ m, at a distance of 3000m. [2]. In this paper we present the results of our measurements, of SO₂, N₂O and hydrocarbon gas content, relative to CO and CO₂ in the 2.5 to 5.5 μ m waveband.

Smoke and flame ejected by an industrial plant pipe was taken as observation object in our experiments. IR spectral measurements were

carried out across a distance of 3000m, in summer season and in clear weather conditions. Over 20 IR spectrograms have been obtained; the averaged spectrogram is shown in relative units in Figure 3.

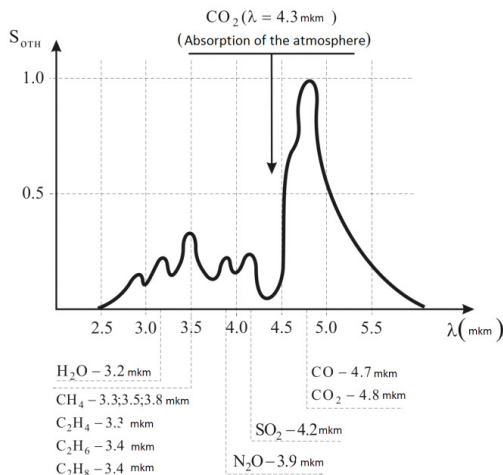


Figure 3. IR spectrogram of hot gas of industrial emissions into the atmosphere.

Note that, due to low spectral resolution ($\approx 3\%$), hydrocarbon line group's lines merge into a single band, however their integral intensity may be still compared with that for CO and CO_2 , gases, which is important for qualitative analysis in ecology. An intensive CO_2 absorption band of the atmosphere is distinctly visible in Figure 3 at a wavelength $4.3 \mu\text{m}$. [3]. CO and CO_2 , gases are known to be the major combustion component, which is reflected in our spectrogram in form of high-intensity emission band at $4.7 - 4.8 \mu\text{m}$ wavelength.

We have selected the $3 - 5 \mu\text{m}$ waveband for measurements because (1) it is one of the main "Transparency windows" in the atmosphere, and (2) because low-concentration gas pollutants, such as hydrocarbons, N_2O , SO_2 etc., possess more or less intensive oscillatory (rotation) spectra in this very range [3]. One may easily identify some groups of these molecules by their emission bands. Most distinctly visible is the hydrocarbon group, with maximum at $3.5 \mu\text{m}$, which is explained by the fact that natural gas was present in the fuel of the plant.

At a flame temperature above 2000K, the emission bands of H_2O , CO and CO_2 , gases are broadened to an extent when their spectrum in the range 3 - 5 μm becomes continuous [6]. However, at temperatures below 2000K the bands may be resolved, which was indeed observed in our experiment. Comparison of the maximum intensity wavelength in the obtained spectrum with that for the black body radiation, we have found that the effective temperature of gas ejection was in the interval 500 - 600k. Relative content of ejected gases (to CO - CO_2 , group) per unit time was evaluated by means of an integral intensity ratio for each gas. One may see that hydrocarbons content is respectively 2 and 3 times higher than SO_2 and N_2O content, and on the other hand it is 4 times less than CO - CO_2 , group. The obtained results on IR spectrometry of hot gas ejections provide important information about the extent of atmospheric pollution. Our proposed method and used equipment make possible fast determination of content for various hot-gas ejections, by passive spectral measurements in 3 - 5 μm and 8 - 14 μm wavebands.

Conclusion

In IR Radiometer Systems for decrease of overall dimensions and rise of sensitivity, usually, using entry objectives of a type Cassegrainian [4],[5] as is executed in our operation. But the matching developed by us Spectral Radiometer "Sipan-A" with closely existing analog [8] shows at least two clear advantages of the instrument, described in the present paper. With the purpose of the greatest elimination of chromatic aberration in the optical "Sipan-A" system two pairs projections mirror of objectives are applied, and in second, for the extension of instrumentation functionality's on ring disks of a mode of light filters (on region from 3 to to 5 μm , and from 8 to 14 μm) are placed half-ring variable cline light filters [8] for spectral measurements of spectrum.

Distant IR Spectral analysis of hot gas ejections (both industrial firms, and various vehicles) have huge value, in particular at ecological monitoring of an environment.

References

- [1] R.S., Asatryan, et.al. , International Journal of IR and MM Waves, **24**, 1035. (2003).
- [2] R.S. Asatryan and V.Gevorkyan. In: Proceedings of Ecological and Hydro meteorological Problems in Large Cities and Industrial Zones '2002, Sankt-Petersburg, International Conference, (In Russian 2002), 95.
- [3] V.Zuev, Atmosphere Transparency for Visible and Infrared Beams, Sov. Radio Publ. Moscow (In Russian, 1966).
- [4] V. Kozelkin, Fundamentals of infrared Engineering, Mashinostroenie Publ., Moscow (In Russian, 1974).
- [5] Yu. Abrahamyan, et.al. YSU Publ. Yerevan (In Russian 2000).
- [6] A. Kadishevich, Simplest methods of Flame Emission Capacity Measurement. Moscow, Oborongiz. (In Russian, 1954).
- [7] Dual Cannel Radiometer, US. Patent-3476914.
- [8] Branch Standard OST3-5683-84 (In Russian) (1984).

Infrared optical/electronic method of monitoring environmental CO₂ and H₂O vapors

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Knowledge of atmospheric in the infrared (IR) waveband is essential in IR remote sensing at long distances, giving information on "Optical weather" conditions during field tests of thermal-vision equipment and other apparatus, as well as for the measurement of concentrations atmospheric CO₂ and H₂O vapor on the horizontal path.

In ecological researches of a terrestrial atmosphere rather great value measurements of quantity water vapor and carbonic gas in an environment have. On the basis of the experimental data received on measurements of a spectral transparency of atmosphere in the wave lengths from 2.5 up to 5.5 μm where there are strong band of absorption water vapor (on 2.7 μm) and carbonic gas (on 4.3 μm) and with help of existing empirical dependences between a spectral transparency and quantity of absorbing molecules it is possible to determine concentration H₂O vapor and CO₂ on a site of measurements.

Introduction

In ecological researches of a terrestrial atmosphere rather great value measurements of quantity water vapor and carbonic gas in an environment have. On the basis of the experimental data received on measurements of a spectral transparency of atmosphere in the wave lengths from 2.5 up to 5.5 μm where there are strong band of absorption water vapor (on 2.7 μm) and carbonic gas (on 4.3 μm), and with the help of existing empirical dependences between a spectral transparency and quantity of absorbing molecules it is possible to determine concentration H₂O vapor and CO₂ on a site of measurements.

The present paper is devoted to representation of results and discussion of the given measurements infrared spectral transparency of the atmosphere. In the wave lengths region from 2.5 up to 5.5 μm carried out with the help developed by us universal IR Spectral Radiometer "Sipan-A", in detail described in [1].

The conventional method of measurement of IR spectral transparency of atmosphere on the large distances is reception of a spectrum known IR source with external modulation of radiation when signals from a source are small in comparison with the contribution of radiation of a background. Advantage of this method consists that drift and fluctuations of radiation of a background do not influence final results of measurements. In work [1] techniques of power graduation of instrument, and also measurement methodology of point and extended IR sources are in detail analyzed. Corresponding relation for spectral calibration characteristic ($k(\lambda)$) of the Spectral Radiometer and interest of spectral contrast representing in this case for us point IR source are received by following relation:

$$W(\lambda) = \Delta S(\lambda) \cdot \omega \cdot l^2 / k(\lambda) \cdot \tau(\lambda, l) A \quad (1)$$

Where, $\Delta S(\lambda)$ a difference between signals "source+background" and "background"; ω solid angle of the "Sipan-A" system (equal 3mrad); l - distance from a researched source up to instrument; $\tau(\lambda, l)$ spectral transparency of atmosphere on distance l ; the A - area of a radiating surface of a source.

Research methods

We have carried out experiments in the middle latitudes of the European part of Russian, in summer. 1500m long horizontal path was selected. The measuring equipment included a standard black-body source (at a temperature 1270K), a chopper positioned in front of the source and IR Spectral Radiometer "Sipan-A". Structural diagram of experiment and measuring equipment is shown in Figure 1.

In this experiment, the internal reference source of "Sipan-A" system was not used, due to existence of the external source with chopper.

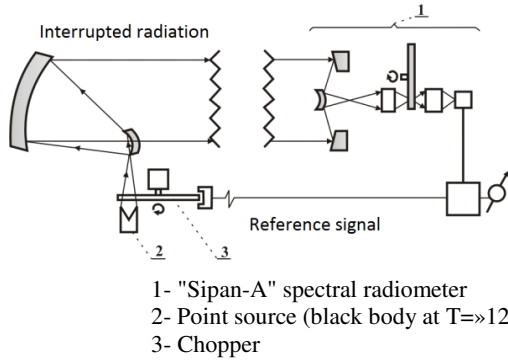


Figure 1. Structural diagram of the measurement equipment

The system was synchronized by a radio signal transmitted from the source position and repeating the modulation half-periods. As results, the synchronous detector output was equal to a difference of signals "source + background" and "background". The same spectral measurements were repeated for a short path (200m). Using the brightness spectrum relations (1) for a point IR source we have obtained the spectral transparency of atmosphere $\tau(\lambda)$ at a distance 1500m. Averaged results over 30 spectral measurements are shown in Figure 2. Solid curves in Figure 2. Correspond to calculated value [2], [8], while crosses show the results of our measurements. Simultaneously, meteorological parameter values were measured at the receiver site (humidity, pressure and temperature).

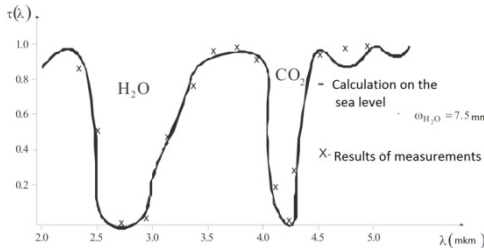


Figure 2. Spectral transparency of the atmosphere on a horizontal track length $L=1500\text{m}$.

Results and discussions

As a result of numerous experiments, Elder & Strong [3] have suggested the following empirical relation for water vapor absorption in narrow wavebands $\tau(\lambda)$, valid for horizontal paths at altitudes up to 3000m:

$$\tau(\lambda) = t_0 \cdot k_1 \cdot \lg \omega_{H_2O} \quad (2)$$

Here ω_{H_2O} is the thickness of condensed vapor layer (cm), t_0 and k_1 are empirical constants for the considered λ [3]. Using the relation (2), with constants given in, as well as the results of our measurements, we have found the condensed layer thickness $\omega_{H_2O} = 4.5$ and 9.5 mm respectively in the 1.9 - $2.71 \mu\text{m}$ and 2.7 - $4.3 \mu\text{m}$ wavebands. Thus the average value of ω_{H_2O} for a distance 1500 m makes 7.0 mm, which is close to the value 7.5 mm obtained from in-situ measurement of meteorological parameters. The solid curve $t(A_1)$ in Figure 2. was drawn for that very value (7.5 mm). Multi-year theoretical and experimental researches by the group of authors [6], [7] has shown that atmospheric transparency in general is a function of the absorbent mass (condensed layer), effective pressure and wave number: $\tau = \tau(\omega, P_E, u)$. The effective pressure $P_E = P_N + B \cdot P_a$, where P_N is the line-broadening nitrogen pressure, B is self-broadening coefficient of the absorbing gas having pressure P_a ($B=6$ or 2 respectively for H_2O and CO_2). The same authors have shown that spectral transparency is well described by the relation

$$\tau_u = \exp(-\beta_0 \cdot W^*) \quad (3)$$

Where β_0 is the absorption coefficient per unit equivalent mass W^* , which depends P_E . Papers [6], [7] indicate that this dependence may be taken in form

$$W^* = (\omega \cdot P_E^{2k})^{1/2} \quad (4)$$

Where k and 1 are parameters depending on $\omega \cdot P_E$ and u , the values of which are given in [5]. For computation using the results of our experiments (in the $4.3 \mu\text{m}$ waveband of CO_2) it was convenient to

introduce the equivalent masses separately for the center and peripheral parts of the waveband:

$$W_1^* = (\omega_{CO_2} \cdot P_E^{0.96}) \text{ (center)} \quad W_2^* = (\omega_{CO_2} \cdot P_E^{0.7})^{0.64} \text{ (periphery)} \quad (5)$$

W_1^* - was used in (2) when $W^* \leq 0.7$ and W_2^* otherwise

$$\tau_u = \exp(-\beta_{1u} \cdot W_1^*) \text{ (center)} \quad \tau_u = \exp(-\beta_{2u} \cdot W_2^*) \text{ (periphery)} \quad (6)$$

The values of coefficients $-\beta_{1u}$ and $-\beta_{2u}$ are given in [5]. Effective pressure P_E in the relations (3)-(6) were expressed in atmospheres ($\sim$ latm. at a sea level), while min centimeters. CO_2 content in the 1500m long path was found from the relations (5) and (6) for the t_0 values measured in the $4.3\mu m$ waveband. The averaged value of CO_2 content was equal $\omega_{H_2O} = 4.2cm$. Numerous researches dealt with CO_2 content in atmosphere, the results being given in the monograph [9]. Although there is a 6-fold difference between the minimum and maximum concentration of CO_2 , one may assume it constant and equal to 0.03% in volume while calculating its IR absorption (large deviations from this value are extremely rare). Note that our calculated value of CO_2 volume Concentration, equal to 0.028% stays in good agreement with the value obtained from multi- year investigations.

Conclusion

Received results of IR spectrometric measurements of atmospheric CO_2 and H_2O vapor can provide the significant information on structure of atmospheric gas pollution. The measurement methodology developed by us and the applied equipment represent an opportunity of carrying out of an operative estimation of the contents of different gases with the help passive spectrometry in the wavebands from 3 up to $5\mu m$ and from 8 up to $14\mu m$.

References

- [1] R.S., Asatryan, et.al. , International Journal of IR and MM Waves, **24**, (2003), 1035.
- [2] P. Cruse, et.al Fundamentals of Infrared Engineering, Trunsl. From Engl., Moscow. (1964)

- [3] Elder, T. & Strong, J. Journal Franklin Institute, **255** (1989), 189.
- [4] B.Golubitski, Inter- University Workshop on Spectral Transparency of Atmosphere, Tomsk (In Russian) (1965).
- [5] B. Golubitski, & N. Moskalenko, Izvestia AN SSSR, ser. Physics of Atmosphere and Ocean, **4**, (In Russian) (1968), 56.
- [6] N. Moskalenko, Izvestia AN SSSR, ser. Physics of Atmosphere and Ocean, **5**, (In Russian) (1969), 962.
- [7] N. Moskalenko, & S. Mirumyants, Izvestia AN SSSR, .ser. Physics of Atmosphere and Ocean, **6**, (1970) 1110.
- [8] A.V. Pavlov, Optical/Electronic Instruments Energy Publ, Moscow (In Russian) (1974).
- [9] V. Zuev, Atmosphere Transparency .for Visible and Infrared Beams, Sov. Radio Publ. Moscow (In Russian) (1966).

Real-time monitoring of glucose concentration by a microwave biosensor

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We used an electromagnetic microwave cavity sensor for real time measurement of the glucose concentration in biological (grape) samples. We could determine the concentration of glucose in the range of 50 - 500 mg/dl at the resonance frequency near 4.7 GHz with a bandwidth of 300 MHz. The change in the D-glucose concentration in sample brings microwave reflection coefficient S_{11} change and resonance frequency shift. The *in-vitro* results show the measured *signal-to-noise ratio* of about 44 dB, and the minimum detectable signal level of about 0.02 dB/(mg/dl).

The results clearly show the sensitivity and usefulness of this microwave sensor for these types of biological investigations. This proposed system provides a unique approach for real contactless glucose monitoring and, it may serve as a non-invasive glucometer for the calibration of different glucose levels.

1. Introduction

Glucose biosensors are major important tools for clinical monitoring and the food industry [1,2]. A sample, economical and accurate method for the measurement of glucose in a selective, sensitive and quantitative manner is urgently required. A glucose biosensor is a device that converts the biological recognition events into a signal that can be future pressed. This signal is then usually converted into an electrical signal.

Glucose biosensor can take many forms, relying on electro-chemical, optical, piezoelectrical, thermal or mechanical principals [3-

6]. In practical, any biological transducer could be used in a biosensor for the measurement of glucose. However, in practice, electrochemical methods have dominated. Device for this type offer suitable sensitivity and reproducibility and importantly, can be manufactured inexpensively. Immobilization of the enzyme of the electrode surface is a key technique in the fabrication of enzyme electrode. The electrochemical method that is often employed is amperometry. Thus, amperometric enzyme electrode with glucose oxidase bound to electrode transducers have therefore investigated in great detail. Recent commercial development of these devices allowed to measured glucose concentration with enough exactly. However, the device is largely limited to monitoring patterns and trends in blood sugar, and it requires daily calibration by a standard glucose meter. Its response is also affected significantly by motion, perspiration and temperature. And finally, because of safety consideration such oxygen-independent devices should be more appropriate for in-vitro glucose testing than to in-vivo monitoring applications. Looking for new techniques with high sensitivity and efficiency for the detection of glucose concentration is of importance in biosensor construction.

In present article other type of a biosensor which one works in microwave range is discussed. We monitor the glucose concentration in three different type grape samples (red, black green) by measuring the microwave reflection coefficient S_{11} using a microwave cavity sensor. The change of the glucose concentration is directly related to the change in the reflection coefficient due to an electromagnetic interaction between the resonator and blood sample.

2. Analysis and Design

2.1. Design of waveguide cavity sensor

The fabricated glucose sensor based on the design of a waveguide cavity resonator is shown in Fig. 1 (A) and a detailed description is given in Ref. [7].

Underlie of the sensor the $\lambda/4$ is resonant cavity configuration with TE_{011} mode. Also, $\lambda/4$ resonator easily couple with propagation

input/output ports. Having a high Q factor (>2000) and small loss, it allows to use for tune voltage-controlled oscillators.

To determine the glucose concentration changes, we measured the microwave reflection coefficient S_{11} and the resonant frequency shift of the microwave resonator using a network analyzer. Subsequent changes in electromagnetic coupling between the cavity and the sample cause changes in the magnitude of S_{11} and shift the resonant frequency. This allows characterization of the electromagnetic properties (dielectric, conducting, volumetric etc.) of the sample.

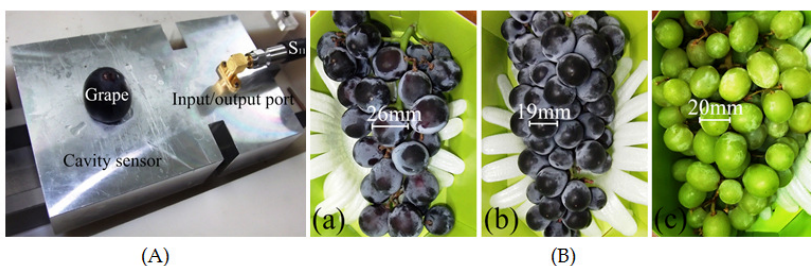


Fig. 1. (A) Resonant cavity sensor with grape sample under test. (B) Grapes samples used in experiments: (a) red, (b) black, and (c) green types.

Grape sample directly located in gap that make in top part of resonator cavity, and therefore the slightest variation of physical properties of a material discovering inside entails variations of a resonance condition, and as a consequent to mismatching coupling mode. Thus the resonance frequency and amplitude varies. And it depends on a percentage ratio of glucose in sample on matching with water. At that, it is diminished frequency rates, and value of amplitude increases. Any further variation of a dielectric permittivity in a grape (due to changes of glucose concentration) brings change in reflection (S_{11}) coefficient of sensor. As a result, shift resonance frequency and matching condition. This change permits to measure glucose concentration.

We used three types (and three samples for every types) of grapes samples with different glucose concentrations and sample linear sizes (Fig. 1 (B)). The linear sizes and reference glucose concentrations for samples show in Table 1. Reference value for glucose level measured

using standard commercial glucometer (Palette PR-101 with 0-45 % (0-450 mg/ml) range of Atago Corp.). The samples conventionally distributed in three groups with high (red grapes), middle (green grapes), and low (black grapes) glucose levels. As reference we measured all grapes juice (without any seeds and pulps) in the same plastic cylindrical tube with 14 mm × 13 mm sizes and with 1 mm wall thickness. All measurements were done at 22 °C.

Table 1. Characteristic sizes and reference glucose concentrations for grape samples.

<i>Sample</i>	<i>Characteristic size (mm)</i>	<i>Glucose concentration (mg/ml)</i>
red 1	26.9	176
red 2	28.7	179
red 3	26.2	202
black 1	19.3	138
black 2	18.5	155
black 3	18.3	162
green 1	21.5	128
green 2	18.6	129
green 3	20.2	141

2.2. Microwave reflection coefficient

The microwave reflection principle can be explained by the plane-wave solution model. An interpretation of the reflection coefficient S_{11} magnitude in terms of medium parameters is written as [8]

$$S_{11} = 20 \log \left| \frac{|Z_{in}| - Z_0}{|Z_{in}| + Z_0} \right|, \quad (1)$$

where Z_0 is the characteristic impedance of the probe tip and Z_{in} is the complex input impedance of the sensor/sample system and it is function on electromagnetic parameters of sample, for example dielectric permittivity, i.e. concentration of grapes.

The dependence of dielectric permittivity on glucose concentration is expressed with the molar increment δ and given by [9]:

$$\varepsilon_g(\omega) = (\varepsilon_0' + c\delta') - j(\varepsilon_0'' + c\delta'') = (\varepsilon_0' - j\varepsilon_0'') + c(\delta' - j\delta''), \quad (2)$$

where c is the concentration of glucose, $\delta = \delta' - j\delta''$ is the increase in permittivity when the glucose concentration is raised by 1 unit: $\delta' = 0.0577 \text{ (mg/dl)}^{-1}$ and $\delta'' = 0.0015 \text{ (mg/dl)}^{-1}$ and $\varepsilon_0 = \varepsilon_0' - j\varepsilon_0'' = 74.37 - j16.518$ is the complex permittivity of de-ionized (DI) water at 4.75 GHz [10].

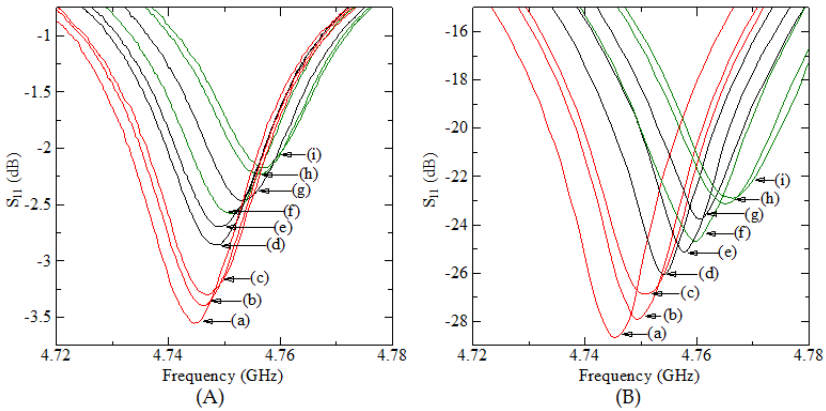


Fig. 2. (A) Measured microwave reflection coefficient S_{11} profiles for grape samples: (a) red 1, (b) red 2, and (c) red 3, (d) black 1, (e) black 2, (f) green 1, (g) black 3, (h) green 2, and (i) green 3. (B) Measured microwave reflection coefficient S_{11} profiles for grape juice in the same plastic cylindrical tube with 14 mm $\times$ 13 mm sizes: (a) red 1, (b) red 2, and (c) red 3, (d) black 1, (e) black 2, (f) green 1, (g) black 3, (h) green 2, and (i) green 3.

3. Results and Discussion

Figure 2 (A) shows measured microwave reflection coefficient S_{11} profiles for the all grape samples at the resonant frequency of about 4.76 GHz. As the frequency of the microwave source is swept, each trace in Fig. 2 exhibited a minimum corresponding to the standing wave mode for that particular resonator-sample combination. The matched

resonance curve for the reference sample has a minimum level of -0.2 dB, which is the reference level for the microwave reflection coefficient S_{11} in these measurements. As the glucose concentration increased, the dielectric permittivity of sample increased and the reflection coefficient S_{11} decreased as shown in Fig. 3 (left axis). Note that the microwave reflection coefficient S_{11} is directly related to the sample electromagnetic parameters (in this case, dielectric permittivity). Here, this relation thus the electric distribution (not shown here) behavior shows the reverse dependence on glucose concentration due to the “two chambers” construction and mode structure of the resonant cavity.

The microwave response is sensitive to the change of the sample volume, temperature and humidity [7]. To distinguish the insertion of glucose concentration effect to others, we measured the separated grape juice (from same measured grape samples) with same 2 ml volume at 25 °C and 50% relative humidity. Figure 2 (B) shows measured microwave reflection coefficient S_{11} profiles for the grape juice with same volume at the resonant frequency of about 4.75 GHz. The matched resonance curve for the sample filled with de-ionized water (i.e. glucose concentration was 0 mg/ml) has a minimum level of -15.1 dB, which is the reference level for the microwave reflection coefficient S_{11} in these measurements. Again, as the glucose concentration increased, the dielectric permittivity of sample increased and the reflection coefficient S_{11} decreased as shown in Fig. 3 (right axis). Obtained result shows that the volumetric effect was weaker compare to the glucose concentration effect, thus the results for grape samples and grape juice samples were same. However, the sensitivity of measurement was higher for juice sample due to dipper insertion of material under test in resonant cavity. The cylindrical tube has more regular shape and so, interaction of the sample and the electromagnetic field in resonator is more effective in this case.

From the linear relationship as a function of glucose concentration $\Delta S_{11}/\Delta c = 0.02 \text{ dB}/(\text{mg/ml})$ or $\Delta S_{11}/\Delta c = 0.0017 \text{ 1}/(\text{mg/ml})$ in the linear scale for the grape samples and $\Delta S_{11}/\Delta c = 0.08 \text{ dB}/(\text{mg/ml})$ or $\Delta S_{11}/\Delta c = 0.0028 \text{ 1}/(\text{mg/ml})$ in the linear scale for the grape juice samples. The root-mean-square (rms) statistical noise in reflection coefficient S_{11} was

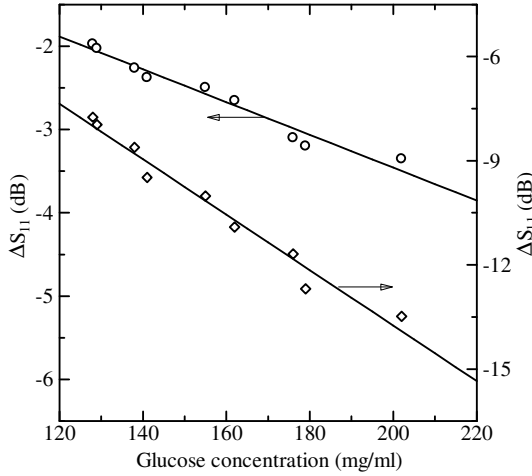


Fig. 3. Measured change in microwave reflection coefficient minima S_{11} vs. glucose concentration for grape samples (left axis) and grape juice in the same plastic cylindrical tube with 14 mm $\times$ 13 mm sizes (right axis). The data was estimated from Fig. 2.

about 10^{-5} in the linear scale [11]. The measured signal-to-noise (SNR) was about 44 dB and 49 dB for the grape samples and for the grape juice samples, respectively. The smallest detectable changes in concentration based on a criterion of SNR of 44 dB and 49 dB were about 10 mg/dl and 4 mg/ml, respectively.

4. Conclusion

A new type microwave cavity sensor has been developed for determination of glucose concentration in grape (and other fruit) samples. The microwave cavity sensor is a novel non-invasive glucometer with a minimum 4 mg/ml detectable change in sample. The results show the sensitivity and the usefulness of the cavity sensor for sensing and monitoring of glucose concentration. It can be useful for the real-time measurements of glucose concentration, and potentially it is an interesting approach for noninvasive in-vitro measurement of human glycemia.

Acknowledgements

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5. References

- [1] J. Wang, , Electroanalysis. **13** (2001) 983.
- [2] S. Zhang, G. Wright, Y. Yang, Biosens. Bioelectron **15** (2000) 273.
- [3] F. G. Tseng, K. H. Lin, H. T Hsu, Ch. Ch. Chieng, Sens. Actuators A **111** (2004) 107.
- [4] J. McKee, B. Johnson, IEEE Trans. Instrum. Meas. **49** (2000) 114.
- [5] M. Pan, X. Guo, Q. Cai, G. Li, Y. Chen, “Sens. Actuators A **108** (2003) 258.
- [6] A. Subramanian, P. Oden, S. Kennel, K. Jacobson, R. Warmack, T. Thundat, M. Doktycz, Appl. Phys. Lett. **81** (2002) 385.
- [7] Kim S, Melikyan H, Kim J, Babajanyan A, Lee J, Enkhtur L, Friedman B, Lee K. Diabetes Res. Clin. Pract. **96** (2012) 379.
- [8] D. M. Pozar, Microwave Engineering. 3rd ed. (Anderson-Wesley, New York, 2005).
- [9] H. Melikyan, E. Danielyan, S. Kim, J. Kim, A. Babajanyan, J. Lee, B. Friedman, K. Lee, Med. Eng. Phys. **34** (2012) 299.
- [10] D. Lide, Handbook of Chemistry and Physics. 85th ed. (CRC Press, Philadelphia, 2004).
- [11] J. Kim, A. Babajanyan, A. Hovsepyan, K. Lee, B. Friedman, Rev. Sci. Instrum. **79** (2008) 086107.

Change of physical-chemical properties tumoral DNA irradiated by low power millimeter waves

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By the methods of spectrophotometry and densitometry thermostability and density of water-salt solutions of DNA, irradiated by low power millimeter coherent waves with frequency 64.5 GHz have been investigated. It is shown that depend on time of irradiation the thermostability of DNA and density of its solutions are increased. It is expected that under influence of millimeter electromagnetic radiation the hydration of DNA and being present in solution ions of Na^+ increase in consequence of which physicochemical characteristics of DNA are changed.

The changes in tumoral DNA after MM-irradiation

It is known that with the help of melting differential curves (MDC) it can be distinguished DNA tumor sarcoma (tDNA) from DNA isolated from the liver of healthy mice (hDNA). MDC of tumor DNA are shifted relatively MDC of the liver DNA to lower temperatures, and in the MDC of tumor DNA there are appeared the additional peaks in the 52-60°C, which is absent for MDC of liver DNA of healthy animals (Fig.1). The effect of MM radiation with a frequency of 42.2 GHz is investigated in vivo on the structure of DNA secondary structure of sarcoma 37. The following table shows the values of temperature (T_m^0) and interval (ΔT^0) of melting and content of 5-MC in the studied samples of DNA. The interesting us parameters characterizing the primary and secondary structures of DNA, under the action of MM in 30 minutes waves undergo certain changes (see Table 1). Without going into the nature of the observed changes in the parameters of melting and the content of 5-MC in the DNA of the tumor and liver of healthy animals, we examined the effect of MM-radiation on the

structure of DNA in vivo, based on the nature of the changes in the parameters of melting and the content of 5-MC.

As can be seen from the Table 1, the tumor DNA has the high level of methylation (4.7 mol.%), which after 30 minutes influence of MM-radiation becomes (2.2 mol.%) close to the corresponding value for healthy DNA (1.9 mol.%). The received results are correlated with the spectrophotometric data (Table 1 and Fig.1). Under the influence of MM-radiation the values of T_m^0C and ΔT^0C of tDNA are changed and approach to the corresponding values of hDNA (Table 1 and Fig.1). The experimental data presented in the Fig. 1 and Table 1 show that, it is quite possibly, 30 min of the MM-radiation leads to the activation of specific molecular mechanisms of cells, resulting in decreased undesirable structural changes in the tDNA, resulting in inhibition of tumor growth. Let's analyze the melting differential curves shown in Fig.1. The characteristic for MDC tDNA low-temperature peaks in the region of 54-62⁰C and MDC of tDNA under the action of the MM-radiation almost disappear in the form and become close to those for MDC hDNA, but the curve is still shifted to lower temperatures compared to the MDC hDNA. The shift of MDC of tDNA as a result of exposure to the direction of MDC hDNA is provided, apparently, due to MM radiation is decreased the proportion of tumor cells in the tumor.

Thus, the correlation data between the ability of MM-radiation to modify the structure and content of 5-MC in tumor DNA in vivo and inhibition of tumor growth, allow to assume that the MM-radiation with a frequency of 42,2 GHz has antitumor activity. The MM-waves general toxic influence on the experimental animals with sarcoma 37 without cytostatics is negligible. Antitumor effect of coherent MM-waves obtained without drugs, shows promising development of millimeter therapy for clinical oncology in the treatment of malignancies.

The change of the thermostability of irradiated DNA

Tumoral and healthy DNA's physical-chemical characteristics changes have been investigated under radiation at frequency 64.5GHz, in correspondence with resonance frequency of oscillations of

hexagonal molecular structures of water. In this case was used 1 Hz amplitude modulation .

Studies have shown that the form of the melting curves, the values of T_m and ΔT , do not exhibit a certain dependence on the duration of post-irradiated term, since after irradiation both about 12, and 24 h, these parameters are within experimental error. It is found that, depending on the duration of exposure the thermostability of DNA increases, which is more pronounced for tDNA (Table 2). Upon irradiation for 90 min T_m tDNA is increased by about $1,5^{\circ}\text{C}$, while the ΔT is decreased. Perhaps, the irradiation leads to the ordering of water molecules associated with the macromolecule, especially in AT-rich regions, which in turn affects the compaction of the macromolecule, and this, in turn, affects the T_m and ΔT .

To confirm this fact there are received MDC of irradiated and non-irradiated DNA. Fig. 2 shows the MDC of irradiated for 90 min and non-irradiated tDNA. As can be seen from the figure, MDC tDNA is shifted toward the high temperatures in comparison with non-irradiated tDNA. A similar increase of T_m was also obtained for hDNA, DNA of calf thymus, but this parameter is less ($1,0^{\circ}\text{C}$) than at tDNA. With increasing of exposure duration (> 90 min) T_m and ΔT both of hDNA and tDNA practically do not change, which, in all probability, is due to the fact that the water structuring degree does not undergo further changes. The values of melting parameters for hDNA and tDNA are summarized in Table 2. As can be seen from the table data, the dynamics of changing of T_m and ΔT for tDNA is more pronounced than for tDNA in case of irradiation with low intensity MM-waves during the increasing of exposure duration.

It should be noted that the T_m and ΔT of unexposed hDNA and tDNA do not coincide: T_m tDNA is about $0,5^{\circ}\text{C}$ lower than hDNA, while ΔT is higher for tDNA (Table 2). This is apparently due to the presence of "defective" parts in tDNA molecule arising as a result of methylation and subsequent enzymatic dezamination of cytosine and transformation its to thymine, which leads to the formation of unstable guanine–thymine pair. As a result, the locally denatured regions are formed in DNA molecule, which leads to the reduction of T_m tDNA.

The conformational transitions into the hypermethylated parts of DNA molecule are possible as well. Due to the above mentioned structural differences there is more pronounced change in tDNA hydration during the irradiation, and it provided by the melting temperature increasing.

The assumption that changes in the DNA melting parameters under the influence of low intensity MM-waves are provided by the structure of water, is based on the fact that the resonant absorption frequencies of DNA are in the region of 2 to 9 GHz (Rodionov 1999). Hence, we assume that at a frequency of 64.5 GHz, the changes in the values T_m and ΔT can not be due to the resonance absorption of DNA. Consequently, the increase in the thermostability of DNA during the irradiation by MM-waves with a frequency 64.5 GHz can be caused by their through the water mediated influence. DNA-samples were prepared in the irradiated only water-salt solution (buffer) for the confirmation of mentioned fact. Melting curves obtained for them does not practically differ from the curves obtained by irradiation of DNA solutions within the experimental error. Therefore, it can be assumed that the observed changes in the parameters of DNA-melting caused just by changes in the structure of water arising due to exposure.

The change under radiation the density of aqueous salt solutions of DNA

This is indicated also by the results on the measurement of the density of aqueous salt solutions of DNA in case of irradiating by MM-waves. There were also measured for control the densities of bidistilled water and water-salt solution before and after irradiation. Density of water, 0.1×SSC and DNA solutions was determined on densitometer DMA 4500 Anton Paar (USA), with resolutions 10^{-5} g/cm³. Studies have shown that in case of irradiation by pure water with a frequency of 64.5 GHz, its density does not practically change, while the density of the buffer and the DNA-solution increases. This indicates that the structural state of pure water does not change due to irradiation, since under these medium conditions the water molecules form a most stable, from a thermodynamic point of view, structure, and an increase in ordering after exposure becomes thermodynamically non-profit.

Therefore, the density of water under these conditions should not be changed. In contrast, in case of irradiation of the buffer and the DNA-solution, some of the free water molecules ("not included" in composition of the most common hexagonal structures) is structured around the dissolved ions or macromolecules (increasing the hydration degree). Moreover, in all probability, the water molecules are involved in the formation of additional bonds with the salt ions or with functional and atomic groups of macromolecules, which leads to an increase in size of the ions or macromolecules, and the latter is the cause of density increasing.

The results of measurements of the density buffer and the DNA-solution are summarized in Table 3. As can be seen from the table, there is almost the same dynamics of changing of the buffer and the DNA-solution densities. And the received data are in a good agreement with the results of DNA-melting. There is also studied the density dependence of the temperature of the DNA-solution, in case of irradiation by duration of 90 and 120 min, to detect changes in the structure of water by irradiation, depending on temperature. It is found that with increasing of temperature the density of the irradiated and non-irradiated DNA is reduced, but there is a significant difference between the solution of the irradiated and non-irradiated DNA.

Fig.3 shows the dependence of the density difference ($\Delta\rho$) on temperature T . Here $\Delta\rho$ is the difference between the solution densities for irradiated and non-irradiated DNA. Dependences of $\Delta\rho$ on T at 90 and 120 min of irradiation is not differ significantly. As can be seen from the figure, the dependence of $\Delta\rho$ on T increases slightly in the range of temperature $20 < T < 40^{\circ}\text{C}$, and in the range of $40 < T < 70^{\circ}\text{C}$ there is obtained a sharp decrease in $\Delta\rho$. With further increase in temperature ($T > 70^{\circ}\text{C}$) $\Delta\rho$ sharply increases, due to reveals a minimum at a temperature of about 70°C on the dependence curve, that corresponds to the melting point of DNA. As follows from the spectrophotometric melting curves, denaturation of DNA occurs in the interval of temperature changes $60 < T < 85^{\circ}\text{C}$. Therefore, this dependence may be caused by the fact that the hydration of the irradiated DNA with temperature increasing decreases to a greater extent than in

case of non-irradiated DNA, and at $T=70^{\circ}\text{C}$, when a half of DNA is in a melted state, the reducing of hydration degree of the irradiated DNA is the maximal. The further increase in temperature leads to a sharp increase of $\Delta\rho$. The sharp increase of $\Delta\rho$ at $T>70^{\circ}\text{C}$, to all appearances, is a consequence of the fact that in the single-stranded (ss) state the degree of hydration of the irradiated DNA-molecules is higher, than in the double-stranded (ds) state (~ 5 times), and, on the other hand, the single-stranded DNA-molecules probably becoming the "centers of crystallization" for the water molecules, so that the density of system "water-irradiated ss-DNA" is increased in comparison with non-irradiated ones. It is assumed that during irradiation of some part of the "free" water molecules, which were involved in the hydrate structure of DNA, is released after exposure, leading to a sharp decrease in the density of the water-DNA system, while an analogous phenomenon does not occur in the case of non-irradiated DNA, and the density of the latter undergoes minor changes. Further, with an increase in temperature there is an increase of the lability both of macromolecule and hexagonal structures of water, so that, in all likelihood, these structures are involved in the hydration shell of macromolecule, and this system density is increased.

Table 1. The content of 5-methylcytosine and DNA melting parameters under the influence of MM-radiation at 42.2 GHz

Experimental conditions	Source of DNA	5-MC mol. %	ΔT $^{\circ}\text{C}$	T_m $^{\circ}\text{C}$
Animals with sarcoma 37	tumor	$4,7\pm 0,1$	$7,6\pm 0,1$	$70,6\pm 0,1$
Healthy animals	liver	$1,9\pm 0,1$	$6,6\pm 0,1$	$71,8\pm 0,2$
Animals with sarcoma 37 MM-therapy effect 30 min.	tumor	$2,2\pm 0,1$	$6,9\pm 0,1$	$71,7\pm 0,2$

Table 2. Temperature and range of DNA melting obtained from of healthy rats liver and tumor sarcoma 45 at 64.5GHz

Time of irradiation, min.	_hDNA		_{s-45}DNA	
	$T_m, ^\circ\text{C}$	$\Delta T, ^\circ\text{C}$	$T_m, ^\circ\text{C}$	$\Delta T, ^\circ\text{C}$
0	69.4 ± 0.1	7.2 ± 0.2	68.8 ± 0.2	7.9 ± 0.2
30	69.4 ± 0.1	7.2 ± 0.2	68.9 ± 0.1	7.9 ± 0.2
40	69.5 ± 0.2	7.1 ± 0.2	69.0 ± 0.1	7.8 ± 0.2
60	69.9 ± 0.1	7.0 ± 0.2	69.8 ± 0.1	7.8 ± 0.2
90	70.3 ± 0.2	7.0 ± 0.2	70.2 ± 0.2	7.6 ± 0.2
120	70.4 ± 0.2	6.9 ± 0.2	70.2 ± 0.2	7.5 ± 0.2

Table 3. Magnitude of solutions density (g/cm^3) before and after exposure of MM-radiation

Time of irradiation, min.	Buffer	Buffer + DNA
0	0.999201 ± 0.000005	0.999232 ± 0.000004
30	0.999220 ± 0.000005	0.999242 ± 0.000005
60	0.999241 ± 0.000004	0.999269 ± 0.000004
90	0.999253 ± 0.000004	0.999291 ± 0.000005

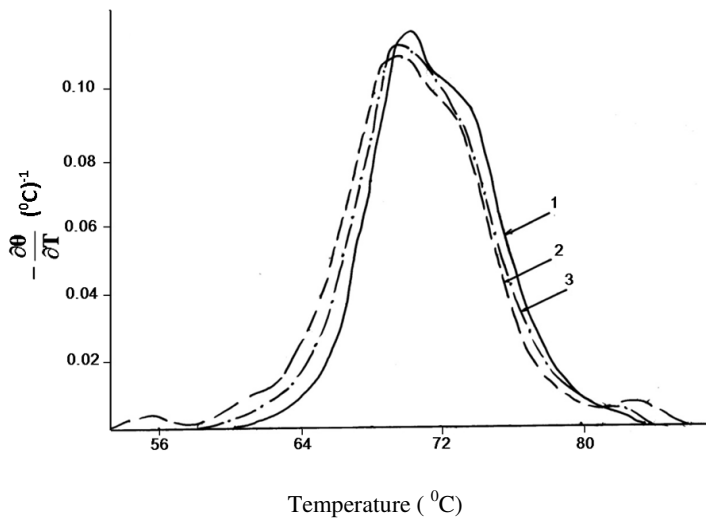


Figure 1. Differential melting curves DNA of healthy mice(1),DNA of sarcoma-37 (2) and DNA of sarcoma-37 after 0.5 hour by millimeter electromagnetic waves irradiation at 42.2 GHz in vivo (3).

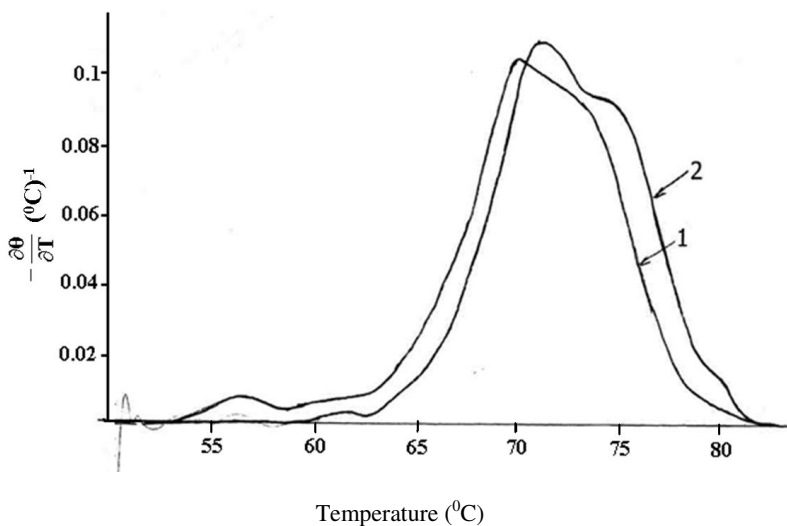


Figure 2. Melting differential curves of nonirradiated (1) and irradiated for 90 min by 64.5GHz with 1 Hz amplitude modulation (2) Sarcoma 45 DNA

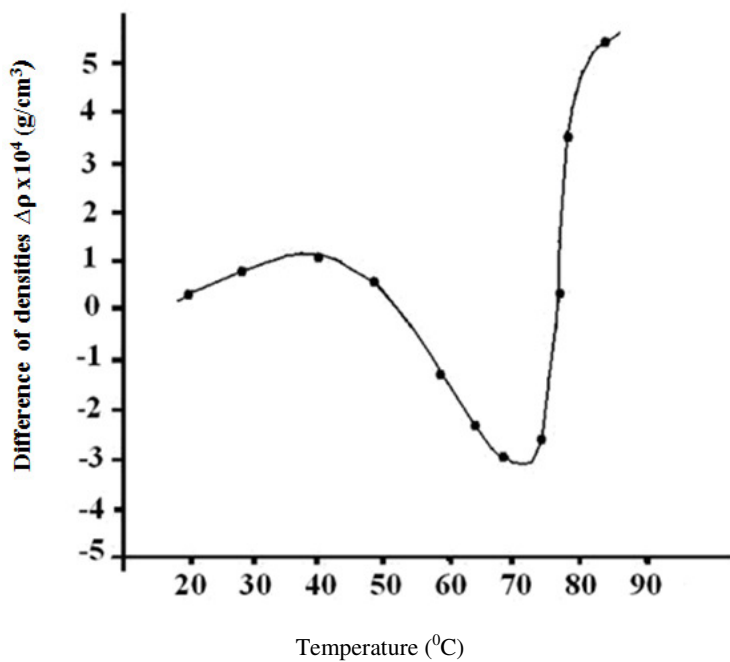


Figure 3. Curve of dependence of $\Delta\rho$ —difference of density of solutions irradiated for 90min and non-irradiated DNA on temperature

Stability of Bilayer Lipid Membrane Under Influence of Low Intensity Millimeter Waves

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In the presented work we have experimentally studied the low-intensity (non-thermal) influence of electromagnetic waves in the millimeter and decimeter ranges on the stability of bilayer lipid membrane (BLM). It is shown that their action leads to a decrease in the BLM-stability in an electric field. The experiments indicate that the effect of millimeter waves leads to an increase in the number of pores in the BLM. Decrease in the BLM-average lifetime is more pronounced under the action of non-resonant frequency millimeter waves. It is also shown that the effect of decimeter waves is connected both with increase in the number of pores in the BLM and with decrease of the pore formation work.

It is known that the most probable mechanism of action of electromagnetic waves on the biological objects is associated with their interaction with plasmatic membranes of cells and water. The results of numerous experimental works indicate a high sensitivity of membrane processes to the millimeter and decimeter electromagnetic radiation, but despite this, the mechanism of radiation influence remains unclear [1,2]. Due to the extreme complexity of biological membrane, it seems advisable to investigate the effects of electromagnetic radiation on a model membrane system. The model system, which allows us to investigate the influence of radiation on the lipid frame of the biological membrane, is a bilayer lipid membrane (BLM).The choice of BLM is dictated also by the fact that it is the structural basis of biological membrane, and the stability of the biological membrane is almost entirely determined by the stability of its lipid matrix. In this connection, it is interesting to study the influence of millimetric and decimetric waves on the stability of BLM.

Results and discussion

The change in the BLM-average lifetime as a function of increasing of the potential difference in the absence of radiation was studied at first (Fig. 1, curve 1). Then, the potential difference on the average lifetime of the BLM has been experimentally studied after 10 minutes of exposure of millimeter waves with frequencies of 61,22 GHz (curve 3) and 65 GHz (curve 2). Frequency of 65 GHz is the resonant one for water and biological media [2].

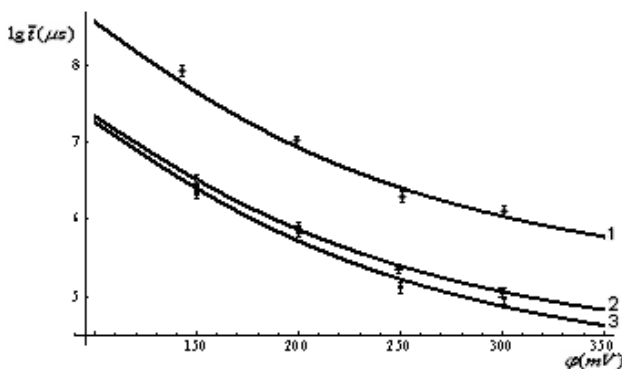


Fig. 1. The dependence of the BLM-average lifetime on the potential difference: 1 – control, 2 and 3 – after 10 minutes of the impact of millimeter waves with frequencies of 65 GHz and 61.22 GHz, respectively. Points – experimental data, solid lines – theoretical curves.

Fig. 1 (curves 2 and 3) shows that the BLM-lifetime decreases after irradiation with millimeter waves. Moreover, the decrease in the BLM-average lifetime for the non-resonant and resonant frequencies is different. The decrease is more pronounced on the non-resonant frequency, than at the resonant one, and with an increase of the potential difference on the BLM, this difference increases.

The loss of stability of the BLM in the electric field is due to appearing of the inverted pores in the BLM and their development up to the critical sizes. The numerical values of the important parameters that affect the BLM-stability can be determined from a comparison of the experimental points in Fig. 1 with the theoretical curve, which can be

obtained from well-known expression for the BLM-mean lifetime derived in [3],

$$\ln \bar{t}(\phi) = A + \frac{B}{1 + \frac{C\phi^2}{2\sigma}} \quad (1)$$

$$A = \ln \frac{(kT)^{3/2}}{4\pi n D \gamma \left(\sigma + \frac{C\phi^2}{2} \right)^{1/2}}, \quad B = \frac{\pi\gamma^2}{\sigma kT},$$

σ - is the surface tension of BLM; γ - is a line tension of edge of the pore in the BLM; n - is the number of pores on the membrane;

D - is the pore diffusion coefficient in the space of radii;

ϕ - is the potential difference on the membrane;

k is the Boltzmann constant;

T - is a temperature;

C - is the reduced electrical capacitance, which is determined by the relation

$C = C_0 (\varepsilon_w / \varepsilon_m - 1)$, where $C_0 = \varepsilon_0 \varepsilon_m / h$ is the membrane specific electric capacity;

ε_w - is the water permittivity;

ε_m - is the BLM-dielectric permittivity;

ε_0 - is the permittivity of vacuum.

The calculations showed that for both frequencies the parameter B practically coincides ($B=3,74$ (control), $B=3,25$ (65 GHz) and $B=3,48$ (61,22 GHz)). In our experiments the error in the determination of the parameter B was approximately equal to $\pm 0,1$, and the error in the determination of the parameter A was approximately equal to ± 10 . The values of the parameter A are different ($A=335,86$ (control), $A=142,17$ (65 GHz) and $A=111,48$ (61,22 GHz)). Since the parameter A is related to the number of pores on the BLM, and parameter B is related to the work of formation of pores with critical sizes, then the obtained results indicate that the decrease in the BLM-average lifetime under the

influence of millimetric waves is connected with increase in the number of pores in the BLM.

Then the influence of the decimeter waves on the BLM-stability was investigated. As in the case of millimetric waves, initially was investigated the change in the average BLM-lifetime as a function of the potential difference increasing in the absence of decimetric waves (Fig. 2, curve 1). Thereafter, the influence of potential difference on the BLM average lifetime has been experimentally studied, when radiation of millimetric waves are exposed to BLM (Fig. 2, curve 2 and 3). Fig. 2 shows that the influence of electromagnetic radiation of the decimeter waves also leads to a decrease in the BLM-average lifetime. Moreover, the effect of decimeter waves for 5 minutes (curve 2) is practically the same as in case of the action for 10 minutes (curve 3).

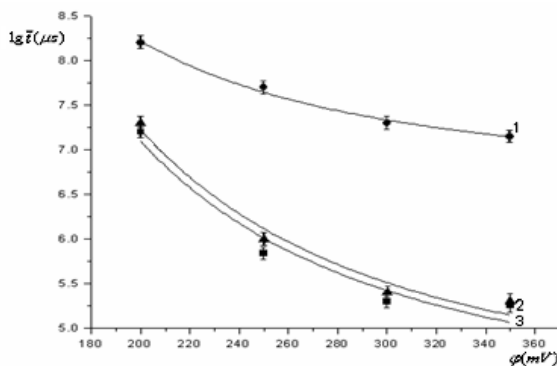


Fig. 2. The dependence of the BLM-average lifetime on the potential difference: 1 – control, 2 and 3 – after 5 min and 10 min of exposure of the decimeter waves with a frequency of 980 MHz. Points – experimental data, solid lines – theoretical curves.

In a result of comparison of the theoretical curves with the experimental data, there were identified the following values for the parameters A and B: B=3,02 (curve 1), B=7,3 (curve 2) and B=7,06 (curve 3); A=1536,9 (curve 1), A=89,2 (curve 2) and A=86,95 (curve 3). The obtained results indicate that influence of decimetric waves to the BLM leads both to a decrease in the work of critical pore formation, and to

increase in the number of defects in BLM, which in its turn results in a decrease in the average lifetime of BLM [3].

Thus, in the presented work we have experimentally studied the low-intensity influence of electromagnetic waves in the millimeter and decimeter ranges on the stability of BLM. It is shown that their action leads to a decrease in the BLM-stability in an electric field. The experiments indicate that the effect of millimeter waves leads to an increase in the number of pores in the BLM. Decrease in the BLM-average lifetime is more pronounced under the action of non-resonant frequency millimeter waves. It is also shown that the effect of decimeter waves is connected both with increase in the number of pores in the BLM and with decrease of the pore formation work.

References

1. N.D Devyatkov, M.B.Golant, and O.V.Betskiy. Millimeter waves and their role in vital activity processes. Radio and Communication press, Moscow, 1991.
2. V.I.Petrosyan, N.I.Sinitin, V.A.Elkin, N.D.Devyatkov, J.V.Gulyayev, O.V.Betskii, L.A.Lisenkova, and A.I.Gulyayev. Role of resonance molecular-wave processes in the nature and their use for the control and corrections of a condition of ecological systems. Biomeditsinskaya Radioelectronica, N 5-6, may-june 2001.
3. I.G.Abidor, V.B.Arakelyan, V.F Pastushenko, L.V. Chernomordik, and Y.A.Chizmajev. Mechanizm of electric breakdown of BLM. Reports of Academy Sciences USSR, vol. 245, no 5, 1979.

Method of cardiac activity extraction in L-Band CW radars

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Abstract

Azimuthal scanning of the human body by CW Doppler radar is considered. The method of separation of oscillation activities of various internal organs in the microwave Doppler spectra is proposed. Two possible setups with fixed human body + scanning antennas and fixed antennas + rotating human body have been investigated.

Introduction

Microwave Doppler radar sensors of biological activity are effective tools for both diagnostic purposes and for use in security and search for life signs during natural disasters [1-3]. It is well known that spectra of Doppler radar signals reflected from the internal organs represent complex mathematical objects [2]. If the simple detection systems of biological activity require only a decision making concerning the presence or the absence of a life, while the non-contact diagnostic systems require to carry out separation of activities caused by various organs. We propose mathematical method and processing algorithm for possible separation of oscillatory activities caused by difficult organs in microwave Doppler spectra. Such an express diagnostic method can be a promising candidate in telemedicine applications.

It is well known, that the spectrum of the signal reflected from oscillating target is identical to the spectrum of FM signal[4,5]:

$$R_x = B e^{i \omega t} e^{i[m \cos(\Omega t) + \Psi]}, \quad m = \frac{2v_m}{\lambda}$$

Such consideration can explain the periodic-like spectrum of pulmonary and cardiac activities. Here modulation index m is determined by the amplitude of oscillator $2v_m/\lambda$. Spectrum of FM signal with modulation index m less than 0.5 represents itself main tone and only two sidelobes with opposite phases.

While for the modulation indexes m greater than 1, there are theoretically an infinite number spectral components, but we can neglect the members of Fourier series with $n \geq m + 2$ [4].

The bandwidth of FM signal for $m \gg 1$ looks like: $2\Delta\omega \approx 2\omega_D = 2m\Omega$, i.e. the spectral width itself is proportional to the amplitude of oscillator velocity.

It should be noted that only longitudinal oscillations, or oscillations having significant longitudinal component can contribute to the Doppler spectrum. In view of the obvious difference in the symmetry of mechanical activity of the heart (more isotropic) and respiratory organs (“strictly polarized”), the velocity of the oscillations should also be subject to the same symmetry. Thus, the spectral width of respiratory activity should depend on the receiver angle (Fig.1).

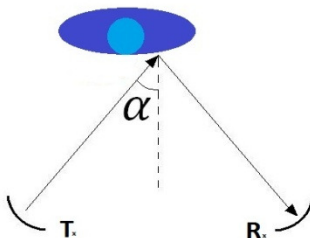


Fig. 1. Experimental setup. Tx – transmitter, Rx – receiver (top view)

The direction itself can be taken into account by modifying the expression for the modulation index (3) $m = \frac{2\pi v_m}{\lambda} \cos\alpha$, where α is the receiver angle.

It should be noted that the change in pulmonic amplitude spectrum is nonmonotonic with a change of receiver angle α . The latter is due to the fact that the receiver angle α is included in the respective argument of the Bessel function, which is known to have an oscillating and alternating form. The foregoing makes sense only for the respiratory components of the spectra. Due to the inherent isotropic symmetry, the spectra of cardiac activity should not significantly depend on the

receiver angle α . Thus, changing component of the Doppler spectrum can be attributed to respiratory activity.

Measurement and Data Processing Methods

CW microwave Doppler radar has been used with power less than 1 mW at 1 GHz band. First experimental series has been carried out with fixed human body and scanning antennas. Separate receiving and transmitting antenna were located at an relative angle 2α . The measurements were performed at three different mutual orientations of the transmitting and receiving antennas: 0° , 90° and 135° (Fig.2).

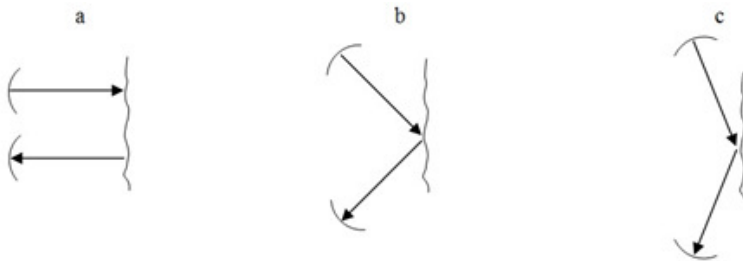
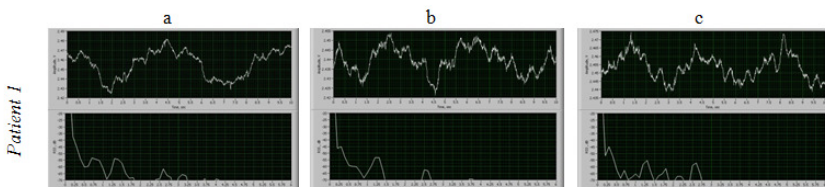


Fig. 2. Different spatial orientations of Tx and Rx antennas (a. $\alpha = 0^\circ$; b. $\alpha = 90^\circ$; c. $\alpha = 135^\circ$). Top view

Fig. 3 shows the results of two series of measurements corresponding to two different patients. Three different relative orientations of the antennas have been implemented for each of these measurements. In order to extract spectral components which are independent of the direction, the averaging of the spectra over different antenna orientations has been produced (Fig.4).



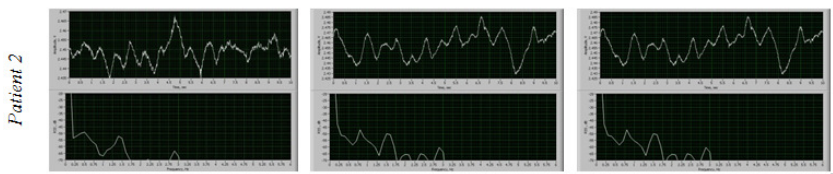


Fig. 3. Typical mixed cardiac/pulmonic activities for two patients. Uppers – time domain; lowers – power spectrum of received signal strength (RSS). (a. $\alpha = 0^\circ$; b. $\alpha = 90^\circ$; c. $\alpha = 135^\circ$).

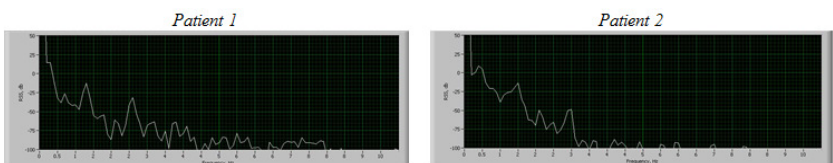


Fig. 4. Averaged spectra over azimuthal angle.

Second series correspond to setup with fixed antennas of $2\alpha=135^\circ$ and rotating human body. Three different orientations of the human body have been implemented for each of these measurements. The schema of three relative orientations is presented below (Fig. 5).

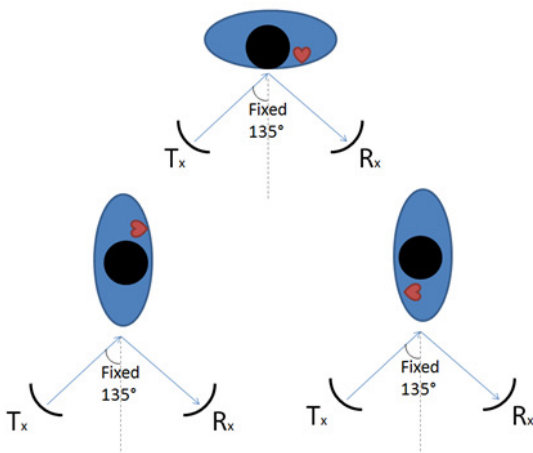


Fig. 5. Three relative orientations of human body and antennas

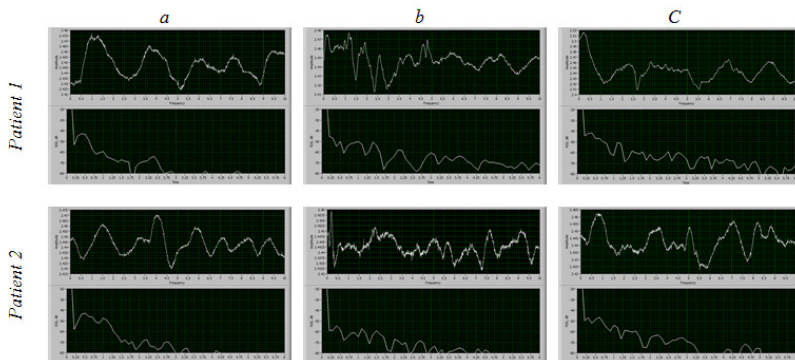


Fig 6. Typical mixed cardiac/pulmonic activities for two patients. Uppers – time domain; lowers – power spectra of received signal strength (RSS) for rotating human body + fixed antennas.

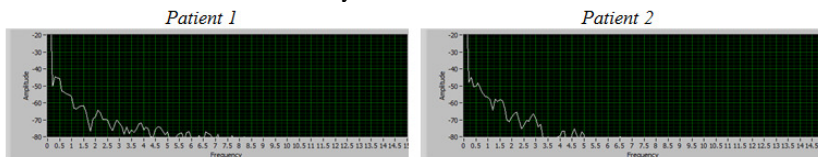


Fig 7. Averaged spectra for rotating human body + fixed antennas over azimuthal angle

Discussion and Conclusion

As it can be seen, the method of averaging of Doppler spectra both over the azimuthal scanning of antennas (Fig.4), and rotating human body (Fig.7) allows us to isolate the spectrum of cardiac activity from a broadband and intense spectrum of pulmonic activity. For comparison, Fig.8 represents an isolated spectrum of cardiac activity which is obtained by breath holding. Qualitative agreement with the results of this method is obvious.

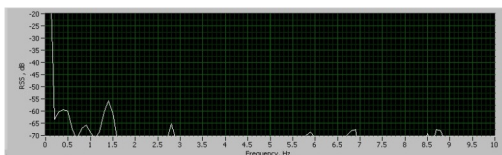


Fig 8. Typical isolated cardiac activity. Power spectrum of received signal strength (RSS).

References

- [1] R.L.Yadava , “RF/Microwaves in bio-medical applications”, Proc. of 8th International Conference on Electromagnetic Interference and Compatibility (INCEMIC-2003), pp. 81-85, 2003.
- [2] J. Lin and C. Li , “Wireless non-contact detection of heartbeat and respiration using low-power microwave radar sensor”, Asia-Pacific Microwave Conference, APMC-2007, pp.1-4, 2007.
- [3] A. Hakhoumian, H. Hayrapetyan, S. Martirosyan, A.Muzhikyan, N. Poghosyan and T. Zakaryan, “L-Band doppler radar for heartbeat sensing”, Proc. of the Int. Conf. On “The Teqnique of microwave and THz waves and its application in biomedical and radar technologies and in remote sensing”, pp. 91-93, 2010.
- [4] D.Calcultt, L.Tetley, Satellite communications: principles and applications, Butterworth-Heinemann, 1994.
- [5] H.Hayrapetyan “Processing Algorithm for Separation of Cardiac and Pulmonic Activities in Reflected Microwave Doppler Spectra”, Transactions of IIAP of NAS of RA, Mathematical Problems of Computer Science, vol. 36, 2012, pp. 99-103.

The Agro Products Sorting System Based on Degree of Maturity

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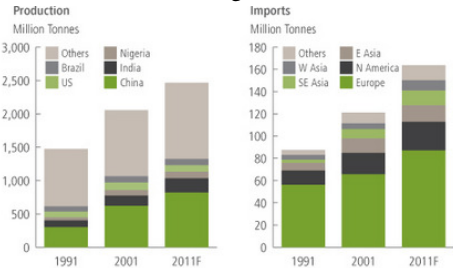
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The current material contains theoretical information associated with usage of NIR-band in social life especially in agricultural engineering. It also covers some specific details related to the new device developed by IRPhE's Applied Radiophysics Laboratory's team. The main purpose is an evaluation of fruits-vegetables' maturity and quality before/after harvesting using NIR waves' properties.

1. Fresh product handling issues

Year by year the volume of productized agro material increases dramatically due to the population growth on Earth. The figure below shows the results of the survey(table1). A large portion of all fresh products is lost after harvest worldwide. The main causes are physiological (wilting, shriveling, chilling injury, etc), pathological (decay due to fungi and bacteria) and physical (mechanical injury), being these causes in many instances interrelated, i.e. mechanical injury can lead to postharvest decay in many cases. Losses are estimated at 20-40% in developing countries and 10-15% in developed countries, depending on the crop. Just in the EU an estimated 4 billion EUR is lost due to postharvest losses and reduced quality of fruit. Here is some evidence:

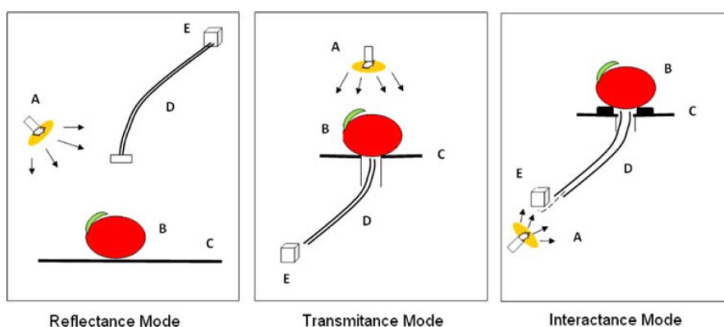
Table 1. Vegetable (tomato, cucumber, pepper, beans) postharvest losses in Southeast Asia range between 13-20%



A new sorting and evaluation system was introduced based on NIR technology as a simple and low cost solution to the above mentioned challenge. Near infrared technology has been extensively and effectively employed in a variety of fields both for research and application utilities, including food, agricultural, chemical, pharmaceuticals, textiles, polymers, cosmetics and medical spheres , though it might have inherent limiting factors. Compared to the silence of the one and a half century after the discovery of NIR, increasing number of scientists and engineers have been devoting themselves to exploiting this old but ‘young’ technology.

2. Sorting system via fruit’s quality indicator

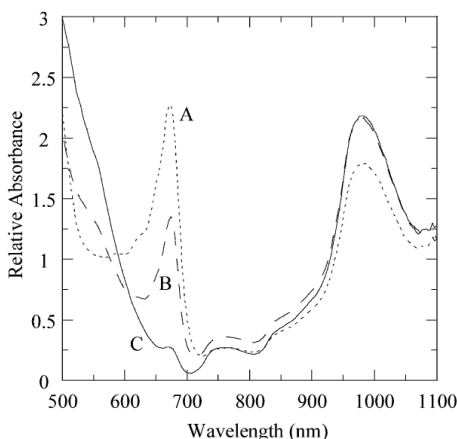
The various components of quality are used to evaluate fruits and vegetables. Quality is classified into external and internal components. Appearance, flavour, texture, nutritive value, and defect factors are generally recognized as the five quality factors of fruits and vegetables. A portable device was prototyped by IRPhE’s Applied Radiophysics Laboratory’s team which evaluates and sorts fruits /vegetables according to their maturity level. The hardware uses a reflection mode from existing three acquisition configurations showed in picture 1.



Picture 1. Sketches : NIR source (A); Object (B); Fruit holder (C); Optic fiber (D); Detector (E).

In reflection mode the information content refers to the reflected signal’s power volume. Due to the absorbance property, part of trans-

mitted energy absorbs by the object (agro product) the other part reflected back. The deviation of absorbance value depends on maturity degree and transmitted waveform length as illustrated in picture 2.



Picture 2. Typical absorbance spectra for (A), mid-harvest (B) and late harvest (C).

NIR sorting equipment works as follows.

A NIR source is focused to illuminate a piece of fruit as it passes under the NIR system. Some of the light penetrates the fruit and is retransmitted. (This effect can be observed by holding a fruit to a bright lamp in a dark room. The fruit will glow at a distance from where the light is shining on it.) The amplitude of the transmitted/reflected light is affected by the internal properties of the fruit and contains information about the internal properties of the fruit. A high maturity fruit will absorb more light at certain wavelength than a low one fruit. Initially when fruit or vegetable is raw the distance among cells in flash is relatively close in contrast to a well-maturated one. Gradually with maturity the distance starts increasing. Accordingly the absorbance starts increasing too. Chlorophyll also plays a vital role as its absorbance value is around 680nm.

Beer's law states that the absorbance is directly proportional to the concentration.

$$A = e \times b \times c \quad (1)$$

Where A is absorbance (no units, since $A = \log_{10} P_0/P$)

e- is the molar absorptivity with units of L mol⁻¹ cm⁻¹

b -is the path length of the sample. We will express this measurement in centimeters.

c -is the concentration, expressed in mol L⁻¹

The amount of radiation absorbed may be measured in below way:

Transmittance:

$$T = \frac{P}{P_0} \quad (2)$$

%Transmittance:

$$\%T = 100 \times T \quad (3)$$

Absorbance:

$$A = \log_{10} \frac{P_0}{P} \quad (4)$$

$$A = \log_{10} \frac{1}{10} \quad (5)$$

$$A = \log_{10} \frac{100}{\%T} \quad (6)$$

$$A = 2 - \log_{10} 10 \times \%T \quad (7)$$

Where P-radiated power, P₀-transmittance power

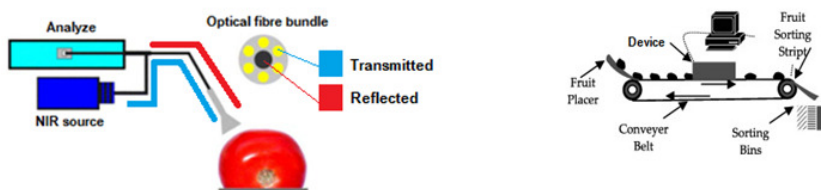
So, if all the light passes through an object without any absorption, then absorbance is zero, and percent transmittance is 100%. If all the light is absorbed, then percent transmittance is zero, and absorption is infinite.

The new developed device uses values of P, P₀ to calculate the absorbance percent and makes a decision of maturity degree. It consists of following components, NIR source (for generating modulated signal), analyzing part (detection, displaying), optical fiber bundle (reflected portion redirection) (picture 3). The device uses this advantage to calculate the absorbance value and makes a decision of maturity degree.

It consists of the NIR source components (for generating modulated signal), analyzing part (detection, displaying) and optical fiber bundle (reflected portion redirection) (picture 3).

Contribution will be done both in programming and hardware design. Here are expected results listed below:

- fast measurement,
- compatibility,
- possibility of two mod functions (production and laboratory).
- This function allows a user to choose whether he/she is going to take a measurement in filed condition as laboratory device or to put in production mode (connect to PC run Software) for line sorting system:
- Portability,
- energy saver,
- Provides solutions for all varieties of fruits,
- Dose not damage produce.



Picture 3. A) Laboratory mod B) Production mod.

Here some technical characteristics of the device:

Table 2. Technical characteristics of the device:

Wave length	980nm
Power consumption	0.3mW
Frequency	80Hz
Modulation index	100%

The main advantage of the presented machine is it will facilitate automatic grading and sorting in a non-destructive method. The whole

machine is cheap compared to other existing solutions. It will also increase export of agro products as it can process large volume in short time and does the quality assessment as per the requirement of export. In our country the entire produce of fruits and vegetables cannot be taken to the storage place because of the lack of processing. But with this machine this can be achieved. It also facilitates the machine sorting where human error is not introduced.

Objective:

This research will develop practical sensors and technologies for quality measurement and grading of fruits and vegetables before, at and after harvest. It also aims to generate new knowledge and understanding of the optical and mechanical properties of fruits and vegetables and their relationship with the physiological factors and quality attributes. A systems approach of integrating sensors development, properties characterization, and models/algorithms development will be applied to attain the following specific objectives:

Objective 1: Develop cost effective sensors and sensing systems to measure and monitor the quality/maturity of individual products.

Objective 2: Develop commercially viable technology to presort and grade fruits and vegetables so as to decrease postharvest handling and storage costs for fruit growers.

Objective 3: Develop technology to accurately and rapidly assess, sort, and grade harvested agro products for multiple internal quality attributes (firmness, flavor, ripeness) and defects.

Fruit maturity measurement will be achieved through integration of NIR technology with the nondestructive firmness measurement method developed in our lab. Algorithms will be developed and integrated into the sensor for real-time measurement of fruit firmness, soluble solids content and other maturity parameters. Laboratory and field tests will be performed to assess the sensor's performance and portability. Commercially viable infield mobile sorting technology will be developed for sorting and grading harvested products into two or three quality grades (fresh market, processing, and cull).

3. References

- [1] V.Andrew McGlone, Robert B. Jordan, Paul J. Martinsen Vis/NIR estimation at harvest of pre- and post-storage quality indices for ‘Royal Gala’ apple Volume 25, Issue 2, June 2002, Pages 135–144
- [2] Yousef Al Ohali Computer vision based date fruit grading system Design and implementation Journal of King Saud University - Computer and Information Sciences Volume 23, Issue 1, January 2011, Pages 29–36
- [3] <http://www.fruitprofits.com/ing/articulo.asp?reg=26>
- [4] Hongjian Lin 卞 Yibin Ying Theory and application of near infrared spectroscopy in assessment of fruit quality: a review.

QPSK mapped OFDM signal processing on Radar applications

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Using QPSK mapped OFDM signal in radar applications it is possible to get from the signal constellations both the target velocity and target distance from the radar station. When in the reflected signal constellation we get the rotated and distorted mapping that is mean that the signal is reflected from the moving target, while on the other hand when the mapping is only rotated around the zero, we have a reflected signal from static target. Getting that signals and passing them trough frequency and arget diustance filter banks we will get the Doppler frequency and time shift and consequently the velocity and the distance of the target.

1. Intorduction

Radar systems develop year by year. And many new techniques have been used to engineer them. One of the latest approaches is using OFDM (Orthogonal Frequency Division Multiplexing) signals in radar systems. A lot of investigations had been done and different method designed to solve various ODFM-Radar applications. One of the approaches designed with using a correlation of received and transmitted signals [1]. Another method is called Novel approach where the proposed algorithm operates directly on modulated symbols [2-3]. There are two major features of OFDM signals which make it applicable in radar applications, which are the signal long duration and the wide spectrum. The first one, signal long duration helps to determine Doppler shift very accurately. On the other hand, wide spectrum of the signal gives an opportunity to find a time shift of the received echo signal. Knowing these two values we can decide the velocity of the target and its distance from the radar station consequently.

2.OFDM in Radar Processing

In Radar processing we can point two general parameters which

describe the accuracy of the radar system: radar range resolution and relative velocity.

$$\Delta r = \frac{c_0}{2B}, (1) \Delta v = \frac{\lambda}{T} (2)$$

With c_0 being the speed of light and B being the total signal bandwidth in (1), while $\lambda = c_0/f_c$, where f_c is the carrier frequency.

OFDM signals consist of orthogonal parallel subcarriers. The whole signal will be.

$$x(t) = \sum_{m=0}^{M-1} \sum_{n=0}^{N-1} s_{mN+n} \exp(j2\pi f_n t) (3)$$

Where s_{mN+n} is a complex modulated symbols, N is the number of subcarriers, M is the number of consecutive symbols, f_n is the individual frequency of subcarriers and

$$f_n = \frac{n}{T}. (4)$$

So after inserting (4) in (1) and (2) we will get

$$\Delta r = \frac{c_0 T}{2N}, (5) \Delta v = \frac{c_0}{MT f_c} (6)$$

From (5) and (6) followed that more subcarriers we have, less range resolution less relative velocity we will get.

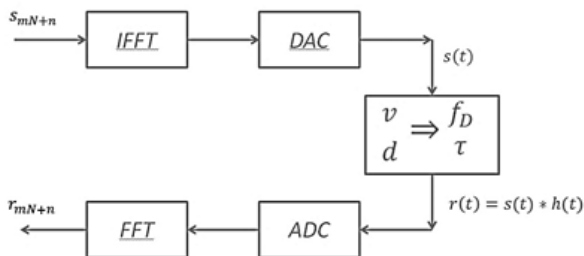


Fig. 1. OFDM Implementation by FFT

OFDM transmission and reception can be implemented with Fast Fourier Transformation (FFT). As shown in Fig. 1, just performing an

Inverse Fast Fourier Transform with the s_{mN+n} symbols, and converting the data from digital to analog we will get $s(t)$ signal which will be transmitted from radar station. After reflecting from the target the echo signal gets a Doppler shift and time shift, which occur because of the velocity of the target and its distance from the radar station. The received signal will be the convolution of transmitted signal and the impulse response. So to get the velocity and the distance of the target from the radar station firstly we should get the Doppler shift and the time shift, which are in the pulse response of our signal. By the way looking on Fig. 1, it looks very easy to implement and OFDM translation and reception by FFT, and it should work, because each block on the transmit site has its corresponding inverse on the receive site, so all the data should be perfectly recovered if our blocks will work perfectly. We must satisfy the condition of orthogonality. In the picture it is hidden in the cover of Fourier transform theory, that's why erroneously it looks very easy to make it work correctly.

3.One target OFDM Radar Simulation using Constellations

Representing OFDM signal in constellations we found a very pretty results. In Fig. 2.a it is shown the transmitted signal constellation, while in Fig 2.b we can brightly see what kind of changings get the signal after reflecting from a target

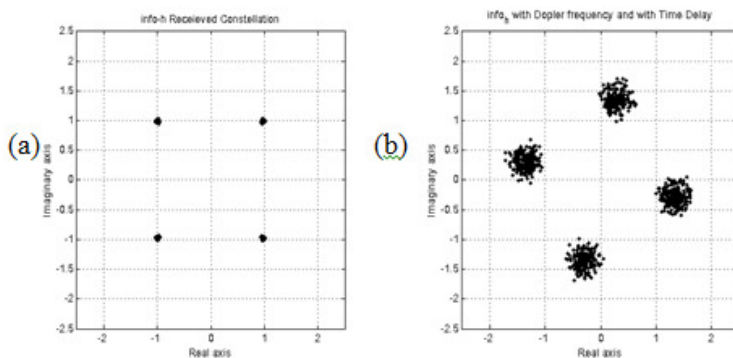


Fig. 2. (a) The constellation of transmitted OFDM signal, (b) The constellation of received OFDM signal with the existence of Doppler shift and time delay

In Fig. 2.b we see that after the impact of Doppler shift and time delay our constellation graph have been rotated and scaled. Simulation results shown that the Doppler shifting make our graph both to rotate and scale. On the other hand the time shifting just rotate the graph. So it is obvious that first we must get the Doppler shifting and consequently the velocity of target and only then time delay and its corresponding distance from the radar station. In Fig. 3.a is pictured the received signal without

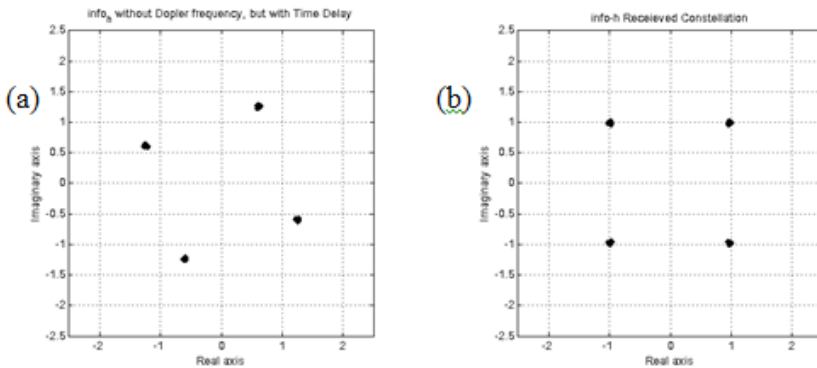


Fig. 3. (a) The constellation of received OFDM signal with the absence of Doppler shift and with the existence of time delay, (b) The constellation of received OFDM signal with the absence of Doppler shift and time delay

Doppler shifting and in Fig. 3.b is the received echo signal without Doppler shift and without time delay. So during these two operations we have got two major measures and then we can easily determine both the velocity and the distance of target form the Radar station.

$$v = \frac{f_D c_0}{2f_r}, (7) \quad d = \frac{c_0 \tau}{2}. (8)$$

Simulations were done in Matlab environment. The velocity range took from 2-62 km/h, the carrier frequency $f_c = 16$ GHz. The frequency bank of filters had a step $\Delta f_D = 60$ Hz. So it is found the Doppler velocity value $f_D = 1.44$ MHz and consequently from (7) we got the velocity of simulated target velocity $v = 48.6$ km/h.

In the same way, after getting the target velocity, we will make passed our signal trough the time delay bank of filters and get the distance between the radar station. The maximum distance from radar station is taken` $d_{max} = 8 \text{ km}$. $\Delta d = 300 \text{ m}$ a. Taking into account that $\tau = 2d/c_0$, we will get $\tau_{max} = 52 \mu\text{s}$ and $\Delta\tau = 2 \mu\text{s}$. Simulations gave the result $\tau = 40 \cdot 10^6 \text{ s}$, and from (8) we got the distance of the target from Radar station $d = 6 \text{ km}$.

Conclusion

As we stated before this method it useful with one target, but it has another positive side. We saw that in the received signal's constellation, if we have only the rotation around the axis, that's mean that the target from which we get the reflected signal isn't moving, and only when we have an additional scaling with the rotation, in that case only we can assume that we have a movable target. So constellations are also useful to vary movable and unmovable targets from each other. What belongs to using an OFDM signals in radar applications, the results also gave the thinking that they are perfectly applicable and are able to solve many problems. Our next investigation will be the using OFDM signals in addition with LFM signals, which is promising to solve the problems with multi target radars.

References

- [1] N. Levanon, "Multifrequency complementary phase-coded radar signal", IEE Proceedings Radar, Sonar and Navigation, vol. 147, no. 6, pp. 276-284, Dec. 2000.
- [2] C. Sturm, E. Pancera, T. Zwick, W. Wiesbeck, "A novel approach to OFDM radar processing", Radar Conference, 2009 IEEE , Publication, pp 1 – 4, 2009.
- [3] Y. L. Sit, C. Sturm, T. Zwick, "Doppler Estimation in an OFDM Joint Radar and Communication System", Microwave Conference (GeMIC), German, pp 1 – 4, 2011.

Analysis of Feeding Systems for Double Reflector Spherical Antennas

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The large meaning and development in a modern science and engineering have got large antenna systems intended for research of the Universe and maintenance to deep space communication with the space ships. Thus, antennas operating with high accuracy in microwave range of wavelengths with large reflecting surfaces are required. That made to overstep from parabolic reflectors to new types of antenna systems with superior basic parameters, as high accuracy, shortest wavelength (mm range), large effective area (S_{eff}), low self noise temperature (T_{sn}), high sensitivity (relation $S_{\text{eff}}/T_{\text{sn}}$), high efficiency of the aperture area (up to 0.7) and others. One of best designs for large antenna systems is the system «Herouni Mirror Radiotelescope». It includes an unmovable main spherical mirror and movable correcting small mirror. In order to use the surface of the antenna possibly efficient, a feed, suitable for the respective wavelength is required. In this article different types of feeding systems will be analyzed, and their appliance, as well as advantages and disadvantages will be discussed.

1. Introduction

In Radioastronomy, as well as in Deep Space Communication, where the microwave ranges of wavelengths is usual, Reflector Antennas gained a big spread. Especially popular are the parabolic antennas, both single and double reflector-systems of Cassegrain and Gregory. Nevertheless, by the construction of such parabolic antennas crucial are: the weight, hardness of the constructing materials, costs, etc. These factors put limitations on parabolic antennas due to the deformation of materials. Thus, parabolic antennas are inefficient in mm wavelengths. As the experience shows, the double reflector spherical antennas with fixed main reflector are free from these

limitations and have a quantity of advantages in comparison with parabolic antennas.

The idea of creating large spherical double reflector antennas for radio-astronomy and deep space communication with unmovable main spherical mirror and mobile correcting mirror was first suggested by P. M. Herouni in 1958. Since that time the elaboration of the theory of antennas of this type has started. In 1985 the construction of a large spherical double reflector antenna, called radio-optical telescope ROT - 54/2,6 (Fig.1), was completed by the efforts of Radiophysics Research Institute (RRI/now – National Institute of Metrology, Yerevan).



Fig.1. General view of the ROT-54/2.6 Double Reflector Spherical Antenna

In order to work as a high-efficiency antenna, a double reflector spherical antenna (DSA) should have a special feeding system. That could be either a horn antenna, or an array antenna as a feed. The issue of the optimal feeding of the secondary correcting reflector is covered in Section II. Different types of single feeds are described, as well as results of research of those patterns are provided.

There are theoretical elaborations of hybrid antennas-parabolic antennas with Phase Array Antennas well-known in the Literature.

The Section III is about constructing of Hybrid DSA with a fast swinging of the beam in the small angle of the field of sight, using two types of arrays.

For expansion of the effective Surface (S_{eff}) of the antenna supplementary linear feed (SLF) could be used, located in the focus of the antenna. The issues of designing, construction, installment and

experimental research of such a SLF in combination with the main feed of the radio telescope ROT-54/2.6 are covered in Section IV.

The Conclusion is invited for making a comparative analysis between those feeding systems.

2. Different types of antennas suitable as a main feed

2.1. General

Peculiarity of Spherical Antennas [1-4] is the absence of the focal point due to the geometry of the sphere. The maximum of the sum of Energy is distributed on a surface, called *caustic*.

There are two possibilities: either to have a correcting secondary reflector or a linear feed at half of the radius along the axis of the reflector. In the second case the elementary irradiators, which are distributed along the surface of the linear feed, collect the energy, reflected from the sphere. In the first case the correcting reflector is reflecting rays from its surface to the focal point. They will be thoroughly referred in this Section.

2.2. Three feeds

As a feed for a DSA could be used an antenna, which creates a spherical phase front in its aperture, which has small sizes and weight, so that installing it in the aperture won't shadow the latter too much. Depending on the wavelength varies also the type of the feed. Three new feeds [6] for a wavelength 3cm was designed, constructed and experimentally researched: open end waveguide, open end waveguide with a Teflon sleeve and corrugated horn (Fig. 2).

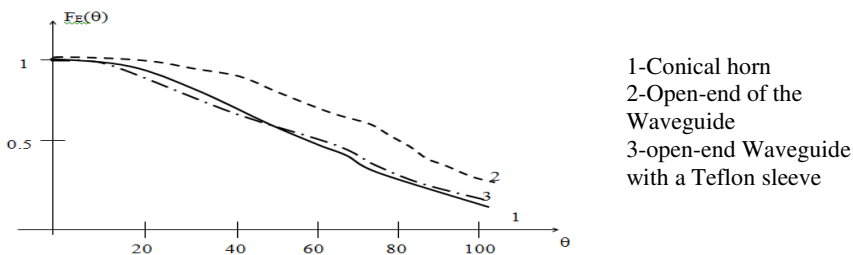


Fig. 2. Measurement results of patterns of single feeds for DSA

In the next figure is the sketch of the corrugated horn, which was constructed from a styrofoam ($n=1.14$).

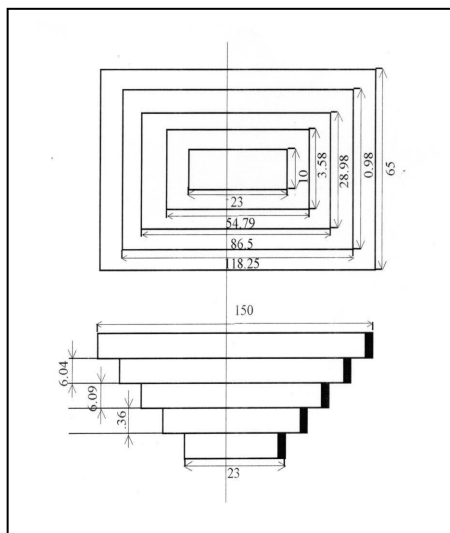


Fig. 3. Sketch of the corrugated horn

2.3. Patterns of the ROT-54/2.6 with a different single antenna feeds

The next Fig.4. is summarizing the results of measurements, represented by means of patterns of constructed feeds.

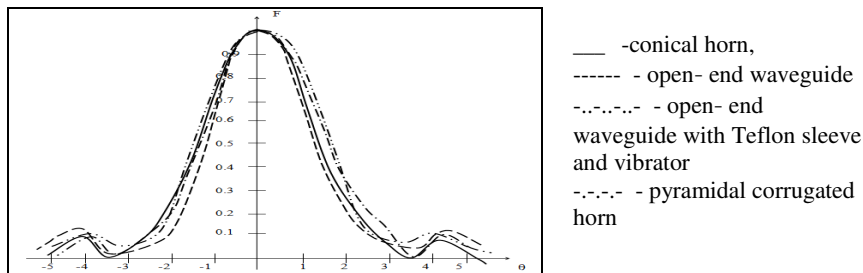


Fig.4.Comparison of patterns of ROT with different feeds constructed

As it can be seen, there is only a slight difference between the patterns of open-end Waveguide and of a Waveguide with a Teflon

sleeve and Vibrator. Moreover, the patterns of these feeds are worse, than those of the antenna with conical horn as a feed. The patterns of the corrugated horn have slight advantages: the level of sidelobes is less, and the width of the main beam is more close to the calculated one. The patterns of ROT-54/2.6, have been measured using the geostationary Satellite “Horizont VI”.

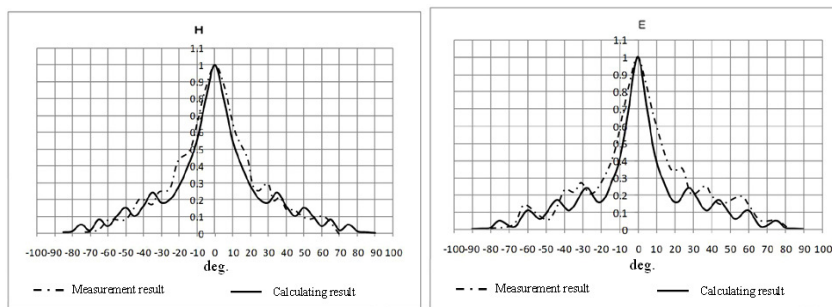
3. Antenna arrays as feeds for correcting reflector

3.1. The frame SWA

The results of measurements and calculations of the frame Slotted Waveguide Array Antenna (SWA) pattern respectively on horizontal and vertical surfaces are presented in Fig. 5.

3.2. Reconfigurable SWA

There has been sketched and made reconfigurable, multi-beam SWA radiator with elliptic polarization for 3cm-length operation wave for ROT 54/2.6 antenna in RRI base laboratory in 2009 [11]. The radiator consists of three frame SWA parts (Fig. 6).



a) horizontal surface

b) vertical surface

Fig. 5 Radiation pattern of frame SWA

For the combined operation of the frame SWA parts there has been designed a supporting construction. Each part has its own power system. Such a selection of the power system enables to provide comutational, simultaneous conducting and receiving as well as multi-beam operation mode.

Frame SWA parts can simultaneously work with similar as well as different frequencies this way providing the receiving of co-independent beams in the antenna aperture. The analysis of the measurement results makes it clear that the RP of each SWA part provides the required RP view of the radiator for Double mirror spherical antennas (DSA).



Fig. 6. Reconfigurable multi-beam SWA

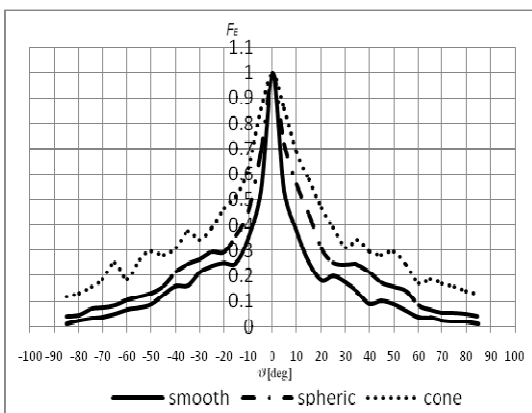


Fig. 7. Radiation pattern of Reconfigurable multi-beam SWA

The supporting construction provides for the simultaneous operation of the SWA parts along the central axis. Due to this peculiarity we provide the receiving of smooth, spherical and cone surfaces of the radiation. As a result of this we get different radiation parameters. The results of the measurements of the reconfigurable frame SWA RP in case of respectively smooth, spherical and cone surfaces are presented in Fig.7. The distance between the antenna parts is changed with 5mm-length steps and basic electrical parameters are

measured. In case of increasing of the dimension of the distance between the SWA parts the width of the pattern main lobe increases. Simultaneously the level of the SWA pattern side lobes increases in case of $119,7^{\circ}$ angle (a radiation angle in the secondary mirror edges). According to the received data the maintenance of the radiation optimal condition of ROT-54/2.6 antenna feed is received when the distance between SWA parts constitutes 25mm.

3.3. Fased array as a feed

Hybrid are called the antennas, in that the width of the main beam and the Coefficient of Directivity are defined by the parabolic reflector, but the deviation of the patterns are provided by a sophisticated feed. In this subsection the option of construction of such an antenna will be discussed: an array antenna with a spherical reflector. Often the limitations of parabolic antennas are caused not by a preciseness of refinement of the surface, but with the speed of controlling the beam.

If in the focus of the main reflector will be an Antenna Array-multi element feed, then in case of a certain position of the smaller reflector the scanning can be done by means of switching (commutation). In this case the antenna is similar to triple reflector antenna, because the Antenna Array is serving as the third antenna [8-10].

The scanning of beam of the whole antenna system could be succeeded by an electronic scan of the pattern of the Phase Array Antenna. The view of the feed have been constructed for ROT-54/2.6 is in the figure 8.

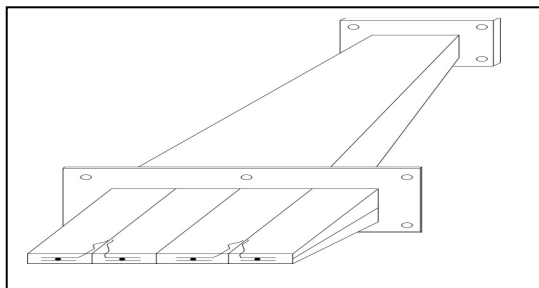


Fig. 8. The sketch of the waveguide made commutating AA.

The results of the measurements of the system showed, that the

patterns of the DSA-AA is able to move in space ~ 18.75 min of arc. The graph shows that in case of deviation of the beam ~ 18 min, the patterns are hardly changing, therefore the Coefficient of Directivity remains unchanged. It was calculated that in case of deviation of the beam to ~ 10 min the patterns of the DSA degrade only 15%, whereas the parabolic antennas experience 50% of degradation.

4. Supplementary Linear Feed (SLF) for ROT-54/2.6.

The non resonance slotted-waveguide with a distance between slots about $\lambda/2$ can also serve as a linear feed for a spherical double reflector antenna. The neighbor slots will excite with a little phase shift according to the wave in the waveguide. For making the phase shift smaller the slots should be made in a chess order. The exact coordinates of slots was calculated, and SWA was designed to function in wavelength $\lambda=20\text{cm}$. SLF of the type slotted waveguide was constructed for the antenna of ROT-54/2.6 and worked in combination with a main feed of a secondary reflector. The slotted-waveguide is hanged along the axe of the main reflector below the second reflector (Fig. 11) [5,7].

For exiting an electromagnetic wave in the waveguide a wire-waveguide transition section was designed according to relevant calculations. The matching in the line was reached by changing the length of phase rotating waveguide section. Hence, the radio signal, incoming to the radiometer, should consist of two components: from the main and supplementary feeds.

Thus, the SLF has patterns, one of the lobes of which is in a plane, directed to the surface of the main reflector. In the plane, perpendicular to the axis of SLF, the patterns are as the ones of Isotropic Irradiator. Exactly such pattern have to have the SLF in order to succeed his function: to collect that part of energy, this doesn't fall to the surface of secondary reflector. In fact the energy, which the rays, crossing the main axe of the DSA after reflection below the point "a" carry, isn't falling to the focus of the DSA (Fig. 9). By installing a SLF along the segment a-c below the focal spot (i.e. slotted-waveguide), and organizing the addition of signals from the main feed and SLF in radiometer, an expansion of the useful Surface will be gained. Thus, the SLF

irradiates a surface of the Main Reflector, which is unreachable for the main feed.

As the SLF is expanding of the antenna surface using factor (SUF), the important question is up to which level the coefficient could be improved. As it can be seen in the Fig.10, the max value for the SUF can be improved up to $0.707R_0$, whereas previously it was 0.6 (R_0 is the radius of the geometrical aperture of the Main Reflector). The usage of the SLF expanded to Radius of the acting aperture from 0.6 to 0.702 . That made the SUF increase by 17.8% .

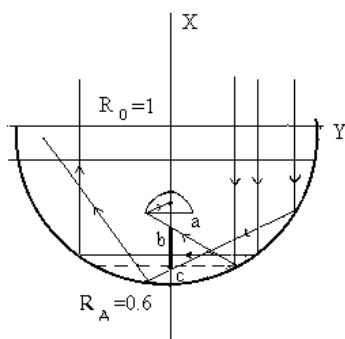


Fig.9. The geometric-optical scheme of a DSA

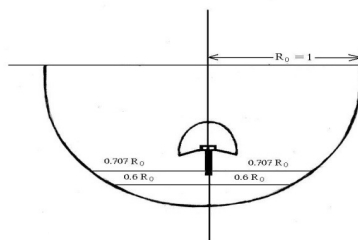


Fig.10. Expansion of the acting surface of the ROT-54/2.6

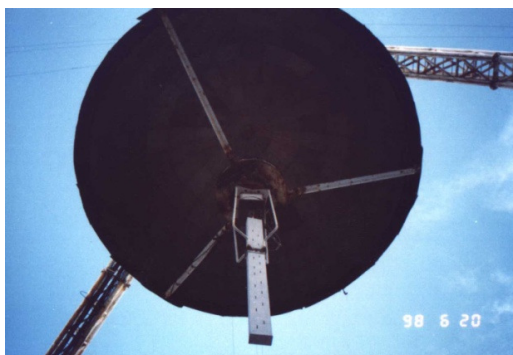


Fig. 11. The SLF, installed on ROT-54/2.6

5. Conclusion

Due to the hard requirements to the antennas used in Deep space communication, where mm range of wavelengths are common, Double

Reflector Spherical Antennas have a couple of considerable advantages against parabolic antennas. In order to enjoy these advantages possibly suitable feeding system should be applied. Therefore in this work feeding systems were compared, and their perspective in solving optimization problems was analyzed. As the measurements show, even a slight change of the patterns of the feed causes a significant change of the amplitude distribution in the aperture. From the feeds, constructed to serve as a main feed, the “one end waveguide” and “open end waveguide with Teflon sleeve” had similar patterns, whereas the feed of type “conical horn” had little better characteristics. The patterns of the corrugated horn had several faintly marked advantages: the side-lobes were lower and the main beam was narrower.

With the use of the suggested multi-beam, reconfigurable frame SWA with an elliptic polarization of radiation field the deformation deviations conditioned by the motion of the secondary reflector can be minimized hereby providing equal amplitude distribution in DSA aperture. It is an optimal radiator for ROT- 54/2.6 antenna. And the antenna surface using factor can be corrected by 87% with the use of SWA-radiator in case of the radiator deviation from the focus location caused by different deviation deformations which in their turn are conditioned by ROT-54/2.6 antenna movement.

In case of use of SLF it is possible to gain 17.8% better SUF without any noticeable degradation of other parameters. Moreover, it was concluded that the construction of hybrid DSA-Antenna Array will improve the antenna control, pointing, as well as tracing the target even further, which will bring a strong advantage against parabolic antennas.

6. References

- [1] P. Herouni, Theoretical problems of doublemirror spherical radiotelesopes. Report at the XI General Assembly of Internat. Astronomical Union, San-Francisco, 8 p., USA, 1960.
- [2] П. Геруни, К расчету больших сферических двухзеркальных антенн, Вопросы радиоэлектроники, XII, N 11, с. 179-197, Москва, 1963.

- [3] П. Геруни, Вопросы расчета сферических двухзеркальных антенн, Радиотехника и электроника, т.19, N1, с. 3-12., Москва, 1964.
- [4] A. K. Head, A new form for a giant radio telescope, Nature, Vol. 179, p. 692-693, April 6, 1957
- [5] A.Sargsyan, Supplementary Linear Feed for “Herouni Glassy Telescope”-s Antenna, XXVI URSI General Assembly, Canada, August, 1999.
- [6] Ա.Սարգսյան, Երկհայելանի սֆերիկ անտենաների ճառագայթիչներ, ՀՊՀՀ տարեկան գիտաժողովի նյութերի ժողովածու, 1999:
- [7] А.Саргсян, Конструирование, изготовление, измерение ДН и установка ДЛО на РОТ-54/2.6, В сб. Трудов ежегодной конференции ГИУА, 2001.
- [8] A.Sargsyan “An opportunity research of the hybrid double reflector spherical antenna with fast scanning of a beam in the field of the sight”, Transactions of IEEE APS/URSI Conference, 2003.
- [9] A.Sargsyan, Architectural Design of the Multiple Beam Multiple Feed Spherical Reflector Antenna Terminal, Proceedings of 29th ESA Antenna Workshop on Multiple Beams and Reconfigurable Antennas, ESTEC, Noordwijk, 2007.
- [10] A.Sargsyan, V.A.Minasyan, H.Abrahamyan, Modelling of Double Reflector Spherical Antennas, IRPhE-50 International conference, - 2010.
- [11] С. Эйрамджян, Г. Абраамян, Синтез и тестирование волноводно-щелевой антенной решетки с использованием оборудования National Instruments и LabVIEW// Образовательные, научные и инженерные приложения в среде LabVIEW и технологии National Instruments // Сборник трудов VIII Международной научно-практической конференции.- М.: РУДН, 2009.- с. 245-248.

CONTENTS

Terahertz time-domain spectroscopy (THz-TDS) applied to non-destructive dielectric quality evaluation of industrial materials	5
<i>Seizi Nishizawa, Mitsuo W. Takeda, Masahiko Tani</i>	
Frequency tunable meta-material element based on superconducting quantum interference filter	16
<i>K. Y. Constantinian, A. V. Shadrin, G. A. Ovsyannikov, and S. E. Bankov</i>	
THz radiation in the LiNbO ₃ crystal tapered waveguide	22
<i>A. S. Nikoghosyan, H. P. Roeser, H. S. Hakobyan, A. Bohr, D. T. Haslam, J. S. López, M. Stepper</i>	
Generation of difference frequency radiation in the field of few-cycle laser pulse propagating in periodically-poled Lithium Niobate	33
<i>G. D. Hovhannisyan</i>	
THz radiation in LiNbO ₃ rectangular plate waveguide	44
<i>A. S. Nikoghosyan and H. S. Hakobyan</i>	
Detection of laser radiation in optically transparent ferromagnet	52
<i>H. S. Hakobyan, A. H. Makaryan, V. M. Mekhitarian, and V. R. Tadevosyan</i>	
Microwave-phonic receivers – New conception in radio-communication	61
<i>H. V. Baghdasaryan, T. M. Knyazyan, A. A. Hakhoumian, and M. Marciniak</i>	
S-Band 800 Watt Gallium Nitride based pallet amplifier for air traffic control radars	70
<i>A. Barsegyan, F. Boueri, and V. Thangam</i>	
Investigation of Carbon-PEEK piezoelectric properties by using a near-field scanning microwave microprobe	77
<i>T. Abrahamyan, K. Karapetyan, C. Jerome, D. Trichet, A. Perronnet, G. Berthiau, A. Babajanyan, Ch. Kim, K. Lee, and B. Friedman</i>	
Design of miniature high power waveguide ceramic filter with improving characteristics	84
<i>J. Kim, Y. Jang, S. Kim, A. Babajanyan, K. Lee</i>	
Reduction of the clutter in non-coherent LFM CW radars	90
<i>A. Hakhoumian, T. Zakaryan, N. Poghosyan, H. Avagyan, E. Sivolenko, and Kh. Tovmasyan</i>	
Increasing of the signal to noise ratio by using threshold devices	96
<i>H. R. Avagyan, H. S. Haroyan, A. H. Makaryan, V. R. Tadevosyan</i>	
Effect of annealing temperature on Copper-Phthalocyanine thin films investigated by a near-field microwave microscope	106
<i>A. Babajanyan, J. Kim, and K. Lee</i>	
Preliminary results of C-, Ku-, and Ka-band multi-frequency radiometric	

measurements of clear air and clouds brightness (antenna) temperatures <i>A. K. Arakelyan, A. K. Hambaryan, V. K. Hambaryan, and A. A. Arakelyan</i>	117
Propagation of a harmonic in time and an arbitrary polarized plane electromagnetic wave through a non-regular one-dimensional media <i>A. Zh. Khachatryan, A. G. Alexanyan, A. S. Avanesyan, and N. A. Alexanyan</i>	127
On the measurements of UHF antenna parameters by two antennas technique <i>V. H. Avetisyan and A. A. Sargsyan</i>	149
The measurements of parameters of UHF antenna array radiators <i>M. V. Markosyan, V. H. Avetisyan, A. A. Hovhannisyan, A. K. Aharonyan, G. Z. Sugyan, A. I. Petrosyan,</i> <i>A. A. Manukyan, E. A. Harutyunyan, and W. H. Tonoyan</i>	156
The problem of automation of electromagnetic compatibility calcula- tions for microwave-link equipment <i>V. H. Avetisyan, A. H. Nikoghosyan, V. V. Grigoryan</i>	162
Distant infrared spectral analysis of industrial hot gas ejection in atmosphere <i>R. Asatryan, S. Asatryan, N. Khachatryan, and L. Sukoyan</i>	166
Infrared optical/electronic method of monitoring environmental CO ₂ and H ₂ O vapors <i>R. Asatryan, S. Asatryan, N. Khachatryan, and L. Sukoyan</i>	173
Real-time monitoring of glucose concentration by a microwave biosensor <i>K. Lee, S. Kim, J. Kim, A. Babajanyan, and B. Friedman</i>	179
Change of physical-chemical properties tumoral DNA irradiated by low power millimeter waves <i>V. Kalantaryan, Yu. Babayan, and T. Qitabchyan</i>	187
Stability of bilayer lipid membrane under influence of low-intensity millimeter waves <i>V. Kalantaryan, V. Arakelyan, and H. Gevorgyan</i>	196
Method of cardiac activity extraction in L-Band CW radars <i>H. Avagyan, A. Hakhoumian, H. Hayrapetyan, N. Poghosyan, and T. Zakaryan</i>	201
The agro products sorting system based on degree of maturity <i>D. Zargaryan, R. Simonyan, A. Ghulyan, H. Pirumyan</i>	207
QPSK mapped OFDM signal processing on radar applications <i>Kh. Tovmasyan</i>	214
Analysis of feeding systems for double reflector spherical antennas <i>A. Sargsyan, H. Abrahamyan, Y. Alkhatib, G. Hakobyan</i>	219

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