SAW Vestigial Sideband Filter for TV Broadcasting Transmitter

TOSHIKAZU KODAMA, KOUJI SATO, AND YOSHIAKI UEMURA

Abstract—**This paper reports the application of an SAW vestigial sideband (VSB) filter to a TV broadcasting transmitter. The filter requires steep cutoff characteristics and large fractional bandwidth. The X-112°$Y\cdot LiTaO_3$ is one of the most suitable substrates for this filter, because it is possible to satisfy specifications without any compensation for typical temperature variations. The first application of $LiTaO_3$ substrates to a VSB filter is discussed in this paper. The filter is designed by a new optimization method using nonlinear programming. Because of the lack of a multistrip coupler and due to a large number of fingers, various second-order effects had to be compensated for. Experimental results are presented, when the filter is installed in a 10-kW TV transmitter.**

I. Introduction

THE VSB FILTER in a television transmitter is a very important component, which determines the broadcast video signal quality. Hitherto, it consisted of a conventional LC filter and a phase equalizer. Besides being bulky, it requires laborious adjustment by a skilled worker. Therefore, it is very expensive and has a long-term stability problem. The merit of replacing it with an SAW filter is self-evident.

Several filters with a low aspect ratio have been reported in the past few years [1], [2]. However, all the filters presented up to now are constructed on $LiNbO_3$. This paper describes the first application of the X-cut 112°-rotated Y-propagation $LiTaO_3$ [3] to a VSB filter for a TV transmitter. Since this substrate has a lower temperature coefficient (18 ppm/°C) than $LiNbO_3$, there is no need to compensate for typical temperature variations. However, the substrate has a lower piezoelectric coupling coefficient than $LiNbO_3$. This means that using a multistrip coupler [4] is not practical. Several methods have been developed for realizing an $LiTaO_3$ VSB filter without multistrip coupler. These include a new synthesis technique using nonlinear programming and compensation techniques for second-order effects, such as bulk wave, diffraction, electrode resistance, and charge distribution effects. Experimental results will be presented for a filter installed in an actual UHF 10-kW TV transmitter.

II. Specifications for a VSB Filter in a TV Transmitter

Typical response requirements for a VSB filter are shown in Fig. 1. The video carrier frequency is 38.9 MHz and the 1-dB bandwidth is 6 MHz. Other principal specifications

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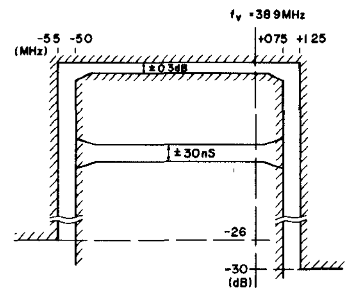

Fig. 1. Response requirements for a VSB filter.

are as follows:

source/load impedance	50 Ω
insertion loss	25 dB maximum
triple transit echo level	−46 dB maximum
VSWR	1.2 maximum
temperature range	0°C–50°C.

III. Filter Design

Since X-112°$Y\cdot LiTaO_3$ has lower coupling coefficient ($k^2 \simeq 0.008$) than $LiNbO_3$, it is not practical to use a multistrip coupler because it requires a large chip area. A filter configuration was adopted which consists of two interdigital transducers (IDT's). One is a uniformly spaced unapodized IDT. The other is a phase-modulated apodized IDT. In this case, an apodized IDT must be carefully designed, because the filter response depends mainly on the apodized IDT. Moreover, a smooth amplitude rolloff due to the uniform IDT must be compensated for in the weighted IDT design. A new technique has been developed which enables optimization using nonlinear programming [5].

Before optimization, the given bandpass filter function is translated down to baseband. This translation yields both in-phase and quadrature time components. Based on the sampling theorem, it can be shown that the impulse response is approximately interpolated by the sample values of these two components. Therefore, the optimization is carried out for these sample values, which are far fewer in number than the fingers in the transducer.

Sample values optimization takes place as follows. The method described in this paper allows the required response to be specified in terms of upper and lower value limits, together with the center values. The problem under consideration is to minimize the sum of the squares of

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errors from the center values, under the constraint that the response error will never exceed the allowed tolerances. This constrained optimization problem can be reduced to an unconstrained optimization problem [6] by sequential unconstrained minimization technique (SUMT). Assume that an actually realizable frequency response is given by

$$A(k)e^{-j\theta(k)}=R(k)+jX(k). \tag{1}$$

The SUMT [7] procedure requires the objective function to be composed of a penalty term as well as the sum of the squares of errors from the center values. This penalty term works as a constraint, and is successively modified as the optimization proceeds. Thus the optimization procedure expressed by SUMT can be written as follows:

mimimize the following function sequentially:

$$F(l)=F_T+F_P(l)$$

where

$$F_T=\sum_k a(k)\left\{\frac{A(k)}{A_s(k)}-1\right\}^2+\sum_k b(k)\left\{\frac{X(k)}{R(k)}-\frac{X_s(k)}{R_s(k)}\right\}^2$$

$$F_P(l)=\sum_k c(k)\left[\left\{\max\left(1,\frac{A(k)}{A_{SU}(k)}\right)\right\}^{2l}-1\right]+\sum_k d(k)\left[\left\{\min\left(-1,\frac{A(k)}{A_{SL}(k)}-2\right)\right\}^{2l}-1\right] \tag{2}$$

where $a(k), b(k),\cdots$ are weighting coefficients and l is a positive integer of a monotonically increasing sequence. Subscripts S, SU, and SL correspond to desired-response center values, upper limits, and lower limits, respectively. Phase response constraints are omitted from (2) in order to simplify the equation. $F_P(l)$ is the so-called penalty function. Thus the above problem can be solved by an ordinary simplex-like method or by the Davidon Fletcher Powell (DFP) method [6]. The DFP method was adopted because it converges faster than the simplex-like method. When the DFP method is used, it is necessary to compute not only the objective function values (2), but also the objective function gradient. Objective function values are easily obtained, because the frequency response given by (1) can be rapidly computed, using usual FFT programs. A new method has been developed for calculating the objective function gradient. The result shows that the gradient can be obtained by double application of FFT processes, when the amplitude and phase characteristics are given as a filter specification. In case a group delay characteristic is given, instead of a phase response, as a specification, the gradient is computed by applying FFT processes four times. In any case, it has been shown that only a short time is required for computing the gradient.

A VSB filter was designed by the above optimization method. As an initial value for optimization, an impulse response, obtained by simple Fourier analysis of the specification, was used.

IV. Second-Order Effects Reduction

The designed filter has an 8 finger uniform and a 280 finger length weighted transducer. Maximum aperture length for the apodized transducer is 20 mm, which was determined according to the impedance requirement and SAW diffraction analysis. Because of lack of a multistrip coupler and a low shape factor, various second-order effects [8] must be compensated for, as described below.

A. *Reflection Due to the Finger Edges*

Fig. 2 shows the electrode configuration used in the filter. The electrodes consist of one-eighth and five-eighth wavelength fingers. Reflections at the finger edges cancel each other, like the well-known double electrode [9]. Although the coupling efficiency is lower than the double electrode, this configuration is superior to the double electrode, with respect to production yield. Moreover, the bulk conversion [10], which occurs at the electrode edges, is smaller than that of the double electrode, because of the fewer electrode edges.

B. *Bulk-Wave Spurious*

The most troublesome bulk-wave spurious response is due to the shear wave, which appears in the upper frequency region of the passband. Recently, several authors have tried to find theoretically a low spurious response cut [11], [12]. However, quantitative evaluation is still difficult. It is more practical to measure spurious levels by an experiment after a short period of theoretical examination.

After the first qualitative examination of $LiTaO_3$ based on Mitchell's method, bulk spurious levels, in comparison with SAW response, were measured for several candidates, selected from temperature coefficient and coupling constant viewpoints [3]. There are two minimum temperature coefficient directions in the X-plane. One is about 112° from the Y-axis and the other is about 147° from the Y-axis. Their temperature coefficients are almost the same values (20 ppm/°C), however, their bulk-wave spurious responses are quite different. It has been found that the 112° direction has a lower spurious response than the 147° direction. This is why the X-cut 112°-rotated Y-propagation $LiTaO_3$ is used.

To reduce bulk-wave effects still further, the bottom of the substrate was roughened by a planetary grinding machine. Silicon carbide (SiC) of 60-μm average diameter was used as the abrasive powder.

C. *Charge Distribution Effects*

In case the filter fractional bandwidth is fairly large, as in the VSB filter, the individual electrode frequency responses due to the charge distribution cannot be ignored

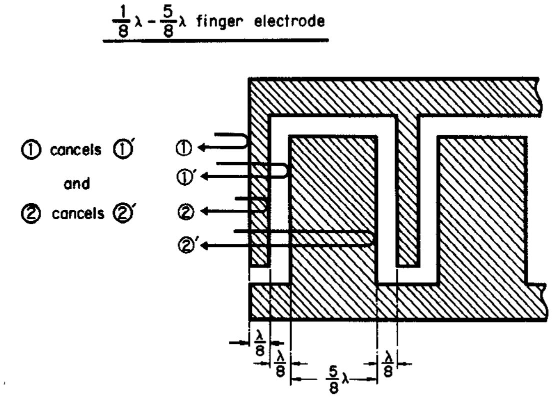


Fig. 2. Electrode configuration used in the filter.

[13]. The charge-based spectral weighting model [14] predicts that a filter amplitude response is nonsymmetrical about the center frequency, even if the filter consists of periodic transducers. This nonsymmetrical response may be corrected by the multistrip coupler [4], because the frequency response for the multistrip coupler can be controlled by changing its period. In the present case, the correction must be accomplished by other methods, because no multistrip coupler is used. A correction method was used, in which the compensation for the finger response was included in the design process. In the optimization process described in Section III, a nonsymmetrical amplitude specification is used. Its curve is gradually corrected by an iteration method, until the final analysis by the spectral weighting model shows a symmetrical response. This correction results in a nonperiodic apodized transducer. Other possible means of compensation include an individual electrode in a double electrode pair weighting [15] or adjustment in matching.

D. *Diffraction Effects*

In SAW filters, the diffraction deteriorates the characteristics, especially to fill the trap, and degrades stopband performance. Several papers have been published which deal with ways to correct these effects [16], [17]. However, the methods involved require a great deal of computing time and, consequently, design cost is high. It is shown that diffraction effect is well compensated for by selecting an appropriate uniform transducer aperture for the maximum aperture of the apodized IDT [18]. There is an optimum finger overlap length for the uniform transducer to minimize the difference between the experimental response and the design goal. In a VSB filter, the aperture ratio for an unapodized IDT to an apodized IDT is set equal to 1.2.

E. *Electrode Resistance Effects*

Electrode resistance effects cannot be ignored, when the aperture length is fairly large, as in this filter. By using a method similar to that developed by Lakin [19], the distributed resistance effects in the finger were examined analytically in case of the electrode configuration shown in Fig. 2. It was shown that an effective aperture weighting factor $\eta(w)$ is approximately given by

$$\eta(w) \simeq \left\{1 + j\frac{L_0^2}{2Q_e}(w-1)^2\right\} w \tag{3}$$

where

- L_0 maximum aperture normalized to the wavelength λ_0;
- w aperture normalized to L_0;
- Q_e $= 1/(\omega C\rho)$;
- C single electrode capacitance $[\mathrm{F}/\lambda_0]$;
- ρ $= \frac{1}{2}\left(8\rho_s + \frac{8}{5}\rho_s\right) = \frac{24}{5}\rho_s$ = effective electrode resistance $[\Omega/\lambda_0]$;
- ρ_s sheet resistance $[\Omega/\square]$.

Equation (3) holds approximately when the radiation $Q(=\omega CR_a$; R_a = radiation resistance per wavelength) for a single finger pair is not so small. It also holds true, when the sheet resistance is not too large. Equation (3) shows that, when the electrode resistance is finite, weighting efficiency is no longer proportional to the aperture length. Also a phase error is introduced by apodization. The electrode resistance effect causes deterioration in the filter response, such as an increased sidelobe level and passband distortion. Modified finger overlap length and position were calculated, using (3).

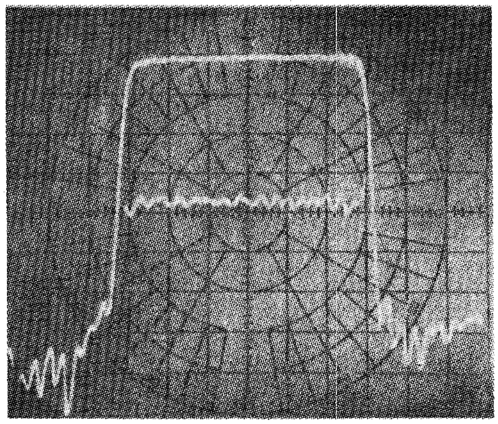
V. Experiments

A. *VSB Filter Response*

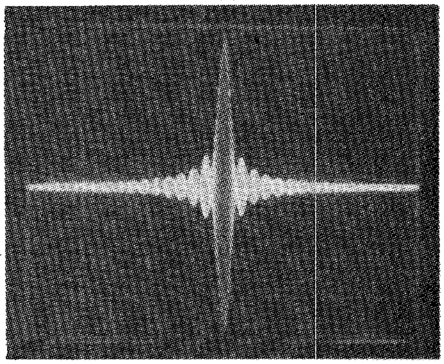
Fig. 3(a) shows representative amplitude and group delay characteristics of an SAW VSB filter, whose center frequency is 36.9 MHz. Fig. 3(b) shows its impulse response. A 3 dB–30 dB shape factor is 1.12. Other measured characteristics are as follows:

passband amplitude ripple	±0.12 dB
passband group delay deviation	±25 ns
insertion loss (tuned)	24 dB
TTE level	−55 dB.

The above measured responses completely satisfy the given specifications described in Section II, for the temperature range from 0°C to 50°C.



(a)



(b)

Fig. 3. SAW VSB filter responses. (a) Amplitude and group delay responses. Vertical scale: 5 dB/div (amplitude), 100 ns/div (group delay). Horizontal scale: 1 MHz/div (center frequency: 36.9 MHz). (b) Impulse response. Horizontal scale: 0.5 μs/div.

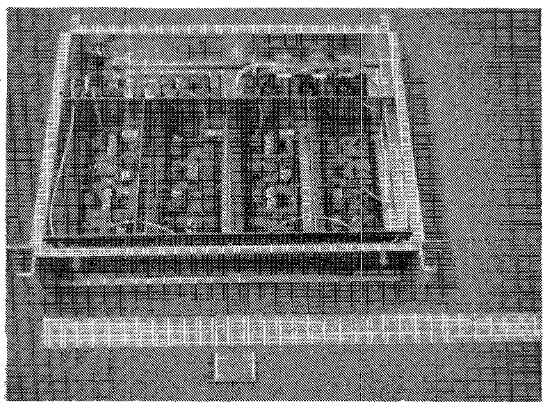


Fig. 4. Comparison between the size of a conventional LC filter and an SAW filter

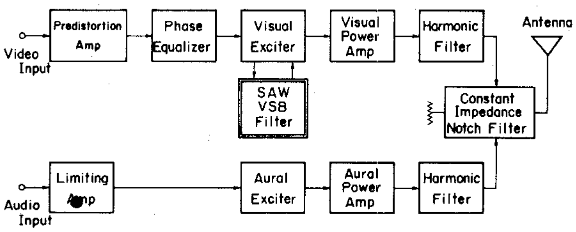


Fig. 5. Block diagram of a television transmitter.

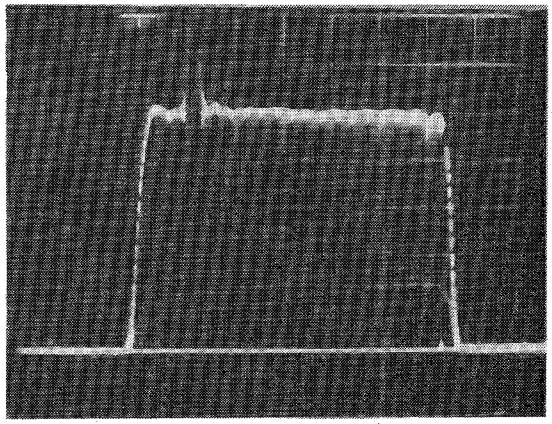


Fig. 6. Overall transmitter amplitude–frequency response. Vertical scale: linear. Horizontal scale: 1 MHz/div. (Video carrier frequency: 189.25 MHz.)

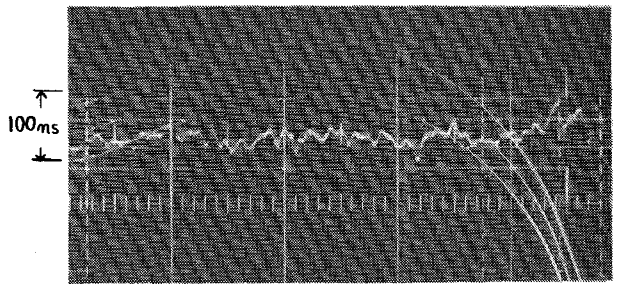


Fig. 7. Overall transmitter group delay response.

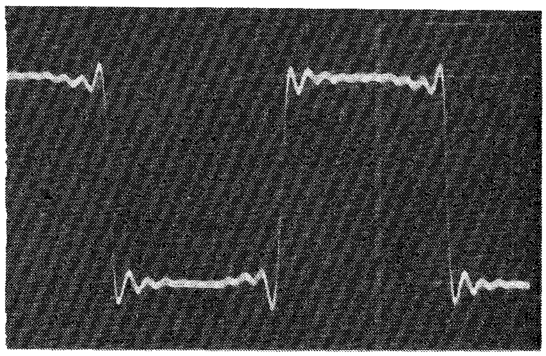


Fig. 8. 330-kHz square wave response. Horizontal scale: 0.5 μs/div.

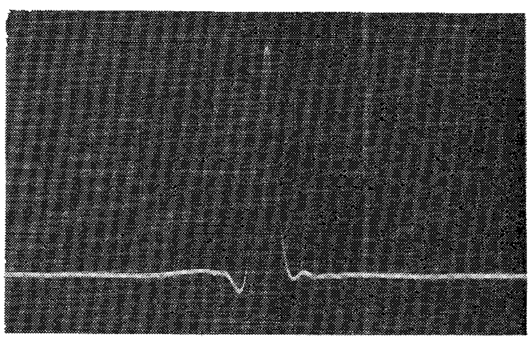


Fig. 9. $2T$ pulse response. Horizontal scale: 0.5 μs/div.

Fig. 4 shows a size comparison between an SAW VSB filter and a conventional *LC* filter. The completed filter is housed in a hermetically sealed package.

B. Overall Transmitter Responses

The filter was actually installed in a UHF 10-kW TV transmitter. Fig. 5 shows a simplified block diagram of the transmitter. The VSB filtering section includes drive and output amplifiers, in addition to an SAW VSB filter.

Several important overall transmitter responses were measured at the output of constant impedance notch filter. A group-delay predistortion circuit (receiver equalizer) was not used in any of the measurements. Fig. 6 shows the overall transmitter amplitude response on a linear vertical scale. Fig. 7 shows the overall group delay response. It can be seen that the group delay deviation is within ± 30 ns. Fig. 8 shows a 330-kHz square wave response. Overshoot is about 3 percent and preshoot is about 10 percent. Fig. 9 shows a $2T$ pulse response, where $2T$ equals 0.2 μs. It is found that the K-rating factor is less than 1.8 percent. The square wave and $2T$ pulse responses were measured using a linear envelope detection circuit.

From the above results, it was found that the filter performance satisfies transmitter specifications.

VI. Conclusion

The first application of X-112°Y·$LiTaO_3$ SAW VSB filter to a TV broadcasting transmitter has been discussed in this paper. The most significant advantages of using X-112°Y·$LiTaO_3$ are that the no compensation is needed for temperature variations, due to the low temperature coefficient, and that the substrate has inherently low spurious bulk modes. However, the square of the substrate piezoelectric coupling coefficient is only one-fifth of that of $LiNbO_3$ substrate. This fact makes it impractical to use a multistrip coupler. In order to realize the $LiTaO_3$ VSB filter without multistrip coupler, several techniques have been developed. First, an optimization method for designing an SAW filter by nonlinear programming was discussed. A new method for calculating the objective function gradients has been developed. It was proved that gradients can be obtained by FFT algorithm, which leads to fast convergence because that makes it possible to use the DFP method. This optimization method is particularly effective for designing an SAW filter having a low shape factor, like the VSB filter. Next, several techniques were introduced for reducing the spurious responses due to the second-order peculiar to the $LiTaO_3$ SAW VSB filter.

The completed filter was installed in a UHF 10-kW TV transmitter and tested. It was shown that overall transmitter responses meet all the specifications required for a TV transmitter.

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