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Short Papers

Surface Elastic Wave Bandpass Filters for Frequency Synthesis

F. S. HICKERNELL, A. J. KLINE, D. E. ALLEN, AND W. C. BROWN

***Abstract*—Surface elastic wave bandpass filter techniques have been applied to the development of a minaturized frequency synthesizer for satellite communications systems. A bandpass filter centered at 247 MHz has been developed exhibiting less than 7-dB insertion loss over a 6-MHz 1-dB band, with sidelobe rejection greater than 45 dB.**

The surface elastic wave bandpass filter is becoming an attractive component for application in communication systems at VHF and UHF frequencies [1]–[8]. The surface wave filter is a fabricationally simple compact planar structure easily integrated into hybrid microelectronic circuitry. It is an extremely versatile component permitting a wide range of desired band shape responses to be developed, because phase and amplitude characteristics can be synthesized independently. It represents a reliable reproducible component that can be produced inexpensively.

Surface-wave bandpass filters were designed, fabricated, and evaluated for use in frequency synthesis for minaturized communications systems.[1] The synthesizer uses frequency multiplication and addition techniques to develop stable coherent reference signals for an ultrasensitive receiver.

The primary requirements for filter performance were a moderate bandpass, low insertion loss, and strong rejection of unwanted sidebands. Equally important requirements were small size and process compatibility and integrability with other hybrid microelectronic circuitry. The surface-wave filter represented an attractive alternative to cascaded high-Q LC networks which would have required shielding and taken up considerable space.

Surface-wave filters were initially developed for evaluation at frequencies of 56, 150, 220, and 244 MHz. The filters were designed with lithium niobate ($LiNbO_3$) as the substrate material. The transducer structure of each filter consisted of a central interdigital electrode having a $\cos^2 x$ apodization with two adjacent electrodes of uniform finger overlap. The interdigital electrodes had equal quarter wavelength linewidths and spacings. The center apodized pattern controlled the main in-band frequency response with the two adjacent patterns having fewer finger pairs and thus a broader band response. The outer patterns also included a separate reflector region spaced by an odd number of quarter wavelengths to test the use of a unidirectional response near band center [9]. The interdigital patterns were photoetched on polished Y-cut plates of $LiNbO_3$ with surface-wave propagation along the Z axis. The electrode metal was evaporated aluminum approximately 2500 Å in thickness.

The four filters were evaluated with the signal input to the center apodized pattern and the output from the adjacent patterns, which were electrically connected together. Series inductors were used to match impedance levels. The insertion loss of the filters was in the range of 5–7 dB. A 3-dB bandwidth of 7 percent was obtained, and the sidelobe rejection was greater than 30 dB. The ripple in the bandpass was less than ± 1 dB.

The filters at 200 and 244 MHz were evaluated using the "quarter wavelength reflectors" to give a unidirectional characteristic to the output transducers. The reflectors were optimized with a series inductance to ground [9]. The beneficial effect of the reflectors was to reduce the loss by an additional 2 dB (a total filter loss as low as 3 dB was obtained) and to reduce the ripple in the bandpass to ± 0.25 dB. There was, however, a decrease in the bandwidth and an additional ripple in the skirt region. The lower sidelobe level outside the bandpass region was unaffected.

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The authors are with the Motorola Government Electronics Division, Scottsdale, Ariz. 85252.
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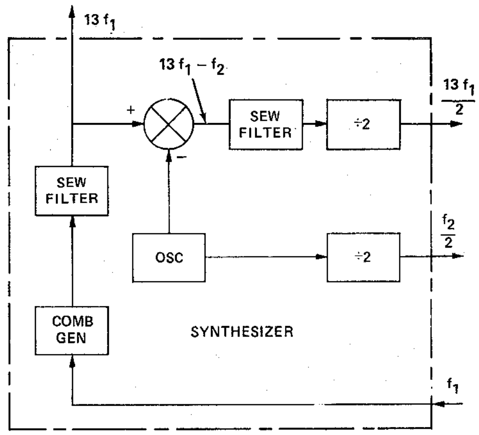

Fig. 1. Synthesizer configuration.

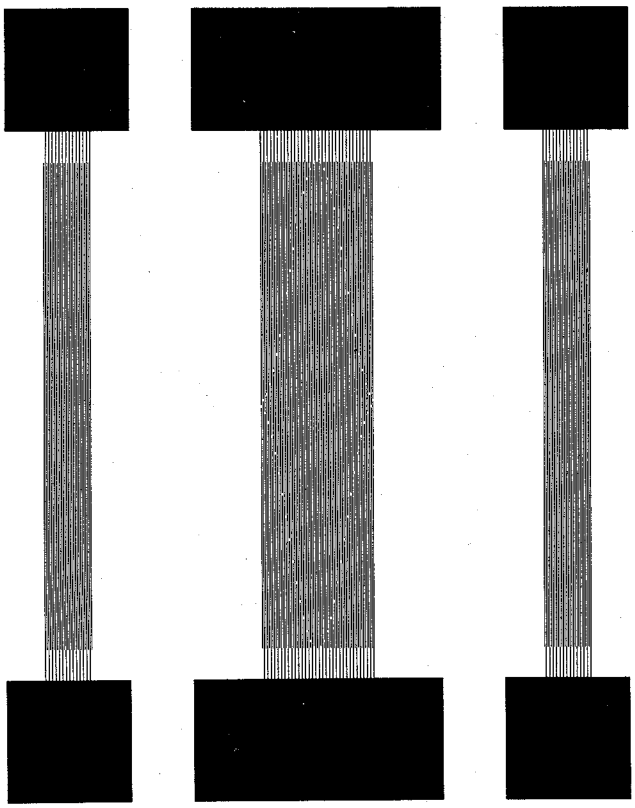

Fig. 2. Mask pattern for 247-MHz bandpass filter.

Based upon the performance of the filters and system requirements, the synthesizer configuration shown in Fig. 1 was selected. The two outputs of the synthesizer from the surface-wave filters are shown in the block diagram. The synthesizer is intended to be used for a range of frequencies requiring that a corresponding band of input frequencies be accommodated.

The filter designed for 220 MHz (used in the $13f_1$–f_2 channel) had characteristics sufficiently good to require no further design changes. The 244-MHz filter ($13f_1$) was redesigned for 247 MHz with a narrower bandwidth and greater sidelobe rejection. The electrode design is shown in Fig. 2. The central pattern had 33 finger pairs with a $\cos^2 x$ apodization. The output electrodes were 14 finger pairs each with uniform finger overlap. One of the outer patterns was offset by 180° with respect to the other. The finger width was 0.140 mil and the finger spacing was 0.135 mil. The finger overlap of the outer patterns was 78.75 mil.

The filter at 247 MHz selects the thirteenth harmonic of a signal fed to a comb generator. Fig. 3 shows the spectrum out of the filter

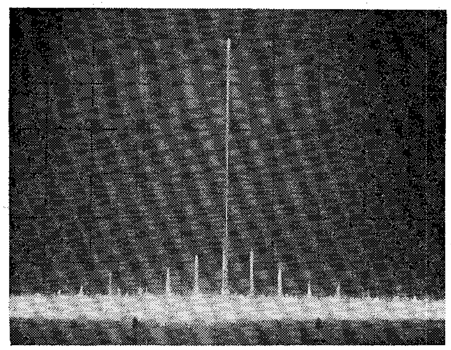

Fig. 3. Spectrum response of 247-MHz filter driven by comb generator. Vertical scale: 10 dB/div., horizontal scale: 30 MHz/div.

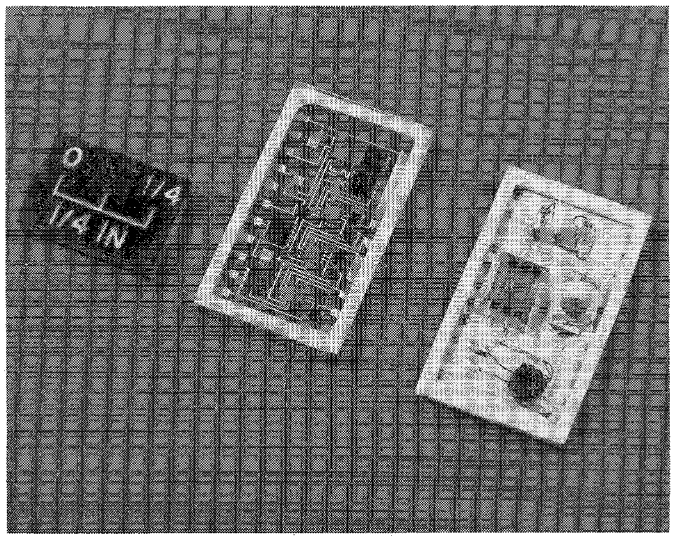

Fig. 4. Filter and amplifier modules used in miniaturized synthesizer.

when used in this manner. The adjacent harmonics (particularly the twelfth and fourteenth) are attenuated greater than 45 dB with respect to the desired thirteenth harmonic. This sideband rejection was maintained over a 2.5-percent change of input frequency and the in-band ripple was held to less than ± 0.5 dB over this frequency range. This frequency tolerance easily permitted operation over a 100°C temperature range despite the temperature coefficient of the filter substrate material. The insertion loss of the filter was less than 7 dB. In operation, the output signal from the filter is taken differentially providing a large rejection to direct capacitive coupling from input to output because of 180° offset of the transducer structures.

The minaturized synthesizer is composed of a number of hermetically sealed modules that perform the basic functions of frequency multiplication, heterodyning, amplification, and filtering. Fig. 4 shows a photograph of the 220-MHz filter unit packaged in module form with coil and transformer matching networks. Also shown is another module containing beam lead electronic components. Six of these $\frac{3}{8}$- by $\frac{5}{8}$-in modules, three of which are filter units, are used for the synthesis of the desired signals.

There are several advantages that surface elastic wave filter technology has brought to frequency synthesizers of the type described in this short paper. Surface-wave filters have easily provided the required miniaturization and hybrid integrated circuit compatibility. They have provided reproducibility, low fabricational cost, convenience of impedance levels, hard mounting for ruggedness, and the alleviation of electrical shielding problems.

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Bandpass Filters Using Nonlinear FM Surface-Wave Transducers

J. C. WORLEY

Abstract—**A generalized FM method of obtaining bandpass filters using surface-wave delay lines is discussed. Two identical FM transducers are used as input and output. The FM function of the transducers is determined by the required passband.**

It is well known that surface-wave delay lines (SWDL) can be used in a wide variety of filter types [1]. One of the most promising areas of application is bandpass filters [2]. Through a relatively simple technology, it is possible to synthesize quite complicated and exact bandpass functions. This short paper discusses one of several possible methods of obtaining this type of filter with a SWDL, a so-called nonlinear FM technique.

The most common and straightforward method of obtaining bandpass filters from a SWDL is a time domain replica approach. In this method one of the transducers is designed so that it has an apodization similar to the Fourier transform of the desired frequency response and all fingers resonant at the center of the band. The second transducer is a standard interdigital transducer with a bandwidth large enough so that it does not affect the overall response of the delay line. For example, if one desired a rectangular passband, the replica transducer would have a $\sin x/x$ apodization.

This approach has the distinct disadvantage that there are fingers in the transducer which have small overlap, some approaching zero, which in turn leads to large diffraction losses. This can possibly be neglected if one uses a focusing material such as $LiNbO_3$. However, this restricts the type of material which can be used, eliminating *ST* quartz, $Bi_{12}GeO_{20}$, and ceramics.

A solution to this difficulty can be found in signal processing theory. It is possible to code the two transducers with an FM signal and then by arranging them as time translates of each other, i.e., so they will autocorrelate. A frequency response results which is not an FM but just the spectrum amplitude of the FM with a linear phase. In other words, with this method the bandpass response of the delay line can be determined by the manner in which the center-to-center spacings of the fingers are graded, independent of the overlap of the fingers. A delay line of this type is shown in Fig. 1. The design shown in Fig. 1 does use a small amount of finger apodization near the transducer ends. However, this apodization is used only to smooth out the effects of passband ripple.

To obtain the desired frequency response, one must specify the corresponding FM function. This requires the Fourier transformation of a frequency domain envelope function into a time domain FM function with an arbitrary envelope. This can be done through the use of stationary phase [3]. A detailed theoretical derivation of this principle is given in [2], and only the pertinent equations will be shown here. The principle of stationary phase leads to the relation

$$\phi'(t) = 2\pi f(t) \tag{1}$$

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The author was with Sperry Rand Research Center, Sudbury, Mass. 01776. He is now with Microsonics, Inc., Weymouth, Mass. 02188.

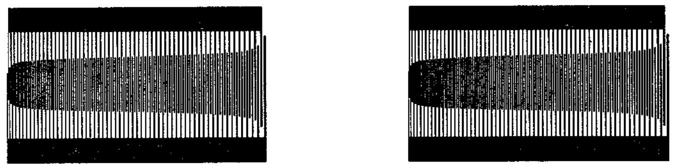

Fig. 1. Transducer pair for FM bandpass filter.

where $\phi(t)$ is the phase function describing the FM and $f(t)$ is a function derived as shown in the following paragraphs. A second equation, derived from Parseval's theorem

$$\int_{-\infty}^{f} |U(\xi)|^2 \, d\xi = \int_{-\infty}^{t} |u(\eta)|^2 \, d\eta \tag{2}$$

completes the equations necessary to find the desired phase function. Note that $U(\xi)$, which is the square root of the bandpass function, and $u(\eta)$, which is the square root of the time response of one transducer, are specified by the designer, and thus the time envelope of a single transducer is independent of the frequency response of the delay line. $\phi(t)$ is found by integrating (2) to give two new functions

$$P(f) = Q(t)$$

solving for f

$$f(t) = P^{-1}(Q(t))$$

substituting this into (1)

$$\phi'(t) = 2\pi P^{-1}(Q(t))$$

and integrating,

$$\phi(t) = 2\pi \int P^{-1}(Q(t)) \, dt + C.$$

As an example, consider the case of a rectangular passband and a constant time envelope:

$$|U(f)|^2 = 1, \qquad -\frac{\Delta f}{2} \le f \le \frac{\Delta f}{2}$$

$$|u(t)|^2 = \frac{\Delta f}{T}, \qquad -\frac{T}{2} \le t \le \frac{T}{2}.$$

Integrating (2) with these amplitude functions gives

$$\frac{\Delta f}{T} \int_{-T/2}^{t} d\eta = \int_{-\Delta f/2}^{f} d\xi$$

$$\frac{\Delta f}{T}\left(t + \frac{T}{2}\right) = f + \frac{\Delta f}{2}.$$

Inverting the preceding equations and performing the integration indicated in (1) leads to

$$\phi(t) = \pi \frac{\Delta f}{T} t^2 + C$$

which is the phase function for a linear FM signal at baseband. This may be placed on a carrier by adding $\omega_0 t$ to the right side of the preceding equation.

An example of spectrum amplitude and time amplitude functions which lead to a nonlinear FM function is

$$|U(f)|^2 = (1/\pi W)/\sqrt{1 + (f/W)^2}$$

$$|u(t)|^2 = 1/\sqrt{T}, \qquad 0 < t < T$$

where W is the 3-dB bandwidth [2]. The corresponding phase function is

$$\phi(t) = 2\pi W \int \tan\left[\frac{\pi}{2}\left(1 - \frac{t}{T/2}\right)\right] dt + C.$$

Fig. 2 shows the theoretical and experimental response of a delay line with a pair of transducers defined by the preceding phase function. The dashed curve is the desired response, and the solid curve is the theoretical response of the delay line as predicted by an equivalent circuit model. The circles are experimental points. Unfortunately, the leakage of the delay line was sufficiently high so