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MINIATURIZED IF CDMA SAW FILTERS BASED ON THREE-CHANNEL STRUCTURES

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Abstract - The size reduction is a permanent tendency in the development of IF SAW filters for CDMA mobile phones. The implementation of three-channel structures is known as a method, which provides reducing IF-filter size. In three-channel SAW filters, the triple transit signal serves as useful one. The direct acoustic signal is compensated due to the 120^o phase shifts between the channels. However, a strong transverse acoustic coupling occurs between the adjacent channels. It results in a strong distortion of the frequency response of such filters, degradation of selectivity and increased phase ripples in the pass-band. In this paper, a modified structure of the three-channel filter, which is free of the above shortcoming, is described. The modified structure has been used in filters for CDMA system: 210.38 MHz, 220.38 MHz and 85.38 MHz in packages SMD 5.0x5.0 mm^2, 5.0x7.0 mm^2 and 9.0x5.0 mm^2, respectively.

1. INTRODUCTION

IF SAW filters for the CDMA system must meet specific requirements: relatively high pass-band width in combination with narrow transmission-band width, high selectivity, linear phase and fairly low insertion loss. For mobile phones, packages of minimum size are also needed.

In recent years, several SAW IF filter techniques have been developed to reduce size; for example, the Z-path filter concept [1]. To reduce size with respect to inline structures, the Z-path concept has been combined with recursive design technique resulting in a Z-path filter. Both SPUDTs of the filter have weighted SAW excitation and reflection. The reflection function changes the sign in order to have longitudinally coupled resonant cavities and to provide steep slopes of the frequency response. Due to the folded propagation path, Z-path filters utilize the available chip length more efficiently than filters with inline structures. The longitudinally coupled acoustic cavities within the transducers increase the length of impulse response, providing more freedom of the design [2].

Another technique uses a two-channel structure with a reflector grating located between the SPUDTs in each channel. In this structure, the direct-transit signals are cancelled and the acoustic path is effectively folded [3]. In the generalized two-channel RSPUDT structure, the frequency response of each track is wider than the final response of the filter [4]. In the passband, the signals in the upper and lower channels are nearly in phase and add constructively, while in the stopband they cancel each other, thus providing improved selectivity of the filter. Thus, a possibility of keeping the filter bandwidth narrow, with fairly wide passbands in each channel, allows the reduction of the filter size.

In a three-channel filter described in [5,6], the phase shifts $2\pi/3$ between channels were introduced, by means of slightly different lengths of acoustic channels. As a result, the first-transit signals (or the direct acoustic signals - DAS), $H_i^D(\omega)$, cancel in such a structure, while the triple-transit signals (TTS) $H_i^T(\omega)$ add up in phase [5,6]. The SAW reflections from the transducers form the filter frequency response and yield the substantial reduction of the bandwidth, without increasing transducer length. However, the size reduction of such a filter increases by the transverse acoustic coupling (TAC) between closely adjacent channels [5,6], which distorts amplitude and phase characteristics of the filter. The present paper describes a modified three-channel structure free of this drawback.

2. MODIFIED THREE-CHANNEL STRUCTURE

Fig.1 shows schematically a possible layout of the modified three-channel structure. Each of acoustic channels I, II, III involves the input and output SPUDTs with an acoustic waveguide between them. As in the initial structure [5,6], SPUDTs are connected electrically parallel at the input and output. The initial differences between the effective electro-acoustic phases of the extreme channels I and III, with respect to that of the middle channel II are

$$\Delta\varphi^{e}_{II-I} = +2\pi/3 \pm 2k\pi \text{ and } \Delta\varphi^{e}_{II-III} = -2\pi/3 \pm 2k\pi, \quad (1)$$

respectively, where k =0, 1, 2, 3, 4 …. is an integer.
In the modified structure, the channels I and III have extra shifts of the electro-acoustic phases $\Delta\varphi^{k}_{II-I} = -\pi/6 \pm 2k\pi$ and $\Delta\varphi^{k}_{II-III} = +\pi/6 \pm 2k\pi$, with respect to the phase of the channel II. The phase shifts $\Delta\varphi^{e}_{II-I}$ and $\Delta\varphi^{e}_{II-III}$ occur because the output SPUDT 2 in channel I and the output SPUDT 6 in channel III are shifted in the direction of SAW propagation (X-direction) at distances $-\lambda/3 \pm k\lambda$ and $+\lambda/3 \pm k\lambda$, relative to the output SPUDT 4 in channel II, respectively. Here λ is the SAW wavelength at the

filter central frequency ω_0. Additional phase shifts $\Delta\varphi_{II-I}^{k}$ и $\Delta\varphi_{II-III}^{k}$ are due to the shift of channel I, as a whole, and of channel III, as a whole, at distances $S1 = -\lambda/12 \pm k\lambda$ and $S2 = +\lambda/12 \pm k\lambda$, respectively, in the X-direction relative to channel II. Thus, the distances between the transducers in the channels I, II, and III are l_1, $l_2 = l_1 + \lambda/3$, and $l_3 = l_2 + \lambda/3$ (Fig.1), respectively.

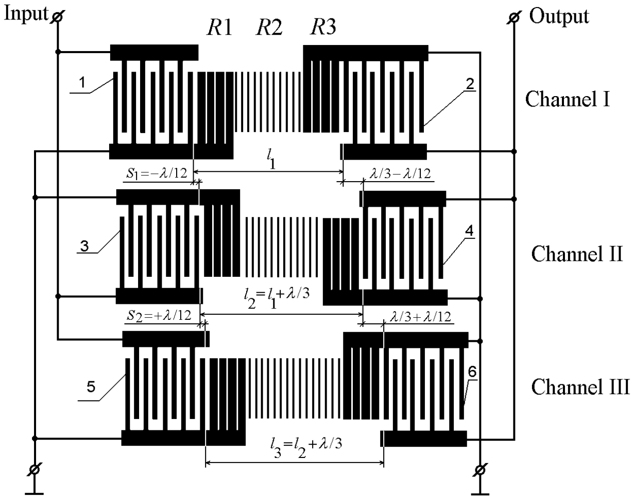

Fig.1. Structural scheme of the tree-channel filter with compensation of transversal acoustic coupling

In general, the phase shifts between the channels can be introduced using electric phase shifters. SPUDTs at the input and output of the filter can be connected in series to increase input and output impedances.

The polarity of the electrode connection with the potential and grounded bas bars in input 1,3,5 and output 2,4,6 SPUDTs are chosen such that

$$\begin{cases} H_3(\omega)\cdot H_2^*(\omega) = H_1(\omega)\cdot H_4^*(\omega); \\ H_5(\omega)\cdot H_4^*(\omega) = H_3(\omega)\cdot H_6^*(\omega); \\ H_1(\omega)\cdot H_2^*(\omega) = H_3(\omega)\cdot H_4^*(\omega) = H_5(\omega)\cdot H_6^*(\omega), \end{cases} \tag{2}$$

where $H_i(\omega)$ is the transfer function of the i-th SPUDT, the asterisk means complex conjugation.

In all, 12 combinations of the electrode connection polarities in the input and output SPUDTs can be used. Fig. 1 shows one of them: SPUDT 1 (-); SPUDT 2 (-); SPUDT 3 (+); SPUDT 4 (+); SPUDT 5 (-); SPUDT 6 (-).

Reflectionless waveguide grating $R2$ forms an acoustic waveguide which weakens TAC in the region between the input and output transducers. Reflector gratings $R1$ and $R3$ are used to adjust the electro-acoustic phase shift of each channel and to decrease phase ripples of the filter additionally.

To simplify the analysis of the three-channel filter, we use the scheme shown in Fig. 2a. Consider first the filter functioning without regard for TAC. The relative phase shift of DAS in channels I, II, and III is $2\pi/3$. Hence, the DAS transfer functions of these channels are

$$\begin{aligned} H_I^D(\omega) &= H_1(\omega)\cdot H_2^*(\omega)\cdot e^{-i\varphi_1} = H_{12}; \\ H_{II}^D(\omega) &= H_3(\omega)\cdot H_4^*(\omega)\cdot e^{-i(\varphi_1+2\pi/3)} = H_{34}; \\ H_{III}^D(\omega) &= H_5(\omega)\cdot H_6^*(\omega)\cdot e^{-i(\varphi_1+4\pi/3)} = H_{56}, \end{aligned} \tag{3}$$

where $\varphi_1 = \omega(l_1/V_e)$ is the SAW phase shift due to propagation over distance l_1; V_e is the SAW effective velocity.

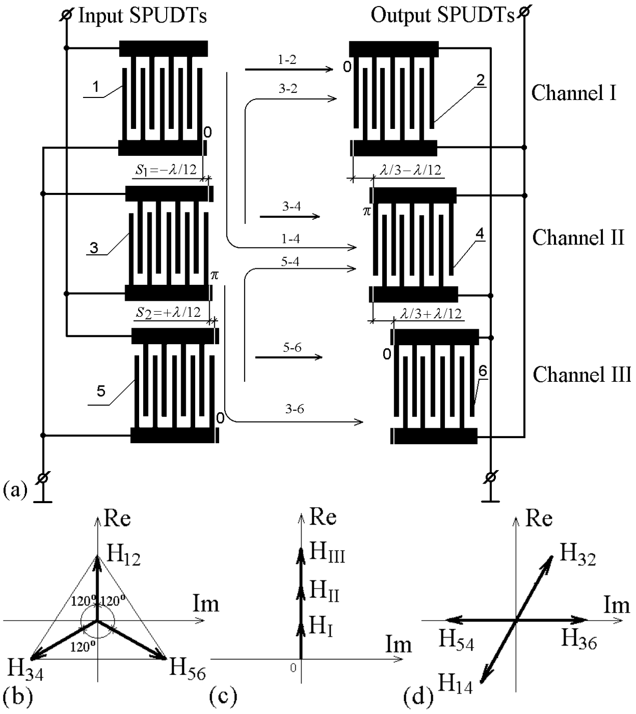

Fig.2. Simplified analysis of filter with transversal acoustic coupling: a - structure of filter; vector diagrams for summation of DAS - (b), TTS - (c), TAS - (d)

If we assume that transducers 1 - 6 are identical, then the DAS transfer function becomes

$$\begin{aligned} H_F^D(\omega) &= H_I^D(\omega) + H_{II}^D(\omega) + H_{III}^D(\omega) = \\ &H_i^2(\omega)\cdot e^{-i\varphi_1}[1 + e^{i2\pi/3} + e^{i4\pi/3}] = 0 \end{aligned} \tag{4}$$

Thus, the summation of DASs from each channel at the filter output results in their mutual cancellation (Fig.2b).

The transfer functions of channels I-III for TTS read as

$$\begin{aligned} H_I^T(\omega) &= H_1(\omega)\cdot H_2^*\cdot R_1(\omega)\cdot R_2(\omega)\cdot e^{-i3\varphi_1} = H_I; \\ H_{II}^T(\omega) &= H_3(\omega)\cdot H_4^*\cdot R_3(\omega)\cdot R_4(\omega)\cdot e^{-i3(\varphi_1+2\pi/3)} = H_{II}; \\ H_{III}^T(\omega) &= H_5(\omega)\cdot H_6^*\cdot R_5(\omega)\cdot R_6(\omega)\cdot e^{-i3(\varphi_1+4\pi/3)} = H_{III}, \end{aligned} \tag{5}$$

where $R_i(\omega)$ is the SAW reflection coefficient from i-th transducer. TTS from the channels add up constructively at the output so that a coherent electric signal appears (Fig.2c). Note that (3)-(5) hold when the polarity of the electrode connection for all transducers 1-6 is the same.

Now, using a simplified method, we analyze the filter with due regard for TAC. This coupling can occur both within the transducer regions and between them [7]. Let an input transducer, e.g., SPUDT 3 in channel II, generates

SAW. Due to transversal acoustic coupling the DAS bounded beam partially falls with the same phase and the relative amplitude P_d onto an adjacent transducer, e.g., SPUDT 1 in channel I; [8]. As a result, in channel I, in addition to direct wave 1-2, wave 3-2 also propagates, coming from channel II. In channel II, there propagate direct wave 3-4 as well as waves 1-4 and 5-4 from channels I and III etc. Taking into account only TAC between input transducers 1, 3, and 5, we can find the DAS transfer functions of channels I, II, and III as

$$H_I^D(\omega) = H_{12}(\omega) + H_{32}(\omega) = \\ = H_1(\omega)\cdot H_2^*(\omega) + P_d \cdot H_3(\omega)\cdot H_2^*(\omega)\cdot e^{-i(\varphi_1-\pi/6)} \; ; \qquad (6)$$

$$H_{II}^D(\omega) = H_{34}(\omega) + H_{14}(\omega) + H_{54}(\omega) = \\ = H_3(\omega)\cdot H_4^*(\omega)\cdot e^{-i(\varphi_1+2\pi/3)} + \\ P_d \cdot H_1(\omega)\cdot H_4^*(\omega)\cdot e^{-i(\varphi_1-\pi/6+\pi)} + \\ + P_d \cdot H_5(\omega)\cdot H_4^*(\omega)\cdot e^{-i(\varphi_1+2\pi/3-\pi/6)} \; ; \qquad (7)$$

$$H_{III}^D(\omega) = H_{56}(\omega) + H_{36}(\omega) = \\ = H_5(\omega)\cdot H_6^*(\omega)\cdot e^{-i(\varphi_1+4\pi/3)} + \\ + P_d \cdot H_3(\omega)\cdot H_6^*(\omega)\cdot e^{-i(\varphi_1+2\pi/3-\pi/6+\pi)} \; . \qquad (8)$$

According to Eqs. (6)-(8), there occurs both the mutual cancellation of the three DAS $H_{12}(\omega)$, $H_{34}(\omega)$, and $H_{56}(\omega)$ and the pairwise cancellation of the signals $H_{14}(\omega)$ and $H_{32}(\omega)$, $H_{36}(\omega)$ and $H_{34}(\omega)$ transmitted from the adjacent channels because of TAC (Fig.2d), i.e.:

$$H_{14}(\omega) + H_{32}(\omega) + H_{36}(\omega) + H_{34}(\omega) = 0. \qquad (9)$$

As a result, only TTS determines the transfer function of the complete filter,

$$H_F(\omega) = H_I^T(\omega) + H_{II}^T(\omega) + H_{III}^T(\omega) \, . \qquad (10)$$

An equivalent scheme model has been used for more accurate simulation of the frequency responses of the three-channel filter. This model considers a pair of coupled acoustic channels as a three-port network [7].

3. EXPERIMENTAL RESULTS

We have designed three CDMA IF filters based on the above-discussed structure. Fig.3 shows the frequency responses of the 220.38 MHz filter. The first (direct) and ninth transit signals are suppressed at DAS=35 dB and NTS=44 dB, respectively. Accordingly the amplitude ripple is low, AR=0.5 dB, within the pass band f_0 ±0.3 MHz. The insertion loss is IL=8.3 dB and the deviation from linear phase is PR=2.3^{o} rms within f_0 ±0.63 MHz. The 5 dB bandwidth is BW5=1.45 MHz and 33 dB bandwidth is BW33=2.27 MHz. The filter selectivity is SR=42-45 dB at f_0 ±1.25 MHz and ultimate rejection is. UR=60 dB at f_0 ±12.5 MHz. The filter uses quartz wafer, cut $yxl/33^o$ and package SMD 5.0x7.0x1.4 mm.

Fig.4 shows the frequency responses of the 210.38 MHz filter, also fabricated on quartz wafer $yxl/33^o$ but mounted in package SMD5.0x5.0x1.4 mm. The levels of DAS=40 dB and NTS=42 dB. Therefore AR=0.3 dB within f_0 ±0.3 MHz and PR=2.0^{o} rms within f_0 ±0.63 MHz. In addition, IL=8.9 dB; BW5=1.41 MHz; BW33=2.3 MHz; SR=42-45 dB at f_0 ±1.25 MHz and UR=55-60 dB at f_0 ±12.5 MHz. TES SPUDTs with electrodes $\lambda/8+\lambda/4+\lambda/6$ [7] have been used in 210.38 MHz and 220.38 MHz filters.

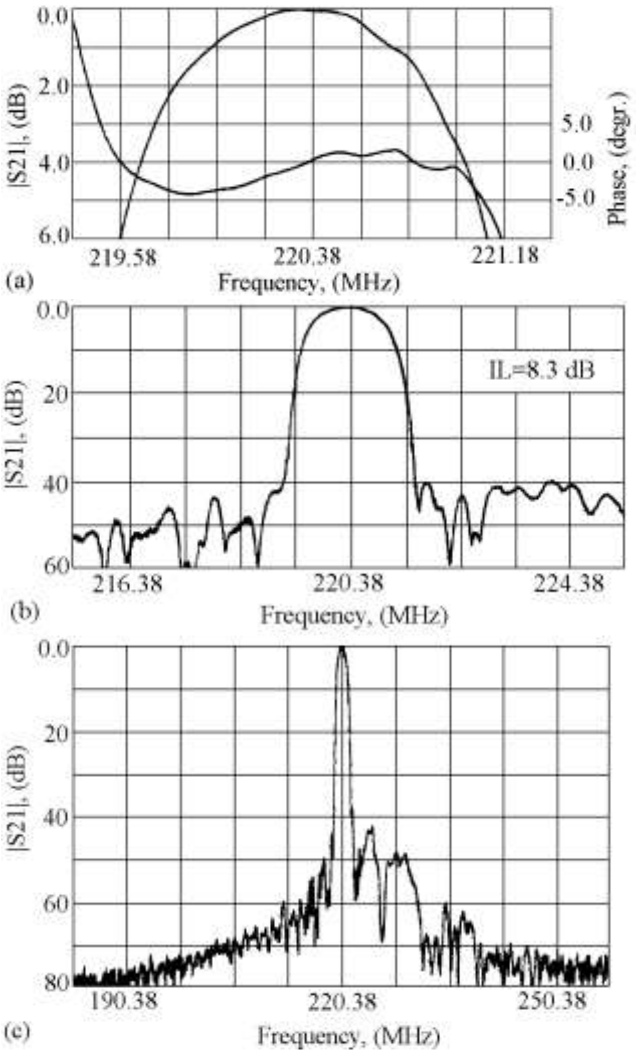

Fig.3. Measured responses S21 of 220.38 MHz filter: a - in pass band; b - in narrow band; c - in wide band

Fig.5 shows the frequency responses of the 85.38 MHz filter on LBO wafer, cut $YXwl/45^o$, in package SMD 9.0x5.0x1.6 mm. This filter uses DART SPUDTs with electrodes $\lambda/8+\lambda/8+3\lambda/8$ and has the parameters AR=0.8 dB within f_0 ±0.3 MHz; PR=5.0^{o} rms within f_0 ±0.63 MHz; IL=6.33 dB; BW5=1.18 MHz; BW33=1.77 MHz; SR=40-45 dB at f_0 ±1.25 MHz and UR=50-60 dB in f_0 ±12.5 MHz. The frequency responses of all filters are measured in the matched 50 Ohm track.

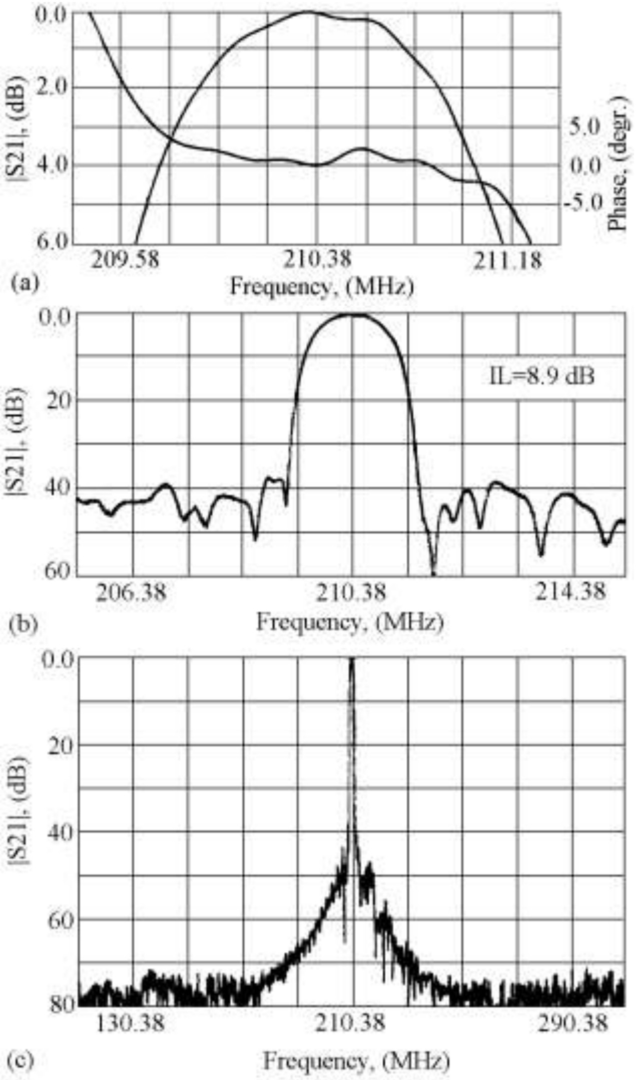

Fig.4. Measured responses S21 of 210.38 MHz filter: a - in pass band; b - in narrow band; c - in wide band

4. CONCLUSION

We have described a modified three-channel structure that allows the size reduction of SAW filters. The transverse acoustic coupling between the closely adjacent channels in this structure can be compensated to decrease the amplitude and phase ripples of the filter. Three CDMA filters, namely, 220.38 MHz; 210.38 MHz , and 85.38 MHz have been fabricated. Their parameters confirm the advantages of the structure proposed.

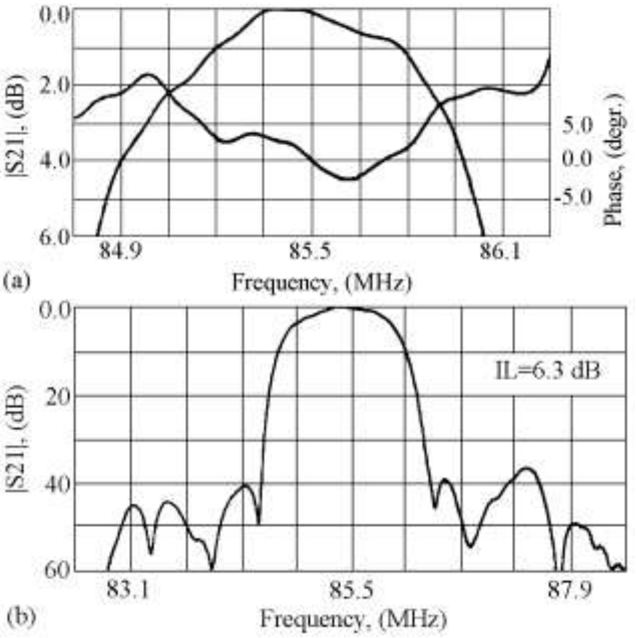

Fig.5. Measured responses S21 of 85.38 MHz filter: a - in pass band; b - in narrow band

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