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# Precision SAW Filters for a Large Phased-Array Radar System

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***Abstract*—The electronically steerable radar (ELRA) at the Forschungsinstitut für Funk und Mathematik is an experimental *S*-band phased-array radar system consisting of separate transmitting and receiving arrays employing several coherent and incoherent signal-processing and data-handling techniques, incorporating multiple beam and multifunction operation for target search and tracking, adaptive interference suppression, and target resolution.**

**This paper deals with the development and application of two types of SAW filters for the IF amplifier channel of the receiving array. Compared to conventional filters with lumped elements, these filters have some important merits. By making use of a special tuning technique, the center frequencies of all filters were adjusted, resulting in an rms deviation of less than 1 kHz. One type of the SAW filters represents an almost ideal approach of realizing a matched filter for rectangular shaped pulses. The conformity of the frequency responses of several hundred filters improved the noise suppression capability of the system. The use of the filters described represents one of the applications where high-quality mass-produced SAW devices have been applied to improve system reliability and performance.**

## I. INTRODUCTION

THE PRIMARY AIM of the ELRA system is to prove the practical feasibility of theoretical studies of signal processing, data management, and system control in a multitarget environment, in connection with phased arrays. To achieve highest flexibility for these widespread demands, separate antennas with active modules for transmitting and receiving have been chosen. The following description of the system should only give an overview. More detailed information on the architecture and specifications can be found elsewhere [1]–[3].

Three-hundred printed dipoles, randomly distributed on planar circular apertures of 28 wavelengths diameter, and 768 dipoles on 39 wavelengths diameter, respectively, form the transmit and receive arrays.

Each transmitter dipole is fed by its own module consisting of a 3-bit diode phase shifter followed by a triode amplifier. One of the 768 receiver channels is illustrated in Fig. 1. A bipolar transistor preamplifier in connection with an image rejection mixer assures a low noise figure of about 3 dB. The IF amplifier uses either a narrow-band or a broad-band SAW filter adapted to the transmitter pulse length of 10 $\mu$s for search and 2 $\mu$s for tracking, respectively. The filtered IF signals are converted down to baseband by a synchronous detector with two orthogonal components for coherent signal processing. The IF reference is fed to the mixer through a 3-bit phase shifter which consists of an 8-channel analog multiplexer connected to a tapped delay line. The outputs of 16 neighboring elements are combined in summing operational amplifiers and converted from analog to digital representation. Parallel trees of binary adders combine these subarray signals forming

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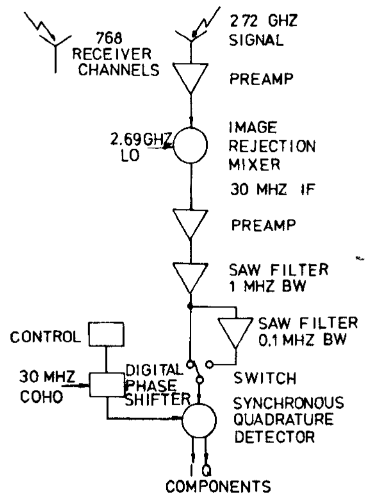


Fig. 1. Block diagram of ELRA dual bandwidth (1, 0.1 MHz) receiver channel.

sum and difference beams. The outputs are processed by an incoherent sequential detector or by a coherent Doppler filter. The whole system is controlled by a hierarchy of different computers.

## II. Choice of Filter Parameters

There are some features of the ELRA system which influence the choice of SAW filters for the IF amplifier.

1) The large number of active modules requires a high degree of stability. Alterations of single-element parameters will gradually worsen the system performance which can only be recovered by permanent supervision and maintenance. To reduce this effort, an automatic procedure for measuring and compensating relative phase drifts between the active modules has been incorporated. But there remain other parameters which cannot be controlled constantly, such as frequency responses, and, therefore, should not show temperature drifts or ageing effects.

2) For the adaptive suppression of directional or correlated noise, a good conformity of all channels is imperative. Nonoverlapping parts of the filter bandpass characteristics will allow to pass noise power which is not correlated from channel to channel and hence cannot be suppressed. For noise suppression, the individual frequency responses do not, necessarily, have to be the matched filter responses but they all should be identical in amplitude and phase.

3) The system uses the following method for generating up to six quasi-simultaneous independently steerable beams. The transmit array serially radiates up to six nearly rectangular, unmodulated pulses, each of them into a different direction. The pulse length depends on the actual task of the radar. The receiving and processing of echoes out of these six directions is possible by cyclic switching of the phase shifters according to these directions and by appropriate switching of the bandwidth according to the pulse length transmitted. To avoid loss of signal-to-noise ratio, the bandwidth of stages ahead of the phase shifter must be matched to the signal bandwidth, whereas following stages require a larger bandwidth corresponding to the switching rate. Therefore, this procedure can neither be applied with RF phase shifters, nor is it possible to filter after beamforming at baseband.

4) Most of the system energy is used in the search mode. Therefore, the receiver sensitivity should be as high as possible. The most efficient filter which maximizes the signal-to-noise ratio is a matched filter whose frequency response is the complex conjugate of the transmitted spectrum. Thus the matched filter for the rectangular 10-$\mu$s pulse should have a sin $X/X$-type frequency characteristic with a 3-dB bandwidth of about 88 kHz, the first nulls being 100 kHz from the center frequency of 30 MHz.

Considering all these points we derived the architecture of the IF amplifier as shown in Fig. 1. When receiving short pulses, only the broad-band filter is effective since the narrow-band filter is bypassed. For the longer pulses both filters are connected in series.

The realization of the narrow-band, matched filter by conventional *LC* devices was very difficult owing to the small relative bandwidth of 0.3 percent. Several weakly coupled bandpass resonators with high-$Q$ ferrite materials had to be used to achieve a fairly good approximation of the main lobe frequency response. But this design suffered from a poor long-term stability, a difficult tuning procedure, and a high temperature sensitivity. Thus we decided to use an SAW filter instead, expecting a higher temperature stability and a better approximation of the desired filter function.

An *LC* broad-band filter, easier to design than the narrow-band filter, could not sufficiently fulfill the demand of conformity. For this purpose we again chose an SAW filter. Since this filter is applied in the radar tracking mode where the estimation of target parameters is more important than the detection capability, we chose a rectangular shaped filter with a 1-MHz bandwidth, sacrificing 1.7 dB of efficiency relative to a matched filter, but obtaining a higher near band selectivity. For our filter, the loss compared to an "optimum" rectangular filter for 2-$\mu$s pulses with a 685-kHz bandwidth is 0.85 dB higher, but the time behavior is better. Nearly the same loss occurs with 1-$\mu$s pulses which will be used in future experiments with phase-coded signals. Thus the 1-MHz bandwidth chosen represents a compromise design for both 1- and 2-$\mu$s pulses. The loss can partly be reduced by digital filtering during signal processing.

## III. The SAW Bandpass Filters

Today's principal materials for SAW devices are lithium niobate and quartz, which have very different acoustic and thermal properties. Lithium niobate, which is about 30 times stronger piezoelectrically but more temperature sensitive than quartz, is generally used for wide-band low-loss applications. Because of the much lower piezoelectric activity of quartz, a low insertion loss is only possible with narrow-band filters. Since temperature stability, which in turn means stability of frequency and phase, was of prime importance in our phased-array radar application, our choice was *ST* quartz, which is 42.5° rotated *Y*-cut, *X*-

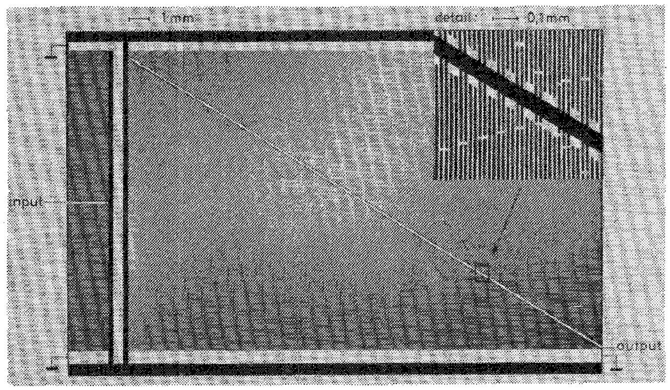


Fig. 2. Wide-bandwidth (1-MHz) SAW filter with sin $X/X$-type weighting function land rectangular frequency response.

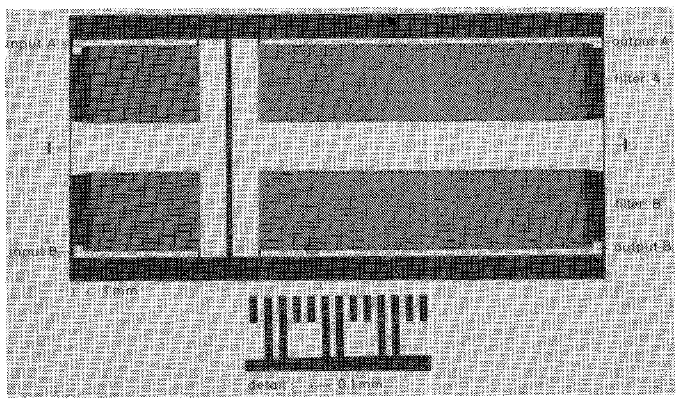


Fig. 3. Two narrow-bandwidth (0.1-MHz) SAW filter structures (unweighted) with sin $X/X$-type frequency response on one $ST$-quartz substrate.

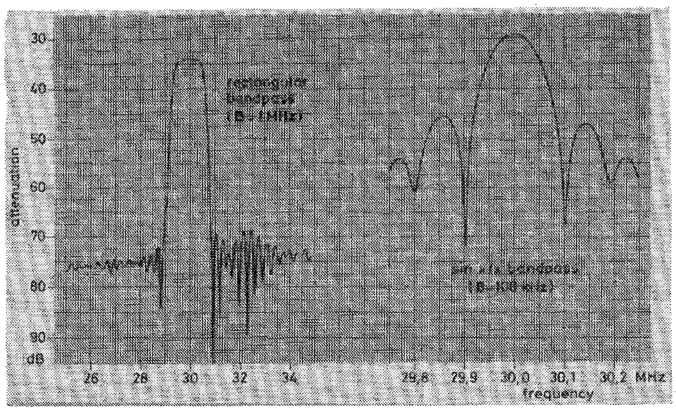


Fig. 4. SAW filter frequency responses (in a 50-Ω system).

propagating. The filters consist of an input and output transducer (reciprocal, except for impedance level) on a polished surface of the $ST$-quartz substrate. Transducers are fabricated by sputtering aluminum and defining the electrode pattern by standard photolithography. For the filters described in this paper, we use one wide-band transducer with few electrodes, and one narrow-band transducer which almost entirely determines the frequency characteristic of the filter. Depending on their design, transducers are capable of operating at odd harmonic frequencies [4]. For instance, a transducer operating at the third harmonic requires fewer and larger electrodes [5]. This is very often advantageous for the lithographic work because the yield is vastly improved. On the other hand, fewer electrodes may cause the transducer impedance level to be too high, causing matching problems and thus high insertion loss.

### *A. The Wide-Band Filter*

In order to obtain a rectangular type passband, it is necessary to use at least one weighted transducer. We chose a sin $X/X$ weighting function with a total of six sidelobes multiplied with a truncated Kaiser function [6]. An in-house-developed computer program [7] was used to select the optimum number of sidelobes for the sin $X/X$ function and the width of the Kaiser function. For example, increasing the number of sidelobes beyond three only lengthens the filter structure without gaining anything in performance.

Diffraction effects due to an increasingly smaller acoustic aperture of the sidelobes can lead to a degradation in performance. Decreasing the number of sidelobes seems attractive because it leads to a saving in substrate material, but unfortunately also to higher ripples in the passband. Narrowing the Kaiser function causes increasing sidelobe suppression and decreasing rolloff at the band edges.

The final design of the wide-band filter is illustrated in Fig. 2. The substrate sized used was $2\times0.75\times0.1$ in ($50\times19\times2.5$ mm). Because of the large bandwidth and the small piezoelectric coupling of $ST$ quartz, it was necessary to use the maximum possible acoustic aperture, which was 150 acoustic wavelengths, to obtain minimum insertion loss. In order to reduce the electrical resistance of the electrodes, the transducers were split into two parts and fed from the center. Double electrodes were used with the fundamental frequency at 30 MHz and a third harmonic at 90 MHz. The frequency response is illustrated in Fig. 4.

### *B. The Narrow-Band Filter*

In order to increase the yield in manufacturing the filters and still maintain a tolerable insertion loss, we chose to design this filter to operate at the third-harmonic frequency. This allowed us to use fewer and larger electrodes as are required for fundamental frequency operation. The number of electrodes is still sufficient to obtain a tolerable insertion loss. The design of this filter is illustrated in Fig. 3. The substrate size was identical to that of the sideband filters. Two filters are placed on one substrate and are separated about 4 mm to keep the acoustic coupling between them low (less than $-40$ dB). The frequency response of this filter is illustrated in Fig. 4.

### *C. Fabrication of the Filters*

Because of the low operating frequency of 30 MHz, the minimum transducer electrode dimensions (finger width) are of the order of tens of micrometers. The total electrode area is however several square centimeters. Our filters were thus made in a single 20:1 reduction step. If a number of filters are made using the same mask, there exists a spread

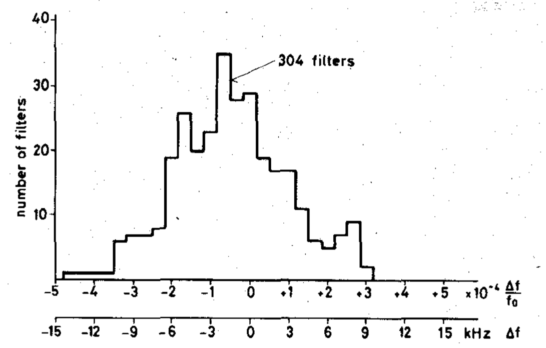


Fig. 5. The distribution of the center frequency of the narrow-band filters (0.1 MHz) after etching of the aluminum pattern.

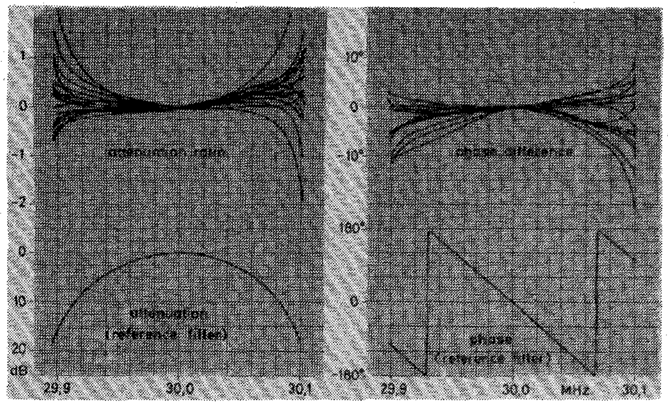

Fig. 6. Attenuation ratio and phase difference of 15 narrow-band filters compared to one reference filter.

in the center frequency as illustrated in Fig. 5. Several masks, differing slightly in the reduction ratio, were made. The mask which yielded filters with center frequencies closest to 30 MHz was then used to fabricate all filters. The substrates were cleaned and vacuum coated with 0.4 $\mu$m of aluminum using an RF sputtering system. Shipley AZ 1350H positive photoresist was used to define the filter pattern. The aluminum was then chemically etched in warm (31°C) 16 $H_3PO_4$:2 $H_2O$:1 $HNO_3$:1 $CH_3COOH$ acid. The etch rate was about 0.075 $\mu$m/min. The yield for the wide-band filters (Fig. 2) with 12-$\mu$m electrodes and 4.5-cm$^2$ electrode area has been 50–80 percent.

We have measured the frequency of several hundred narrow-band filters and plotted the deviation from nominally 30 MHz in Fig. 5. A Gaussian-type spread is obtained with an rms mean deviation of several kilohertz. The data shown illustrate the center frequencies after the aluminum transducers have been etched.

The center frequency of SAW devices may be varied a small amount by depositing thin dielectric insulating films onto the surface of the filters. For our specific case, an *increase* in the center frequency is obtained by depositing a thin layer of insulating material with a *higher* acoustic velocity than *ST* quartz. Such a material is aluminum oxide $Al_2O_3$. Similarly, if the frequency must *decrease*, one must deposit a material with a lower acoustic velocity than *ST* quartz such as zinc oxide ZnO. Insulating ZnO is obtained by sputtering in an argon atmosphere with a few percent oxygen. $Al_2O_3$ may also be sputtered in argon or evaporated with an electron gun. At 30 MHz, the required film thickness for a frequency change of 10 kHz is about 0.1 $\mu$m or less, a very practical value. The filters described here were all adjusted to a frequency of $30 \pm 0.001$ MHz, by monitoring the frequency during the sputter or evaporation process. Since this work was started, others [8] have also used dielectric films in order to achieve small frequency or phase changes.

### D. Filter Performance

Fig. 4 shows the frequency response of the filters around the center frequency, demonstrating the fundamental difference between both types of filters. The maximum sidelobes of the rectangular wide-band filter remain below 35 dB. Ripples in the passband do not exceed 0.5 dB. The main lobe of the narrow-band filter agrees well with the theoretical sin $X/X$ function, whereas the sidelobes differ in height and symmetry. But this effect seems irrelevant to the system performance because the main lobe of the frequency spectrum comprises most of the signal power, and the real transmitted pulse differs from the ideal rectangular pulse by its finite rise and decay time causing a spectrum with a reduced sidelobe level.

In order to get a critical measure of the repeatability of the frequency response, we arranged one of the narrow-band filters in the reference path and another one in the test path of an HP 8407A network analyzer which thus directly delivered the attenuation ratio and the phase difference of both filters. In Fig. 6 we compared 15 different filters with a reference filter. The frequency range of the measurement covered only a section of the principal lobe, because the reference level had to be kept sufficiently high. One can infer the cause of the inaccuracies from the form of the difference functions in the upper part of the diagram. Deviations of the center frequency produce odd functions, and deviations of the bandwidth produce even functions. Other functions point to a mixture of both causes. These deviations have an rms value smaller than 1 kHz. This excellent value cannot be obtained with conventional *LC* filters. Between individual filters, differences of attenuation and phase at the center frequency (tuned out with this measurement) are naturally larger, but of no great importance for our application as we have provided hardware and software means to compensate for the difference at one frequency.

The insertion loss is relatively high but can be reduced by proper matching of the input and output impedances which are complex, mainly capacitive. In general, there are several combinations of matching circuits which compensate the reactive component and simultaneously transform the resistive component of the filter impedance. The matching network must have a bandwidth considerably larger than that of the filter in order to have little influence on the resulting frequency response. Otherwise it could worsen the excellent performance of the SAW filter. So it is

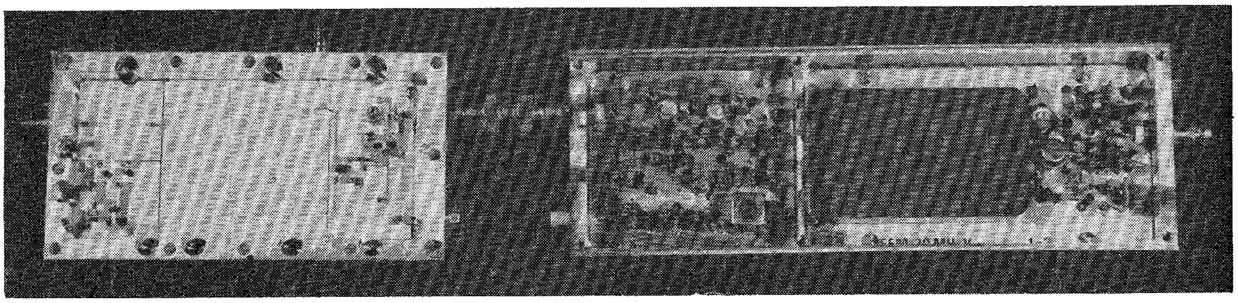

Fig. 7. One of the wide-band (1-MHz) SAW filter modules containing one filter (left), and the image rejection mixer and preamplifier module (right).

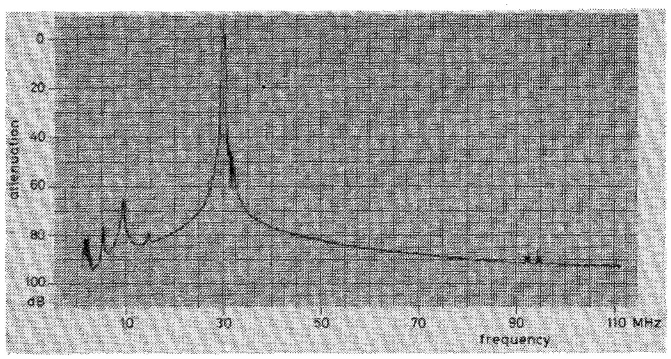


Fig. 8. The frequency response of the wide-band (1-MHz) SAW filter amplifier module.

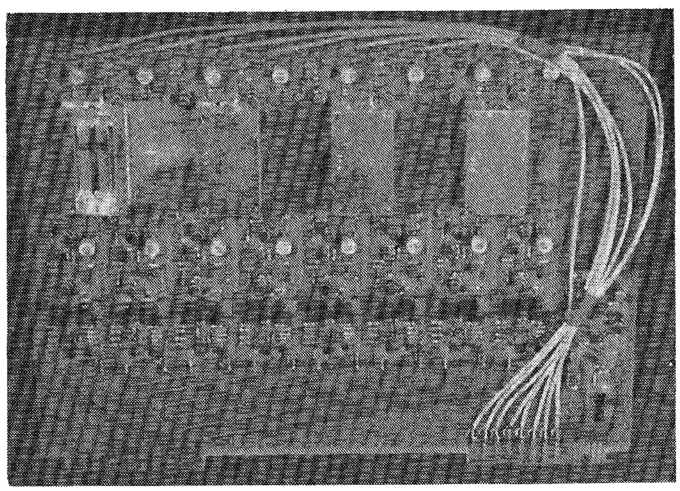

Fig. 9. One of the narrow-band (0.1-MHz) SAW filter amplifier modules, containing input and output amplifier and switches. Four filter packages, each containing two filters, are visible.

often more advantageous to admit a small mismatch, thus increasing the bandwidth at the cost of an increased insertion loss, which can be compensated for in the IF amplifier at these frequencies. For getting a high selectivity and amplification, conventional IF amplifiers use several resonant circuits or bandpass filters with amplifying stages between them. With SAW filters the design of IF amplifiers can be simplified. Since the selectivity is concentrated in the filter, the amplification can be condensed in a few high gain, wide-band devices. One of the 800 modules comprising the wide-band SAW filter is illustrated in Fig. 7. The frequency response of the module (without preamplifier) is shown in Fig. 8. The strong SAW filter response due to volume waves at 48 and 55 MHz, as well as the third-harmonic response at 90 MHz due to the use of split-electrode transducers, are suppressed by the low-$Q$ resonant circuits of the amplifier module. One of the 100 printed circuit boards comprising 4 SAW filter packages (each containing 2 filters on one substrate as shown in Fig. 3) amplifiers and switch is illustrated in Fig. 9.

## IV. Conclusion

The experimental results obtained with a large number of filters prove that the high quality of SAW filters developed in the laboratory can be realized also for larger production quantities. SAW filters are promising alternatives for phased-array radar applications. They improve not only the single-receiver performance by realizing almost ideally the matched filter concept but also the array performance by the excellent reproducibility of the frequency response and by the stability of their parameters.

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# Use of an SAW Multiplexer in FMCW Radar System

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***Abstract*—This paper describes the application of an SAW multiplexer to develop range line resolution in FMCW millimeter-wave radar systems. The basic system design concept as well as test results are presented describing the function of the SAW multiplexer in developing the multiple-range cells in the millimeter-wave terminal guidance seeker. The SAW multiplexer has 16 channels and uses the offset multistrip coupler technique for sorting the acoustic beam into various acoustic tracks according to frequency.**

## I. Introduction

RANGE RESOLUTION in FMCW radar systems has been achieved using multichannel surface acoustic wave (SAW) filter devices. This paper describes the basic system design concepts employed for developing range cells in linear FMCW radar using the SAW filters. The techniques discussed are currently being used in low-power solid-state millimeter-wave radar for air-to-ground terminal guidance of antiarmor munitions. The narrow-band SAW filters are incorporated in the receiver IF sections where each filter represents a range cell that is dimensionally proportional to range. The small size, light weight, low power consumption, and stable performance characteristics make the SAW filter an ideal choice for missile seeker applications.

## II. FMCW Technique

In a linear frequency-modulated continuous-wave (FMCW) radar [1], [2], the transmitter generates a periodic waveform having a linearly varying frequency versus time for each modulation sweep period. The frequency-modulated signal is transmitted from the radar antenna to the target where a portion of the signal is reflected back to the radar. The frequency of the return signal received by the antenna is compared with a sample of the instanta-

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