ARCHIVES OF ACOUSTICS
13, 3-4, 237-253 (1988)

ANALYSIS OF NARROW-BAND DISPERSIVE INTERDIGITAL TRANSDUCERS BY THE SCATTERING-MATRIX METHOD FOR PERIODIC METAL STRIPS

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A model of the interdigital transducer of surface acoustic wave (SAW) is introduced on the strength of the scattering matrix for a single metal strip of the periodic system deposited on a piezoelectric substrate surface. The model describes properties of the transducer within the wide frequency band and takes into account the SAW reflection of electrical origin from the metal strips. The calculations performed for narrow-band dispersive filters show that the SAW reflections from the transducer electrodes distort the filter response not only as a result of the immediate effect upon the filter frequency characteristic, but also through the frequency-dependent change in the transducer admittance. A presented method for analyzing inhomogeneous systems may have some other applications, e.g. in a theory of superlattice.

Wprowadzono model przetwornika międzypalczastego akustycznych fal powierzchniowych wykorzystując macierz rozpraszania dla jednej elektrody periodycznego układu, położonego na powierzchni podłoża piezoelektrycznego. Model opisuje własności przetwornika w szerokim pasmie częstotliwości i uwzględnia odbicie AFP od elektrod, pochodzenia elektrycznego. Obliczenia, przeprowadzone dla wąskopasmowych filtrów dyspersyjnych wykazują, że odbicia AFP od elektrod przetworników zniekształcają odpowiedź impulsową filtru nie tylko wskutek bezpośredniego wpływu na jego charakterystykę, ale także poprzez zmianę admitancji przetworników, zależną od częstotliwości.

1. Introduction

The commonly used equivalent scheme of a pair of metal strips of an interdigital transducer (IDT) [1] is based upon the theory of bulk-wave transducers. That is why, in order to take account of the phenomena connected closely with SAW, this scheme had to be appropriately modified. Thus the SAW reflection has been considered from the metal strips in the transducer fundamental frequency-band [2] and on the harmonic frequencies [3]. The so-called element factor [4] is introduced that describes quantitatively the phenomenon of generating and detecting the surface wave by the transducer electrodes, this phenomenon being substantially different from the analogous electromechanical transformation for the bulk waves.

A different approach will be assumed below to the theory of interdigital transducers, this approach being based upon the strict theory of SAW. For instance, the above-mentioned element factor results naturally from the theory of periodic metal strips situated on a piezoelectric substrate surface [5] (the case of near-surface waves being considered in [6]).

The IDT model introduced in the present paper is based upon the scattering matrix for the periodic metal strips [7] as well as upon circuit theory. The reasoning that permits the theory [7] to be applied to the non-periodic metal strips within the wide frequency-band including the SAW Bragg reflection band, is similar to that in the considerations presented in [8], though the equivalent scheme will not be derived in the present paper as being non-physical for the surface waves. It will be shown that the IDT model based exclusively upon the scattering matrix for the metal strips is sufficient for analyzing the SAW transducers and filters.

In order to simplify matters it is assumed below that the transducer metal strips have a width equal to the spacing between them (this spacing can vary along the transducer). Account will be taken of the SAW reflection brought about exclusively by the electrical interaction of the conducting metal strip with the wave. The mechanical properties of the metal strips can be taken into consideration in the similar way as in Refs [9–11]. Bulk waves are left out of account.

As mentioned above, a strict theory exists only for the periodic system of metal strips. Extension of its results to the case of non-periodic transducers will be effected by comparing the appropriate results of the theory of non-homogeneous transmission lines, these results being shown in the next Section along with the results of the theory [7] shown in the Section after next. The resulting model of the transducer, as presented in Sections 4 and 5, has been used for analyzing the dispersive delay-lines, the analysis being similar to that presented in [12]. In contrast with [12], in Section 6 are analyzed the narrow-band filters with a considerably longer time response.

2. Non-homogeneous transmission line

We now consider a transmission line (Fig. 1) composed of three sections of different but homogeneous lines having the lengths $l/2$, l' and $l/2$. The impedances of these lines are R or R'. The wave velocity in these lines are denoted by v and v' (for simplification it is assumed that $l/v = l'/v'$ and $R \geqslant R'$).

The relation between voltage and current complex amplitudes on the line terminals is the following ($\alpha = \omega l/v$, ω – angular frequency)

$$\begin{bmatrix} i_{n+1} \\ e_{n+1} \end{bmatrix} = \begin{bmatrix} \cos(\tau) & -j\sin(\tau)/Z \\ -jZ\sin(\tau) & \cos(\tau) \end{bmatrix} \begin{bmatrix} i_n \\ e_n \end{bmatrix}, \tag{1}$$

where the phase shift τ is expressed by $(j=\sqrt{-1})$

$$e^{-j\tau}=\frac{(R+R')^2}{4RR'}[\cos(2\alpha)-\Delta^2-j2\sin(\alpha)(\cos^2(\alpha)-\Delta^2)^{1/2}], \quad (2)$$

$\Delta=(R'-R)/(R'+R)$, while the wave-impedance is

$$Z=R[(\cos(\alpha)-\Delta)/(\cos(\alpha)+\Delta)]^{1/2}. \quad (3)$$

This is at the same time the impedance of a semi-infinite chain composed of successively linked two-port networks described by the relation (1).

It should be noted that the wave impedance Z may have a complex value. This occurs when the wave propagating an infinite chain composed of the networks, as depicted in Fig. 1, is in synchronism with the periodic inhomogeneity of such a transmission line. Then $\tau\approx\pi/2$, and as results from (2), the wave number of the wave $r=\tau/(2l)$ also has a complex value ($2l=l+l'$ is the period of the transmission line over the length of which period the wave changes its phase by τ). The frequency range for which the wave number is complex is called a stop-band.

Next we consider the chain (transmission line) composed of different two-port networks presented in Fig. 1 (with different parameters R, R' etc). The network with the number n is described by (1) on assuming the values τ_n, Z_n instead of τ, Z, respectively. We introduce the notion of the wave α^+ propagating to the right and having the wave number r and the complex amplitude a^+, as well as that of the wave α^+ propagating to the left and having the wave number $-r$ and the amplitude a^-. These amplitudes are defined for each network of the chain in the following standard fashion (the wave phases are referred to the centers of networks)

$$\begin{aligned} i_n&=(\alpha^+\exp(j\tau_n/2)-\alpha^-\exp(-j\tau_n/2))/\sqrt{Z},\\ e_n&=(\alpha^+\exp(j\tau_n/2)+\alpha^-\exp(-j\tau_n/2))\sqrt{Z}. \end{aligned} \quad (4)$$

The relations obtained from (1) and (4), between the amplitudes of the waves α^+ and the waves α^- in the successive networks of the non-homogeneous transmission line under consideration can be written in the form (the equality of currents and voltages on the boundaries of the adjoining networks should be taken into account)

$$\alpha_n^+=T_{n-1,n}\alpha_{n-1}^+\exp[-j(\tau_{n-1}+\tau_n)/2]+\Gamma_{n,n-1}\alpha_n^-\exp(-j\tau_n), \quad (5)$$

and similarly for the wave α_n^-.

The phenomena described by the above relation and accompanying the wave propagation in non-homogeneous transmission line are the following:

- change in the wave phase after transmission of the network from one edge to the other is τ_n,
- wave transmission through a connection between the adjoining networks. T_{LE} is the transmission coefficient for the wave passing from the network L to the network E,

$$T_{LE}=2(Z_LZ_E)^{1/2}(Z_L+Z_E)^{-1}. \quad (6)$$

– wave reflection, the reflection coefficient of the wave propagating in the network L from the boundary with the network E is

$$\Gamma_{LE} = (Z_E - Z_L)(Z_L + Z_E)^{-1}. \tag{7}$$

The relations (5)–(7) are also valid in the stop band, where Z_n has a complex value (however note that the above interpretations may not be correct). That is because (4) has been introduced entirely formally (the square-root sign in (6) is of no importance, it must, however, be consistently the same as in (4)). Relations (6) and (7) define the elements of the scattering matrix for the successive networks of the inhomogeneous transmission line.

In conclusion we consider the wave reflection from the boundary between the homogeneous line of the impedance $R = 1$, and the semi-infinite line periodically inhomogeneous, composed of the networks depicted in Fig. 1, and having the wave impedance Z. The reflection coefficient of the wave is

$$\alpha^+/\alpha^- = \Gamma = (1-Z)/(1+Z). \tag{8}$$

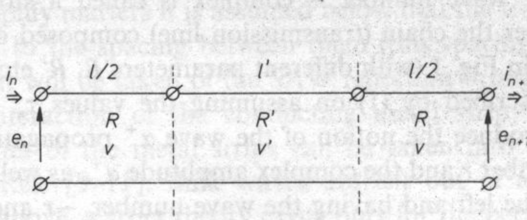


Fig. 1. Transmission line

3. Selected results of the SAW theory

In Ref [7] the periodic system is considered of the metal strips with the period Λ (for simplification the metal strips under study have the width $\Lambda/2$) and situated on the surface of the half-space of a piezoelectric substrate characterized by $\Delta v/v$, v (SAW velocity) and ε_{eff}(effective surface permittivity of the substrate). All the metal strips are grounded with the exception of one having the number m, to which a potential of an angular frequency ω is applied, this potential in the complex form being equal to $V_m \exp(j\omega t)$.

The current that flows into the metal strip of the number n and of the length W is

$$J_n = -j\omega W C_{|n-m|} V_m + W I_n, \tag{9}$$

where the first component is defined exclusively by the dielectric properties of the substrate, and the other component is connected with the surface accustic wave generated at the metal strip m. The quantities occurring in (9) are defined by simplified relation (the simplifications concerning mainly I_0 and E_f)

$$C_n = \frac{1}{\pi}\varepsilon_0(1+\varepsilon_{\text{eff}})\left(n^2-\frac{1}{4}\right)^{-1}, \tag{10a}$$

$$I_n = YV_m \exp(-j|n-m|\tau), \tag{10b}$$

$$Y = E_f/Z, \tag{10c}$$

$$E_f = \omega\varepsilon_0(1+\varepsilon_{\text{eff}})4\pi(\Delta v/v)s(1-s)\sin(\pi s), \tag{10d}$$

$$Z = [(\omega-\omega_3)/(\omega-\omega_1)]^{1/2}, \tag{10e}$$

$$\tau = r\Lambda, \tag{10f}$$

$$r = K/2+\frac{1}{v}[(\omega-\omega_1)(\omega-\omega_3)]^{1/2}, \tag{10g}$$

$$\omega_1 = Kv/2, \tag{10h}$$

$$\omega_3 = \omega_1(1-s(1-s)\Delta v/v), \tag{10i}$$

$$s = \frac{1}{2}\omega/\omega_1. \tag{10j}$$

These relations are valid for $0 < s < 1$, i.e. in the fundamental frequency band.

In the periodic system of metal strips described among others in [7], there occur two waves, a forward-travelling wave of the wave number r and the amplitude α_0 and a backward-travelling wave of the wave-number r–K and the amplitude α_{-1}. These amplitudes are connected by the relation

$$\alpha_{-1}/\alpha_0 = (1-Z)/(1+Z), \tag{11}$$

where Z is defined by (10e).

The relation (11) is made use of in [13] for constructing the boundary conditions for a bounded system of metal strips. Let the wave $\alpha^+ \exp(-jkx)$, $x < 0$, $k = \omega/v$, propagating on the free surface of a piezoelectric, falls on the area covered by periodic metal strips ($x > 0$). In the area $x < 0$ account must also be taken of the reflected wave $\alpha^- \exp(jkx)$, while in the area of the metal strips two waves exist $\alpha_0 \exp(-jrx)+\alpha_{-1}\exp j(K-r)x$ (the factor $\exp j\omega t$ is omitted in this paper). The boundary conditions for $x = 0$ are [13]

$$\alpha_0 = \alpha^+, \quad \alpha_{-1} = \alpha^-,$$

hence, on consideration of (11) one obtains the relation (8). The analogy with the transmission line discussed above is obvious, particularly when noticing that Z defined by (3) can be written in the form similar to (10e) for $\alpha \approx \pi/2$.

4. Mathematical model of metal strips

Analogies between surface waves and waves in the transmission line have been taken into account when introducing the equivalent scheme for transducer metal strips (Refs [1, 8, 14]). It may be stated here that (5) describes the SAW propagation under the system of metal strips, if the parameters occurring there are defined by the relations (10), (6) and (7). In the case of a non-periodic system of metal strips, in the relation (10) Λ_n should be assumed as equal to the "local period" of this system (which is justified only for a nearly periodic system of metal strips, which take place e.g. in the case of narrow-band dispersive filters). For the consecutive metal strips the parameters Y_n, Z_n, and τ_n will occur.

When supplying the metal strips from an external source is taken into account this leads to the modification of the relation (5). It should then be assumed that under the metal strip a change occurs in the amplitudes of the waves α^+ and α^- by a certain value dependent upon the potential of the metal strip. Similarly, the current flowing to the metal strip depends upon the waves travelling under it. The relevant considerations can be found in [7] and their result, on consideration of (5), is contained in the following relations

$$\alpha_n^+ = T_{n-1,n}(\alpha_{n-1}^+ + V_{n-1}(Y_{n-1}/2)^{1/2})\exp[-j(\tau_{n-1}+\tau_n)/2] + \Gamma_{n,n-1}[V_n(Y_n/2)^{1/2} + T_{n+1,n}(\alpha_{n+1}^- + V_{n+1}(Y_{n+1}/2)^{1/2})\exp[-j(\tau_n+\tau_{n+1})/2]]\exp(-j\tau_n), \quad (12)$$

$$I_n = (\alpha_n^+ + \alpha_n^-)(2Y_n)^{1/2} + Y_n V_n,$$

(similarly for α^-, the square-root sign is to be chosen consistently as in (6)).

The relations are illustrated in Fig. 2. They describe the propagation, generation and detection of SAW in the area of the metal strips, the relation (9) taking additionally into account the mutual capacitance of the metal strips. In order to complete the description the value should be adopted yet of the wave impedance on the free surface that limits the system of the metal strips. The discussion in Section 3 justifies the adoption of this value equal to 1. The error of leaving out of account the wave reflection on the border of two areas with different wave numbers ($r > k$) may be usually disregarded.

It should be emphasized that the above depicted model (Fig. 2) takes no account of either the bulk waves or of the elastic properties of metal strips, and the only cause of wave reflections from the metal strips is the short-circuit effect for electric potential under metal strips (this is so-called $\Delta v/v$ reflection). However, both Ref [9] and Refs [10] and [15] indicate that the mechanical properties of the metal strips may be included into the model through appropriate correction of the parameters that occur there.

In conclusion, it is worth noting that the case under study in [7] and [16] of a bounded periodic system of metal strips a part of which have an established non-zero potential may be described as well by the model in Fig. 2 [14], [17].

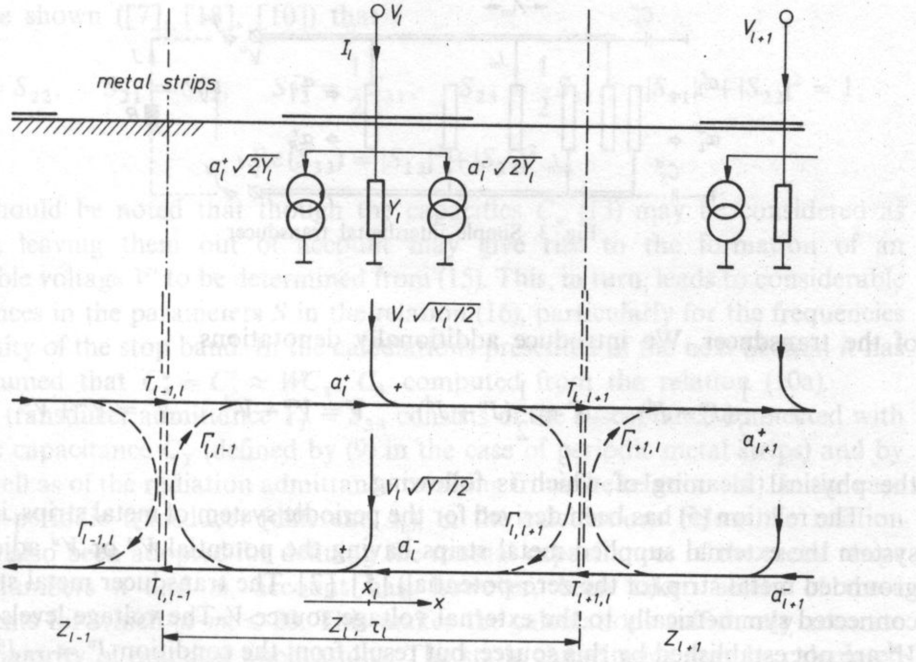


Fig. 2. Metal strip model resulting from the scattering matrix

5. Analysis of interdigital transducers

On the basis of the model introduced in the preceding Section the scattering matrix will be derived below for a simple nearly periodic ($\Lambda_n \approx \Lambda_{n+1}$) transducer (Fig. 3) with the aperture W (the width of the metal strip of the number n is equal to $\Lambda_n/2$).

The quantities $\alpha_{L,R}^{+,-}$ shown in Fig. 3 denote the complex amplitudes of acoustic waves on the left-hand (L) and the right-hand (R) side of the transducer and propagating to the left ($-$) or to the right ($+$). $J^{u,d}$ and $V^{u,d}$ denote both the complex amplitudes of the current flowing to the transducer bars (u $-$upper one, d $-$ lower one) and the potentials of these bars, respectively. In addition it is convenient to introduce total SAW amplitudes as $A = \alpha\sqrt{W}$ with indices L, $+$ and $-$, of the same meaning as in the case of α ($|A|^2/2$ is the power of the SAW beam of the width W).

In accordance with Fig. 3 one obtains

$$J^{u,d} = \sum_{l=l^{u,d}} J_l + j\omega C_x^{u,d} V^{u,d} \tag{13}$$

where $l^{u,d}$ are the numbers of the metal strips connected to the upper or the lower bar

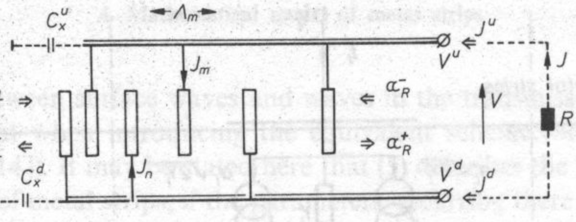


Fig. 3. Simple interdigital transducer

of the transducer. We introduce additionally denotations

$$J = \frac{1}{2}(J^u - J^d), \quad J' = \frac{1}{2}(J^u + J^d), \quad V = V^u - V^d, \quad V' = V^u + V^d \tag{14}$$

the physical meaning of which is following.

The relation (5) has been derived for the periodic system of metal strips, in which system the external supplied metal strips having the potential V^u or V^d adjoins the grounded metal strip (of the zero potential) [5], [7]. The transducer metal strips are connected symmetrically to the external voltage source V. The voltage levels V^u and V^d are not established by this source, but result from the condition $J^u = -J^d$, that is

$$J' = 0. \tag{15}$$

Consequently, the potential V' denotes the magnitude of the non-symmetrical supply to the transducer metal strips, $V' = 0$ when $V^u = -V^d$. In this connection, in the case under investigation of a bounded system of metal strips, situated on a substrate of a finite thickness, the capacity is to be taken into account between the metal strip and the filter housing, this capacity corresponding to the grounded metal strips discussed previously. These are capacities $C_x^{u,d}$ taken into consideration in the relation (13) and in Fig. 3.

The relations (12) which constitute the set of equations for the amplitudes of acoustic waves, currents and potentials of the metal strips in each section of the transducer of a length Λ_n, may be solved recurrently. One obtains relations between the amplitudes $A_{R,L}^{+,-}$ and the quantities as given in (14) these relations being similar to the relation presented below for J'

$$J' = \alpha A^+ + \beta A^- + \gamma V + \delta V'.$$

The condition (15) permits eliminating V'. Ultimately one obtains

$$\begin{aligned} A_R^+ &= S_{11}A_L^+ + S_{12}A_R^- + S_{13}V, \\ A_L^- &= S_{21}A_L^+ + S_{22}A_R^- + S_{23}V, \\ J &= S_{31}A_L^+ + S_{32}A_R^- + S_{33}V. \end{aligned} \tag{16}$$

It can be shown ([7], [18], [10]) that

$$S_{11} = S_{22}, \quad S_{21} = S_{12}^*, \quad S_{13} = \frac{1}{2} S_{31}, \quad S_{23} = \frac{1}{2} S_{32}, \quad |S_{11}|^2 + |S_{22}|^2 = 1,$$

$$\mathrm{Re}(S_{33}) = |S_{13}|^2 + |S_{23}|^2.$$

It should be noted that though the capacities C_x (13) may be considered as parasitic, leaving them out of account may give rise to the formation of an appreciable voltage V' to be determined from (15). This, in turn, leads to considerable disturbances in the parameters S in the relation (16), particularly for the frequencies in a vicinity of the stop band. In the calculations presented in the next Section it has been assumed that $C_x^u = C_x^d \approx WC_1$, C_1 computed from the relation (10a).

The transducer admittance $Y_T = S_{33}$ consists of the susceptance connected with the static capacitance C_T (defined by (9) in the case of periodic metal strips) and by (13), as well as of the radiation admittance resulting from the relation (12). In the case of nearly-periodic transducer (different Λ_n), in the calculations below the relation (10a) has also been adopted for defining the mutual capacitances of the metal strips having numbers n and m, account has, however, been taken solely of these components for which $|n-m| \leqslant 20$. This makes the calculations sufficiently accurate for the majority of practical applications. The static capacitance may be calculated separately (neglecting the SAW, that is for $\Delta v/v = 0$) by applying (9), (10), (13) as well as the relation (16) which is reduced then to $J = j\omega C_T V$.

6. Numerical analysis of narrow-band chirp filters

The dispersive filter under analysis (Fig. 4) consists of two transducers, a broad-band one with $n = 3$ pairs of splitted metal strips ($m = 4$ of a metal strip falls to a wave-length), and a dispersive one (with linear dispersion) characterized by the $B = 4$MHz pass-band and duration of the time response $T = 8\mu$ (or $T = 16\mu s$) the center frequency of the filter being 30 MHz. Analysis has been performed of dispersive transducers with $m = 2$ metal strips to a wave-length and with sectional metal strips ($m = 3$ or $m = 4$ metal strips to a wave-length). The calculations have been carried out for filters with an aperture $W = 4$ mm on the substrate of SiO_2YX or $LiNbO_3$YZ. In the calculations the load impedance R has been assumed to be equal to the impedance of the generator supplying the filter ($R = 50\,\Omega$ or $R = 1\,\Omega$).

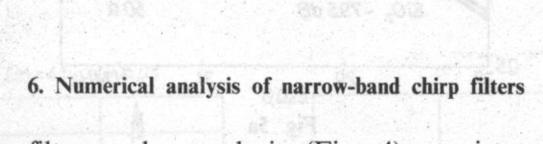


Fig. 4. Configuration of the down-chirp filter

The filter frequency response $H(f)$ is

$$H_l(f) = 2RS_{l3}^{(1)} S_{3l}^{(2)} [(1+S_{33}^{(1)} R)(1+S_{33}^{(2)} R)]^{-1}, \tag{17}$$

where the superscripts (1) and (2) denote the parameters of a broad-band transducer and a dispersive one, respectively, the subscript $l = 1$ for the down-chirp filter, while $l = 2$ for the up-chirp filter. TTS is left out of account in (17). The impulse response $h(t)$ and the compressed pulse $c(t)$ is calculated with the aid of the FFT algorithm:

$$h_l(t) = F^{-1}(H_l(f)), \quad c(t) = F^{-1}(H_1 H_2).$$

Fig. 5a depicts the calculated frequency characteristic of the delay line composed of two broad-band transducers for a substrate of lithium niobate or quartz. The insertion losses, amounting at center frequency to 32. 7 dB and 79. 5 dB increase by about 1.5 dB to a transducer at the band edges. The same transducers have been assumed in the calculations presented below.

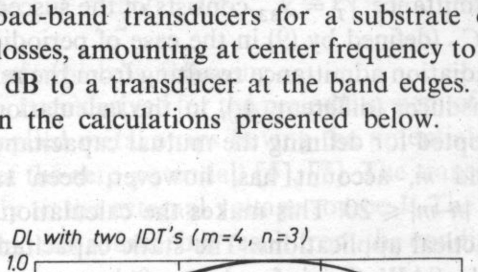
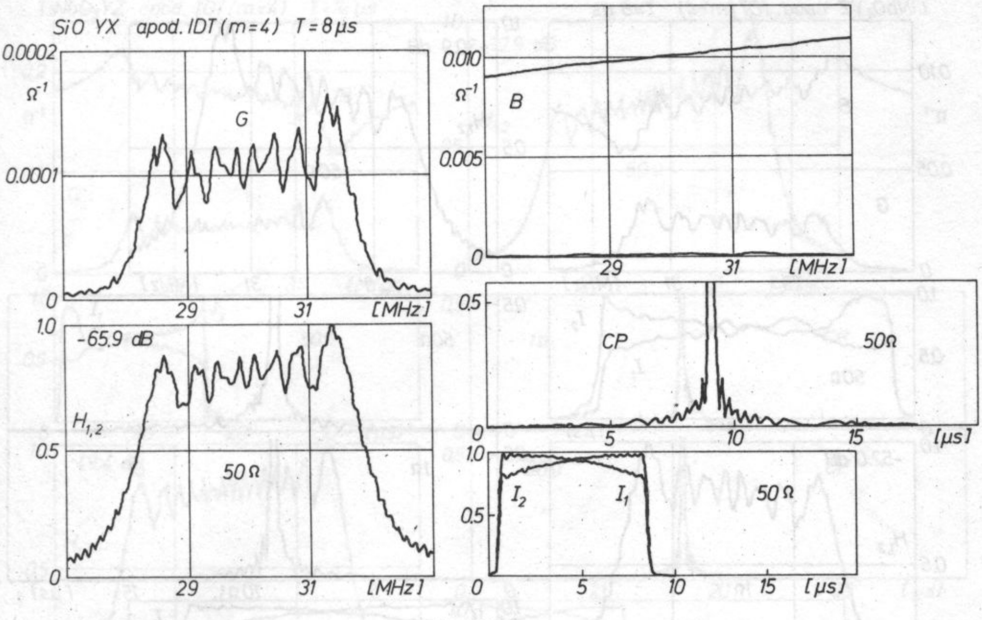


Fig. 5a

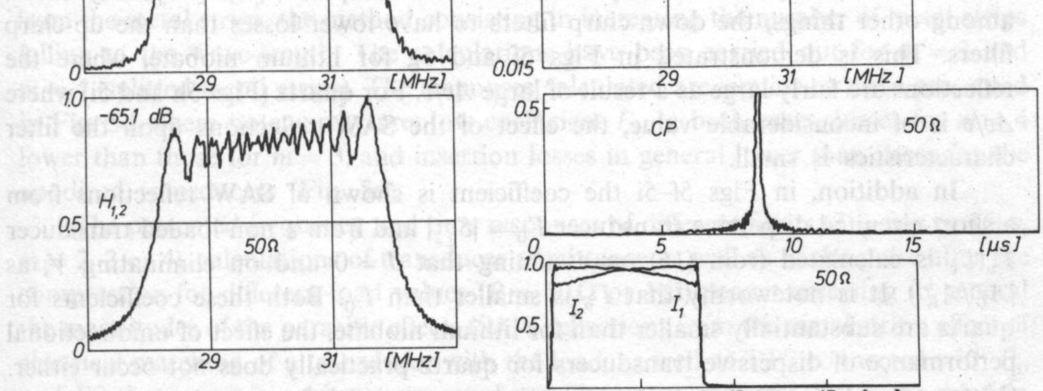
For later comparisons Figs 5b and 5c present in a certain sense ideal responses of dispersive filters in which dispersive apodized transducers of periodic metal strips are applied (Ref [19]). The apodization is calculated on the basis of the relation (7.2) given in [7] ($f_0 = 60$ MHz was applied, that corresponds to four ($m = 4$) metal strips on a wave-length at center frequency of the filter). The synchronous SAW reflections from the metal strips in the pass-band of these transducers do not occur. The frequency and time responses presented are rounded as a result of the above-discussed form of the frequency characteristic of the broad-band transducer. The conductance G and the susceptance B of the dispersive transducers are also shown, as well as the shape of compressed pulse.

The calculated responses of similar filters on a substrate of $LiNbO_3$ (Figs 5d and 5e) are more distorted. This results from an appreciably higher admittance value of the dispersive transducers ($G+jB$), this value being dependent upon frequency. By comparing the filter responses calculated at different loads R it is seen that the effect of the factor $(1+RS_{33}^2)$ occurring in (16), must be considerable for $R = 50\,\Omega$. This



SiO YX apod. IDT (m=4) T=8 μs
0.0002
Ω⁻¹
G
0.0001
0
29
31
[MHz]
0.010
Ω⁻¹
B
0.005
0
29
31
[MHz]
1.0
-65.9 dB
H1,2
0.5
50 Ω
0
29
31
[MHz]
0.5
CP
50Ω
0
5
10
15
[μs]
1.0
I2
I1
50 Ω
0.5
0
5
10
15
[μs]

Fig. 5b



SiO2 YX apod. IDT (m=4) T=16 μs
0.00030
Ω⁻¹
0.00015
0
29
31
[MHz]
0.025
Ω⁻¹
0.020
0.015
29
31
[MHz]
1.0
-65.1 dB
H1,2
0.5
50Ω
0
29
31
[MHz]
0.5
CP
50Ω
0
5
10
15
[μs]
1.0
50Ω
I2
I1
0.5
0
5
10
15
[μs]

Fig. 5c

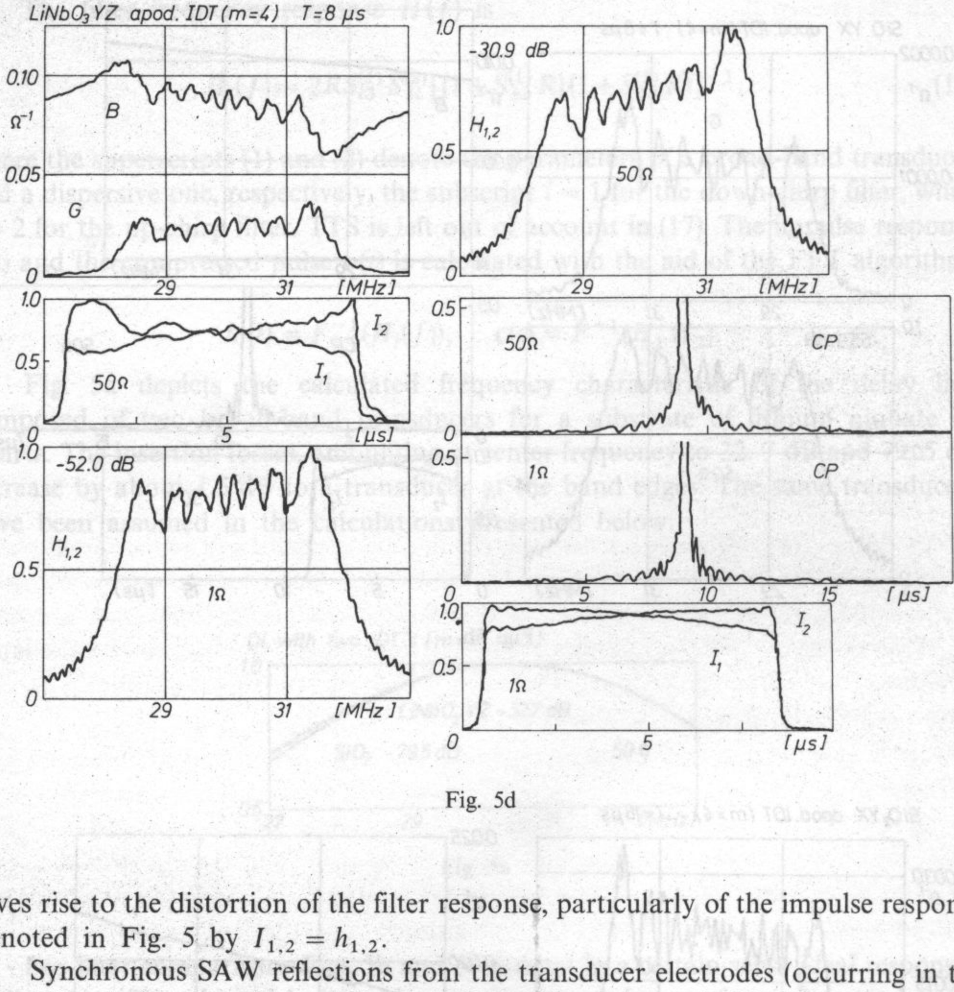


Fig. 5d

gives rise to the distortion of the filter response, particularly of the impulse response denoted in Fig. 5 by $I_{1,2} = h_{1,2}$.

Synchronous SAW reflections from the transducer electrodes (occurring in the case of $m = 2$) may substantially distort the filter response [2], [12], they cause, among other things, the down-chirp filters to have lower losses than the up-chirp filters. This is demonstrated in Figs 5f and 5g for lithium niobate, where the reflections are fairly large as a result of large $\Delta v/v$. For quartz (Figs 5h and 5i) where $\Delta v/v$ is of inconsiderable value, the effect of the SAW reflections upon the filter characteristics is small.

In addition, in Figs 5f–5i the coefficient is shown of SAW reflections from a short-circuited dispersive transducer $\Gamma_0 = |S_{12}|$ and from a non-loaded transducer Γ_v(Γ_v is calculated from (16) on assuming that $J = 0$ and on eliminating V, as $|A_L^-/A_R^+|$). It is noteworthy that Γ_v is smaller than Γ_0. Both these coefficients for quartz are substantially smaller than for lithium niobate, the effect of unidirectional performance of dispersive transducers for quartz practically does not occur either. (Note: only $\Delta v/v$ reflections are taken into account in the paper).

The filter responses as presented in Figs 5f, 5g are considerably distorted in

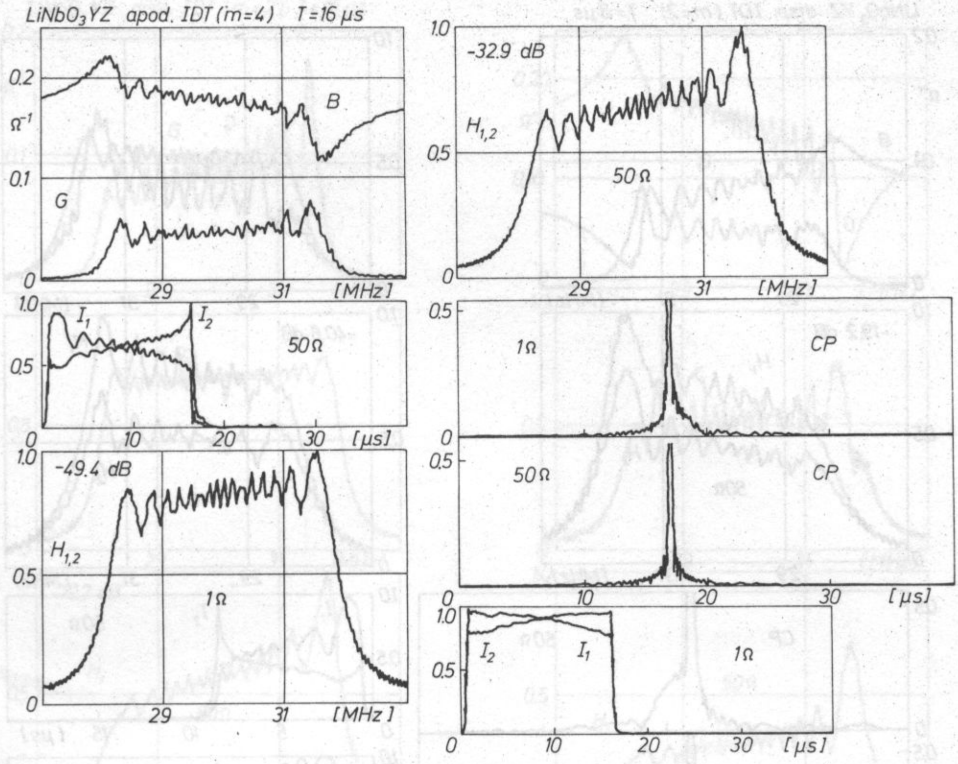


Fig. 5e

relation to "ideal" responses (Figs 5b, 5c). The distortion for $R = 50\,\Omega$ results from the effect of SAW reflections from the metal strips as well as from the above discussed effect of high admittance of a dispersive transducer. Hence the compressed pulse, calculated for $R = 50\,\Omega$ is of an appreciably lower quality than for $R = 1\,\Omega$.

Figs 5j, 5k depicts the work of the method for eliminating the SAW reflections from the metal strips, the method consisting in increasing the number of metal strips falling to the wave-length. The calculations have been carried out for $m = 3$ and $m = 4$ (splitted metal strips). The responses calculated are similar to these presented in Fig. 5d where noteworthy are: low coefficient Γ_0 in both cases, losses for $m = 4$ lower than those for $m = 3$, and insertion losses in general lower than those for the apodized transducers (Fig. 5d).

The presented frequency and time responses of different filters ($T = 8\mu s$ or $16\mu s$, $m = 2$, 3 or 4), calculations of transducer admittance as well as of the pulse after the compression for different load values ($R = 50\,\Omega$ or $1\,\Omega$) permit assessing in general the magnitudes of the parasitic effects (SAW reflections from the metal strips, effect of electrical matching of a transducer with the load, as well as TTS connected with Γ_0 and Γ_v) that occur in arbitrary narrow-band filters (the case of $s < 1$ in the relation (10)).

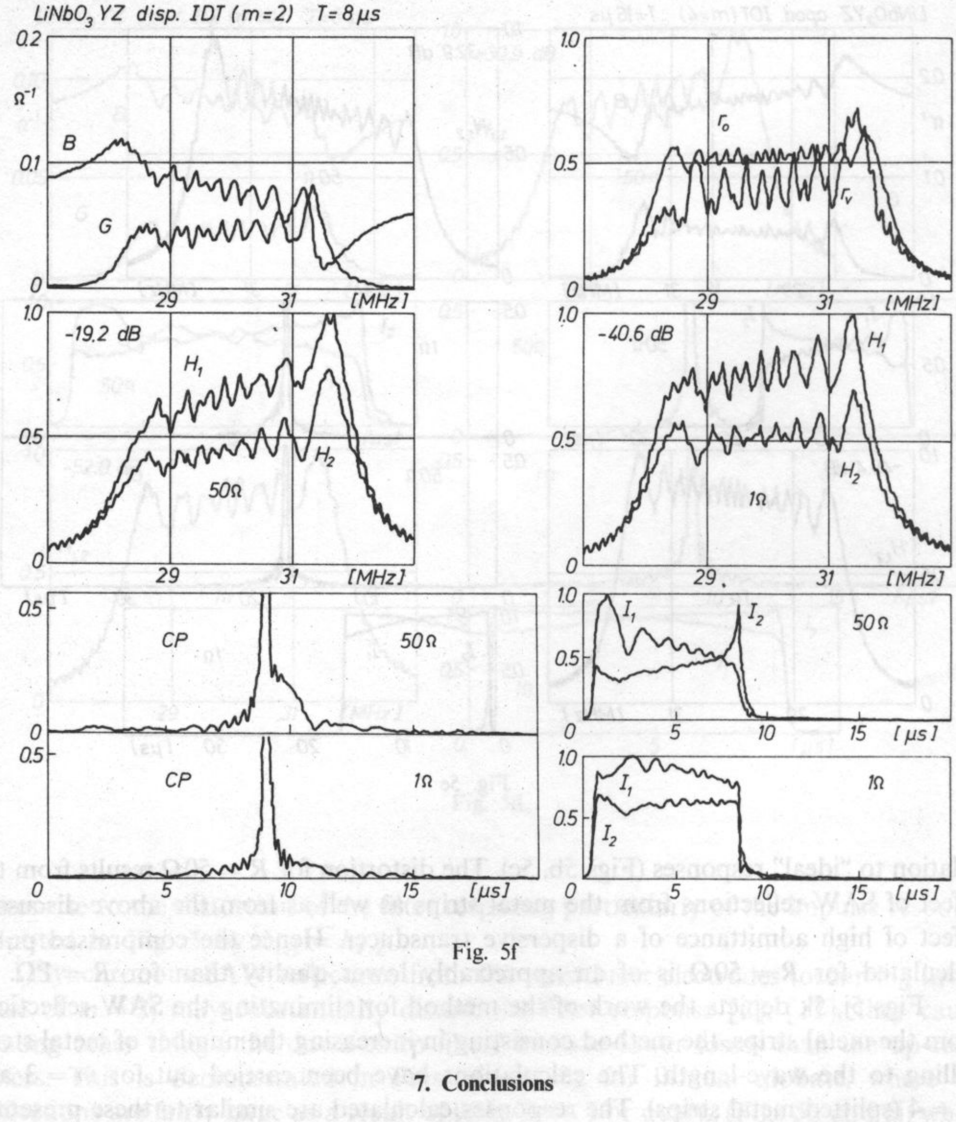


Fig. 5f

7. Conclusions

The above simple model of the interdigital transducer electrodes based upon the scattering matrix, has been derived from the strict theory of periodic metal strips. The model does not include any artificial non-physical parameters and may be readily employed for calculating the scattering matrix of interdigital transducers. The model is open to further modifications such as taking account of mechanical properties of metal strips and bulk waves, and even diffraction [9], [7]. Similarly it is possible to broaden the frequency band of the correctness of the model (for $s > 1$). Application of the above introduced model to the analysis of unidirectional transducers is obvious.

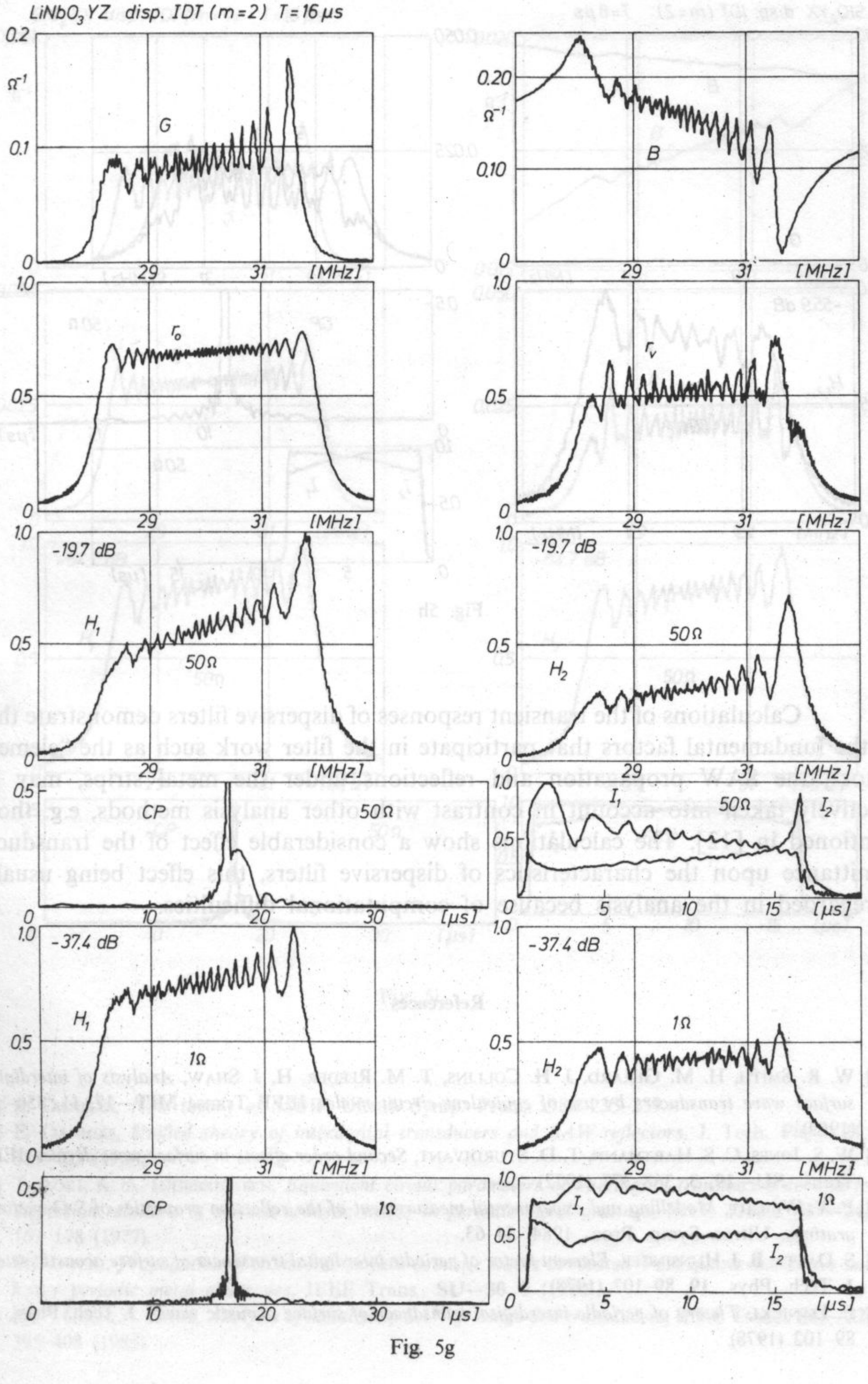

LiNbO3YZ disp. IDT (m=2) T=16 μs
G
B
Ω⁻¹
[MHz]
r₀
rᵥ
-19.7 dB
H₁
H₂
50Ω
CP
I₁
I₂
[μs]
-37.4 dB
1Ω

Fig. 5g

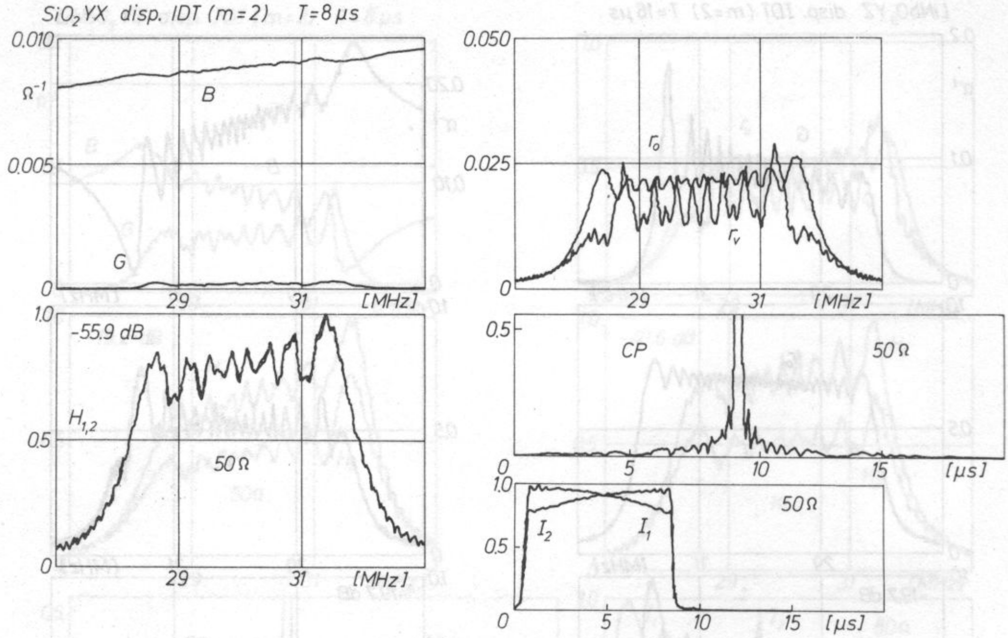


Fig. 5h

Calculations of the transient responses of dispersive filters demonstrate that all the fundamental factors that participate in the filter work such as the "element factor", the SAW propagation and reflections under the metal strips, may be effectively taken into account in contrast with other analysis methods, e.g. those mentioned in [12]. The calculations show a considerable effect of the transducer admittance upon the characteristics of dispersive filters, this effect being usually disregarded in the analysis because of computational difficulties.

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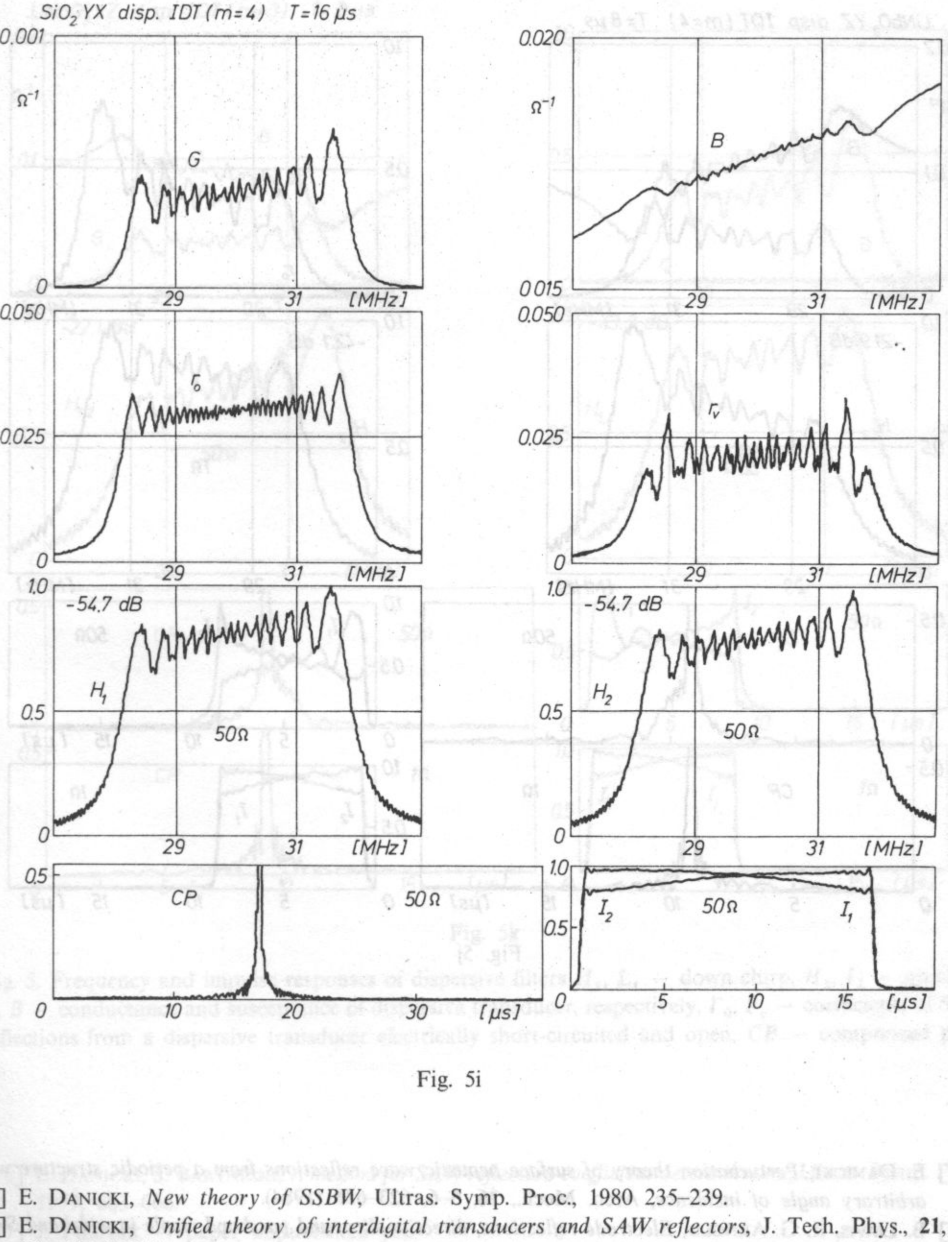

Fig. 5i

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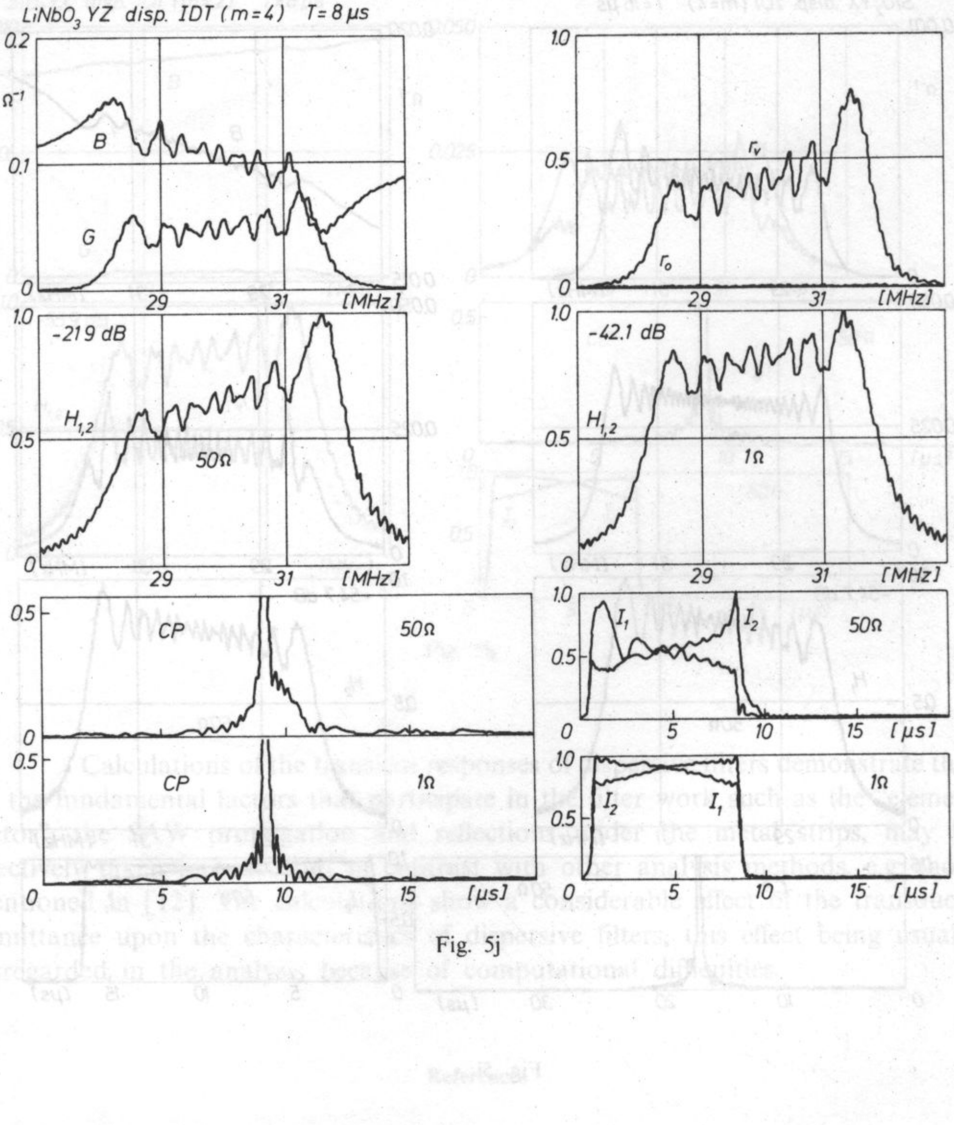

Fig. 5j

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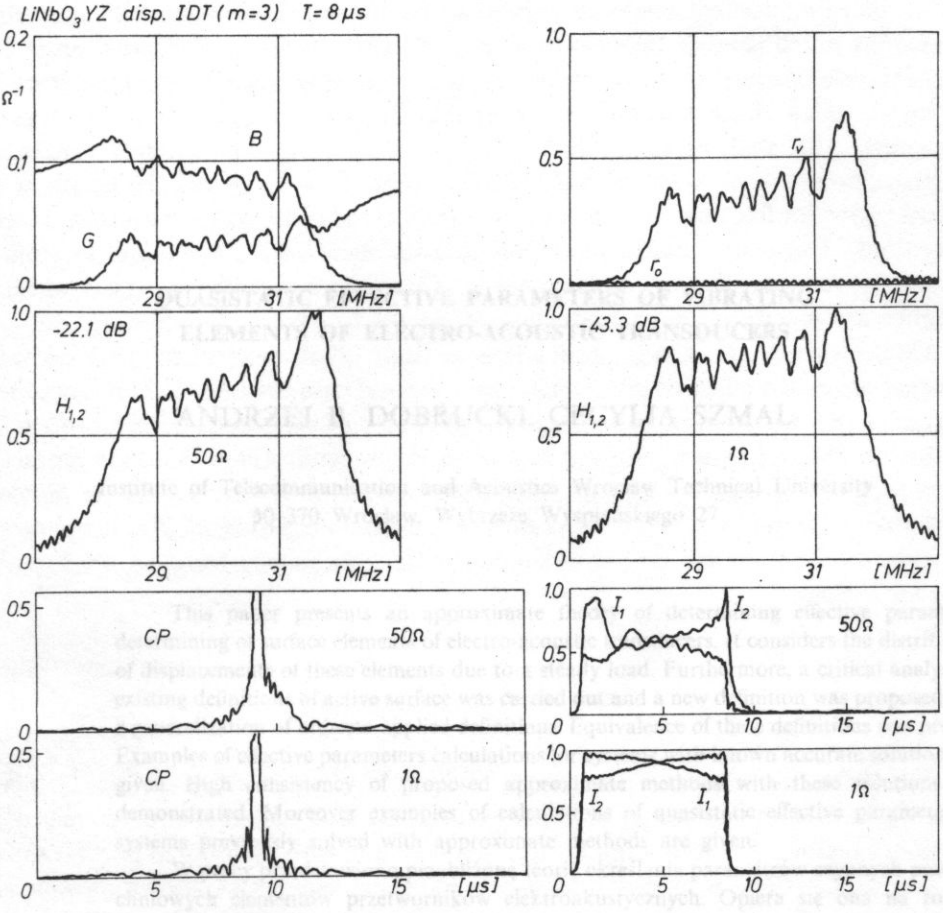

Fig. 5k

Fig. 5. Frequency and impulse responses of dispersive filters H_1, L_1 – down-chirp, H_2, I_2 – up-chirp, G, B – conductance and susceptance of dispersive transducer, respectively, Γ_0, Γ_v – coefficients of SAW reflections from a dispersive transducer electrically short-circuited and open, CP – compressed pulse

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Received on August 19, 1988.