Surface Acoustic Wave Resonators on Quartz

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Abstract—The design of high-\( Q \) low-distortion surface acoustic wave (SAW) resonators requires techniques and data for minimizing acoustic losses and accurate placement of the device resonance. Several aspects of one- and two-port SAW resonator design and fabrication are discussed. It is shown that devices on quartz consisting of recessed aluminum transducers and shallow reflecting groove arrays can be fabricated with very high \( Q \) values. The properties of the shallow-grooved reflector are analyzed both experimentally and theoretically. Experimental values of the surface wave velocities for reflectors, transducers, and the free surface are reported, and transducer placement and maximum cavity length are discussed. It is shown that the maximum cavity length possible, without additional resonant modes, increases as the groove depth decreases. Cavity length corrections required to center a resonance in the reflector stopband are derived, and resonator fabrication for the recessed-transducer/shallow-grooved reflector configuration is discussed. SAW devices with \( Q \) values of 75 000 (at 71 MHz) and resonators at frequencies above one GHz (with \( Q \) values over 3200) have been fabricated.

I. INTRODUCTION

SURFACE acoustic wave resonators [1], [2] have been studied in order to reduce acoustic losses and distortions. It is shown that the recessed-aluminum transducer/shallow-grooved reflector [3]–[5] system yields excellent results (\( Q \) values in excess of 75 000 have been obtained) by reducing bulk-mode losses and transducer reflections. Velocity data are presented which permit reduction of reflector radiation losses and distortions by centering the device resonance in the reflector stopband. In addition, data are presented which allow maximizing the cavity length while limiting unwanted resonant modes to tolerable levels. Maximizing the cavity length is important in achieving higher \( Q \) values, when radiation [6] is limiting, and in permitting maximum length coupling structures in coupled mode resonators [7].

The surface acoustic wave (SAW) resonator [2], [8] configuration to be discussed is shown schematically in Fig. 1. The reflectors, consisting of arrays of shallow etched grooves, are spaced to form a resonant cavity in which one or two recessed aluminum transducers are located. The device substrates were the \( ST (42.75^\circ \text{ rotated } Y) \) cut of quartz in order to obtain temperature stability. The recessed-transducer configuration [9] virtually eliminates transducer reflections and the resultant distortion from this source [10]. Loss and distortion increase when the resonance is not centered in the reflector stopband, since radiation losses are not minimized and the resonance develops asymmetrically. Accurate velocity data are necessary to make the required cavity length corrections to center the resonance. All the devices tested were overlap-weighted (cosine on a 5% pedestal) to suppress transverse SAW modes [11], unless otherwise stated.

Discussion of the fabrication techniques of SAW devices is presented in the next section. The experimental design and testing of the devices are discussed in the following sections.
A SAW transducer within a resonant cavity will contribute to losses through finite electrode conductivity and from bulk-mode scattering associated with surface wave scattering for acoustically reflecting electrodes. Distortions are introduced by transducers due to acoustic reflectivity and by velocity differences between the transducer, the free surface, and reflector sections of the resonator. These effects are to some degree interrelated, so this discussion will center on several types of transducers that have been used. It is shown that the recessed-metal (aluminum) transducer on quartz is effective in reducing losses and distortions in resonators.

Several configurations used are shown in Fig. 2. Configuration (a), metal at the base of a groove, was used in earlier work because the fabrication was simple as the reflectors and transducers were structurally the same. However, Fig. 3 shows the resonance filter response of a two-port resonator fabricated using configuration (a) of Fig. 2, and the response is highly distorted in part due to the high acoustic reflectivity of the transducers. Transverse mode distortion is also present as the transducers were not weighted. The transducer electrodes must thus be of low acoustic reflectivity, and configurations (b)-(e) of Fig. 2 meet this requirement.

Using the split electrode configurations (c) and (d), the maximum device frequencies are limited to about 500 MHz due to optical photolithographic resolution limitations, and the thin metal electrode configuration has been shown to be satisfactory only at frequencies below several hundred MHz.

An electrode configuration, yielding very low acoustic reflectivity, a minimum of fabrication problems, and low bulk-mode scattering loss, is the (fully) recessed aluminum electrode configuration of Fig. 2(e). Recessed transducer electrodes have very low acoustic reflectivity, and these are found to be of significant value in resonators. Evidence of the relative effectiveness of recessed and thin-metal transducers is given in Fig. 4 where the admittance magnitudes (in air) of two one-port resonators are shown. Both devices were fabricated using the same mask, and each had etched groove reflectors of approximately the same depth (1400 Å for the recessed device and 1600 Å for the thin-metal device). The only significant difference in the structures was that the thin transducer was 700 Å thick. All depth and thickness measurements were made using surface analysis equipment with a measurement accuracy of approximately 5%. The admittance curves were
Fig. 3. Two-port resonator resonance transmission response for electrodes with metal at base of groove.

Fig. 4. Admittance magnitudes of one-port SAW resonators fabricated using alternate transducer configurations.

Fig. 5. 800-MHz SAW delay line fabricated using recessed transducers, with quarter wavelength, 1000 Å thick, aluminum electrodes. The transducers had 200 electrodes each.

made with a low capacitance (0.07 pF) probe associated with a network analyzer and a frequency synthesized signal source. The $Q$ values listed in Fig. 4 are thus for electrically unloaded devices. The results in Fig. 4 clearly show the improvement in device performance obtainable using recessed transducers. The recessed transducers, which contribute significantly to improvement of resonators, are shown to be of low acoustic reflectivity also by the response in Fig. 5 of an 800-MHz SAW delay line. The curve of Fig. 5 shows little evidence of distortions due to electrode reflections though significant direct feedthrough ripple is present.
The resonance transmission response of a two-port resonator fabricated with recessed transducers is given in Fig. 6. Note the highly symmetric, low distortion resonance as compared with the result in Fig. 3. Additional data concerning the device parameters and the reflector stopband characteristic are given in Fig. 6. The $Q$ value (14 000) for this device was obtained with the transducers unmatched and loaded by a 50-$\Omega$ network analysis system.

III. ETCHED GROOVE REFLECTORS

The surface wave reflector, which serves to contain the energy in a resonant cavity, loses energy through radiation (transmission through the reflector) and by scattering energy into bulk acoustic modes. Etched groove reflectors on quartz will be treated, including a discussion of loss reduction techniques. It will be shown that efficient reflectors can be made that consist of long arrays (between 500 and 2000 stripes) of shallow grooves ($0.5 < h/\lambda < 1.5$), that is, long arrays of very lightly reflecting discontinuities [3], [4]. This study has been restricted to grooves since it is believed, but it has yet to be proved, that the aging rate for this type reflector will be lower than other types (metal or ion implanted stripes, etc.) when the possible aging mechanisms are considered.

The operational characteristics of a grooved reflector are determined primarily by the depth of the grooves ($h$); the total number of grooves ($N$), that is, the reflector length; and the period of the grooves ($\lambda_0/2$). Secondary factors are the groove profile, which has been shown [13], theoretically at least, to not be critical, and the groove width to period ratio ($2W/\lambda_0$). The effect of this ratio has been studied [6], and it has been found that the maximum reflectivity occurs when $2W/\lambda_0$ is less than 0.5 by a small amount depending on the groove depth.

Bulk acoustic mode losses ($L_{BG}$) which occur when a surface wave is reflected from a discontinuity are known to increase with the size of the discontinuity, specifically [14] $L_{BG} \approx (h/\lambda_0)^2$. An approach to reducing $L_{BG}$ is thus to reduce the groove depth. If each groove is shallow and very lightly reflecting, then in order to obtain sufficiently low radiation losses there must be a large number of grooves—hence, long arrays. An optimum groove depth for a given reflector length is that which yields bulk-mode losses slightly smaller than the most limiting loss mechanism for the frequency range considered. In the lower frequency range ($10 < F \lesssim 300$ MHz), the material losses are fairly low, and extremely shallow grooves may be necessary ($h/\lambda_0 < 0.5\%$) to reach the material $Q$ limit in device performance. This would dictate very long arrays, and the overall device size may thus be the limiting factor. At much higher frequencies ($800$ MHz $< F$) the material losses increase considerably, and it may be possible to have somewhat deeper grooves ($2\% < h/\lambda$ perhaps) and still keep the bulk-mode losses less than losses due to the material. This situation works in the designer's favor also because deeper grooves permit thicker transducer electrode metalization with a consequent lowering of resistivity losses. Additionally, if thicker electrodes are desired than are compatible with the reflector groove depth, the configuration of Fig. 7(b) can be used, rather than that of Fig. 7(a) at the expense of a somewhat more difficult fabrication process. An optimum groove depth for a given re-
The transmission characteristic of a shallow-grooved reflector is shown in Fig. 8 with important parameters labeled and the basic experimental configuration. Though the power reflectivity \( R \) is of primary interest, the transmission response \( T \) is more easily measured. The quantities \( R, T, \) and reflector losses are related by

\[
R + T + \text{loss} = 1, \tag{1}
\]

and an indication of the experimental shape of and location (with frequency) of \( R \) can be had through a knowledge of \( T \). Since the loss due to bulk modes can only be estimated, the measured values of \( T \) can be used only to establish an upper limit on \( R \).

The source of much of the transmission data to be presented was taken from one- or two-port resonators (which have auxiliary transducers added) as shown in the inset diagram of Fig. 8. Rubber cement, an effective and easily removed acoustic absorber, was placed on the outside reflector, allowing one to determine the transmission properties of the reflectors for an operative device. In many cases, the transmission and resonance properties have been superimposed on the same graph (an example of which was given previously in Fig. 6) which is of value in analyzing the performance of a device.

An effective method [15] of analyzing the reflectivity of a groove is to incorporate the effect of the vertical discontinuities into a hypothetical acoustic impedance change of the groove. With this identification either a network analysis [16]-[18] or a coupled mode approach [19] may be used to determine the grating performance. The deviation of the ridge to groove impedance ratio from unity (the quantity \( \varepsilon \)) [18] is assumed to be much smaller than unity, and it is central to an experimental evaluation of reflectors. The absolute value of \( \varepsilon \) may be obtained experimentally (neglecting the effect of losses) from the reflector stopband depth \( |\Delta P| \) of Fig. 8 by use of

\[
|\Delta P| = 8.686 N|\varepsilon| - 6.02 \text{ db}. \tag{2}
\]

Equation (2), which has been derived using the coupled mode [19] approach, is valid for \( N|\varepsilon| > 1 \), where \( N \) is the total number of reflecting elements. Since \( |\varepsilon| \) is small we may assume that it is directly proportional to the relative groove depth \( (h/\lambda_0) \), that is,

\[
|\varepsilon| = \alpha \frac{h}{\lambda_0}, \quad \text{for } |\varepsilon| << 1. \tag{3}
\]

An analysis of the transmission characteristics of numerous reflectors yield the result

\[
\alpha = 0.50 \pm 0.02 \tag{4}
\]

which is slightly smaller than the value (0.54) previously reported [20]. Using (2)-(4) one may readily calculate the radiation loss at stopband center and using (1) obtain an upper limit on the reflectivity. For instance, a reflector with \( N = 1060 \) and a 1% groove depth \( (|\varepsilon| = 0.005) \) will have a radiation loss of 0.01% and a maximum reflectivity of 99.99%.

The reflector need not be this long if it can be shown that another loss mechanism exceeds 0.01%. At lower frequencies \( (f < 100 \text{ MHz}) \) the physical length of reflectors can be prohibitively large, creating a design problem because it is in this frequency range that the radiation and bulk-mode losses may be limiting since material losses are low. In order to decrease the radiation and bulk losses, the grooves must be made more shallow and the reflectors longer. Thus the maximum allowable device size may place a lower limit on the losses and therefore the maximum \( Q \) attainable at low frequencies.

The second reflector characteristic of importance is the SAW velocity at the stopband center \( v_c \). This velocity is found experimentally by measuring the frequency \( (f_0) \) at the stopband center and using \( v_c = f_0 \lambda_0 \), where \( \lambda_0 \) is twice the reflector period. It has been shown [15] that \( v_c \) will deviate from the
free surface velocity $V_F$ by a factor proportional to the square of the normalized groove depth, that is,

$$v_c = V_F \left(1 - K_\phi \left(\frac{h}{\lambda_0}\right)^2\right).$$

The velocity $v_c$ is smaller than $V_F$ due to the reactive storage [15] of acoustic energy at the groove edges. Experimental values of $K_\phi$ have been obtained from devices with different etch depths for frequencies between 70 and 150 MHz. An experimentally verified free surface velocity value (to be discussed shortly) was used in the calculation. The average value found for a large number of devices with etch depths varying from 0.4 to 2.8% $\lambda_0$ is given in

$$K_\phi \text{ (average)} = 10.3$$

with a standard deviation of 1.6. For several devices with very shallow grooves ($h/\lambda \approx 0.5\%$), the computed value of $K_\phi$ deviates from (6) by up to 50%. However, the velocity change from the free surface value is less than 1 m/s for such shallow grooves, and the uncertainties in the free surface and stopband center velocities, estimated to each be ±0.2 m/s, account for this deviation. As the reflector velocity must be known primarily to correct the cavity length to center the resonance, uncertainties in $K_\phi$ lead to higher order errors, which may generally be neglected.

The bandwidth of the reflector stopband, $\Delta F$ of Fig. 8, may be derived from the coupled mode theory. An expression for the normalized bandwidth ($\Delta F_n = \Delta F/f_0$) previously [21] published in a differing notation, is given in (7). Experimental bandwidth data

$$\Delta F_n = \frac{2}{\pi} |\epsilon| \sqrt{1 + \frac{\pi}{N |\epsilon|}^2}$$

are presented in Fig. 9 along with the theoretical curve calculated from (7). The values of $|\epsilon|$ were obtained from the experimental values of $\Delta P$, through (2), except in a few cases where accurate measures of $\Delta P$ were not available. In these latter cases, (3) and (4) were used to determine $|\epsilon|$.

It is seen that there is generally close agreement between theory and experiment with a few data points at considerably wider than expected bandwidths. For a constant transmission factor given by (2) (corresponding to an approximately constant reflectivity), (7) shows that $\Delta F_n$ varies linearly with $|\epsilon|$ and the groove depth. Thus the reflector bandwidth narrows with decreasing groove depth, and this will be shown to allow longer cavities without introducing unwanted resonant modes.

### IV. Resonator Cavities and Velocities

Other resonator parameters studied are the transducer ($V_T$) and free-surface ($V_F$) velocities, transducer placement for maximum coupling, and maximum cavity length (to limit additional resonance modes).

The free-surface velocity has been found by measuring the wavelength from edge to edge, for a system such as is shown in the lower part of Fig. 10. The response curve of Fig. 10 is an example of the data from which calculations have been made. Combining the resonant phase matching conditions

$$2k_R L_{CE} - 2\phi_R = 2M_F \pi,$$

where $k_R = 2\pi/\lambda_R$ is the resonant wavevector, $\phi_R$ is the reflection coefficient phase at resonance, and $L_{CE} = M_F \lambda_0/2$ is the edge to edge cavity length, with a known [18] expression for $\phi_R$, and the velocity $V_F$ is found to be given by

$$V_F = \frac{f_R \lambda_0}{1 + \frac{f_R - f_0}{M_F f_R |\epsilon|}}.$$
In (9), \( f_0 \) is the reflector stopband center frequency of Fig. 8. Using (9), the velocity \( V_F \) has been found to be

\[
V_F = 3157.8 \pm 0.2 \text{ m/s}, \quad \text{ST-}x\text{-propagating quartz} \quad (10)
\]

which equals the published [22] value of 3157.6 m/s. The value of \( V_T \) from (9) has been found not to be sensitive to the value of \( |e| \) but is quite sensitive to the difference \( f_R - f_0 \).

The recessed transducer velocity has been measured for depths in the range \( 0.3\% \leq h/\lambda_0 \leq 1.5\% \). The value of \( V_T \), given in

\[
V_T = 3154.6 \pm 0.4 \text{ m/s} \quad (11)
\]

has been found to be independent of the depth; that is, this velocity is dispersionless over the range of depths tested. The velocity \( V_T \) has been found using two methods. The first involved the measurement of the two-port resonant frequency \( (f_R) \) and application of (12):

\[
V_T = \frac{f_R \lambda_0}{M_F + M_T - M_T V_F} = \frac{f_R - f_0}{M_T \sqrt{M_T f_0 |e|}}. \quad (12)
\]

In (12), which may be derived in a manner similar to (9), \( M_T \) is the total number of half periods in the cavity, and \( M_F \) is the number of free space half periods. The value \( |e| \) used in (9) and (12) was the experimental value when available; otherwise it was calculated from (3). It has also been found that \( V_T \) can be measured to within 0.5 m/s by determining the frequency \( f_M \) of Fig. 10 and using \( V_T = f_M \lambda_0 \), where \( \lambda_0/2 \) is the transducer period. A ripple-free distortionless \( \sin x/x \) response for a delay line consisting of a set of unweighted transducers is necessary in order to measure \( f_M \) accurately. Most of the resonators fabricated include a set of auxiliary unweighted transducers, one on either end of the resonant structure, which are generally made 50 \( h \) long. In addition to being useful in experimental analysis of the reflectors, these auxiliary transducers frequently yield response data sufficiently distortion free to permit calculation of \( V_T \).

It is generally necessary to adjust the resonator free space cavity length to center the resonance in the stopband, thus minimizing losses and distortion. Given a cavity containing transducers totaling \( M_T \) half wavelengths long and free sur-
face totaling $M_F$ half wavelengths before correction, we require the change ($\delta M_F$) in the free-surface length necessary to cause resonance to occur at $f_0$. Applying the phase-matching condition in a manner similar to (8) and noting that at resonance the reflection coefficient phase $\phi_R$ is zero, it is found that $\delta M_F$ is

$$\delta M_F = (M_F + M_T) K_1 \left( \frac{h}{\lambda_0} \right)^2 - M_T \frac{V_F - V_T}{V_F}. \quad (13)$$

The physical additional length ($\delta L_{CE}$, which is generally positive) is then given by

$$\delta L_{CE} = \delta M_F \cdot \frac{\lambda_0}{2}. \quad (14)$$

The transducer position for maximum coupling has been found by noting the response of two-port resonators in which the cavity edge to electrode center spacing was either $I(\lambda_0/2)$ or $(I + \frac{1}{2})(\lambda_0/2)$, where $I$ is an integer. With a reflection phase of $\phi_R = 0$, for the variable to which the transducer responds (the normal component of the potential wave [8]), the required electrode position (Fig. 1) is given by

$$L_{FC} = I \frac{\lambda_0}{2}. \quad (15)$$

It has been demonstrated [12] that transducers in cavities on quartz with grooved reflectors must be placed according to (15) to obtain maximum coupling to the potential wave. This placement is consistent with the requirement to position the electrode centers on the maxima of the resonant potential standing wave. Transducer placement for other types of reflectors [12] is given in Table I for reference purposes.

Since the piezoelectric coupling on quartz is small, metal stripes present a mass loading effect (for gold) or a physical discontinuity analogous to a ridge (for aluminum) whether the stripes are connected or not.

Calculations and experiments have been performed to determine the maximum cavity length. One calculation scheme involved using the equivalent circuit model [23] to determine the filter response of a two-port resonator for designs with specific groove depths and increasing length cavities. This procedure may be understood with reference to Fig. 11. The levels, relative to an established reference, of additional resonant modes were noted for increasing cavity lengths, and the maximum cavity length was defined to be that at which the additional modes reached a specified value (5 dB). In Fig. 11, the maximum $L_{CE}$ is $190 \lambda_0$ for $\epsilon = 0.003$. Similar calculations were performed for other parameter values, and the results are summarized in the equivalent circuit model curve of Fig. 12.

Another computational scheme, based on the coupled mode analysis, was used to calculate the second curve of Fig. 12. This procedure involved the phase matching condition, similar to (8), for the central resonance of mode number $K_c$ and for next higher resonance of mode number $K_c + 1$. The reflection phase of $K_c$ is zero for a resonance frequency ($f_c$) centered in the stopband, but the reflection phase $\phi_{K1}$ at the $K_c + 1$ resonance frequency ($f_{K1}$) is nonzero. It is due to the non-zero value of $\phi_{K1}$ that the first undesired resonances to arise in the device response are those corresponding to $K_c \pm 1$. 

Fig. 11. Calculated response for 2 two-port resonators differing only in length of cavity ($L_{CE}$).
Solving the set of phase matching conditions for $K_e$, it is found that maximum value of $K_e$ is given by [24]

$$K_e(\text{max}) = \frac{\phi K_1(\delta_f, \varepsilon)}{\delta_f} + \pi.$$  \hspace{1cm} (16)

To compute $K_e(\text{max})$ for a given $\varepsilon$ it is necessary to establish the reflectivity value $R(= 0.9$ has been found to correspond approximately to the 5-dB $K_e + 1$ mode level of the equivalent circuit analysis) which determines the normalized frequency deviation $\delta_f = \pi(f_{K_1} - f_f)/f_f$. The maximum cavity length was then taken to be $L_{MX} = K_e \lambda_0/2$, under the assumption that $K_e$ is the same as, or very close to, the number of half wavelengths in the cavity. This condition will generally hold since the various velocities do not greatly differ. $K_e(\text{max})$ was found to be virtually independent of $N$ for $N > 500$, so a value of $N = 1000$ was used to calculate the coupled mode curve of Fig. 12. Experiments were also performed using two-port resonators to determine maximum cavity lengths. Three data points are given in Fig. 12. These data were taken using devices in which approximately 200 grooves adjacent to the cavity in one reflector were filled with aluminum during fabrication to match the transducer structure. These filled grooves were essentially reflectionless and effectively extended the cavity length. The $K_e \pm 1$ mode resonance levels were measured for successively shorter cavities, which were formed by etching the aluminum from sets of grooves farthest from the cavity.

**V. Resonator Fabrication**

The recessed-transducer/grooved-reflector system lends itself to a simple fabrication process. It is necessary in resonator fabrication that no critical photolithograph mask realignment be performed since each of the device components (reflectors and transducer) must be positioned with respect to one another to an accuracy of a few hundredths of a wavelength or better. This accuracy cannot generally be attained using present methods to superimpose masking steps. The recessed-transducer/grooved-reflector system requires no critical mask realignments and will not need a second mask alignment for some designs.

The process steps, which have been outlined previously [3], [4], are presented here in more detail and may be understood with reference to Fig. 13.

**Step (a):** Coat the substrate with photoresist (PR), expose the entire pattern (reflectors plus transducers) in the PR using the primary mask with ultraviolet light (UV), and develop the photoresist pattern. (Process step (a) of Fig. 13)

**Step (b):** Etch the grooves in the substrate over the entire pattern utilizing the PR as a shield. The etching may be performed using one of several methods, such as a directed-ion beam source or a radio frequency (RF) generated source of ions. (Process step (b) of Fig. 13) After etching, one of two possible process paths may be employed, and the path chosen depends on the separation between the reflectors and the transducers. If the reflector-transducer (R-T) spacing is greater than 150 microns or so, it is possible to paint a protective layer of PR over the reflectors. Path 2 may then be followed. If the R-T gap is small, however, the more complex path 1 must be taken.

**Step (c2):** Apply a protective layer of PR (by painting) over the reflector sections only to prevent metal deposition in the reflector grooves. Then metalize to the desired (depth $= 200$ Å of chrome for adhesion plus aluminum sufficiently thick to fill the transducer grooves). Lift off the entire pattern (i.e., dissolve the PR in a solvent), and the final desired structure is produced as shown in (f).

The following steps must generally be followed when the R-T gap is too small to permit painting PR on the reflectors only without also partially covering the transducers.

**Step (c1):** Following step (b), metalize the entire pattern (200 Å chrome plus aluminum thick enough to fill the grooves) and lift off.

**Step (d1):** Apply a secondary layer of PR over the entire pattern, expose only the reflector sections with UV light utilizing a specially designed secondary mask. Note that this masking step is not critical since the R-T gap can always be made at least 5 or 10 microns and the gap will always be greater than or equal to $3/8 \lambda_0$. (Fig. 13, step d1) Develop the exposed PR exposing the recessed metal in the reflector sections only.

**Step (e1):** Chemically etch the metal from the reflector areas. The unexposed PR acts as a shield to protect the recessed metal in the transducers from being etched away. Remove the remaining PR (Fig. 13, step e1) The structure remaining is that desired and is shown in Fig. 13 (f).

Etching of the grooves has been performed by RF sputter etching [26] using argon or a reactive gas [27] (CHF$_3$ for quartz), and using directed ion beams [28]. Ion beam etching,
utilizing argon or a heavier noble gas, is used almost exclusively as the process permits attaining etch rates which are reproducible to better than 2%. Also, the ion beam energy can be controlled by varying the beam voltage. This capability allows the use of beam energies \( E < 300 \text{ electron volts} \), during the etching run or in the terminal phase of the process, that minimize substrate damage, thus reducing the long term aging rate from this source.

Reactive RF sputter etching with \( \text{CHF}_3 \) is a reproducible (±5%) process which we have used to obtain very deep (in excess of 4 microns) grooves. This is possible because the etch rates of quartz greatly exceed (by factors up to 10 or more) those of the masking resist in the \( \text{CHF}_3 \) plasma. Reactive sputter etching has been used to produce deeply etched surface skimming bulk mode [29] resonators \((Q\) values in excess of 2000 have been obtained in this laboratory), and this etching process may be of use for low-frequency resonators. Also, it has been shown [30] that very high \( Q \) SAW resonators can be produced using deeply etched grooves, though a different reactive etching process was used.

We have found that it is difficult to obtain reproducible etch rates using RF sputter etching with argon in large part due to the need to strictly control secondary vacuum system parameters. For instance, very stringent vacuum system operational procedures were necessary to adequately reduce the partial pressure of oxygen. However, with care, etch rates reproducible to approximately 5% may be attained.

Photoresist (Shipley AZ1350 [31] or equivalent) has been used to form the masking patterns. The etch rates of quartz and resist generally differ by less than 30% for all etching processes, except the \( \text{CHF}_3 \) sputtering as noted above. Adequate heat sinking of the substrates is necessary to ensure the resist does not overheat and flow or bake on.

### VI. Results Achievable

The techniques described here have been applied to produce resonators with high \( Q \) values and at frequencies above one GHz. In Fig. 14, the resonant admittance characteristic of an electrically unloaded one-port resonator is shown. The series resonant \( Q \), calculated by a method discussed previously, [5], is 75 300. This device had an \( 80-\lambda_0 \) cavity, \( \lambda_0 = 44 \text{ microns} \), and the reflectors were etched to \( h/\lambda_0 = 1\% \) and contained in excess of 1000 grooves. The transducer was \( 12 \lambda_0 \) long and cosine weighted. The response, taken in vacuum, was centered in the stopband through application of the cavity length correction (14) during design.

The response shown in Fig. 15 is that of a two-port resonator filter with a resonance at 1.009 GHz. The loaded \( Q \), in vacuum, between 3-dB points is 3250, and this is to be compared with a material \( Q \) of about 10 000 [6] at 1 GHz. This device had two 15-\( \lambda_0 \) transducers in a 90-\( \lambda_0 \) cavity. The wavelength was 3.12 microns (0.78 micron lines and spaces), and the etch depth was 600 \( \mu \). The device fabrication was performed with optical techniques, following the procedures of Section V using low-reflectivity chrome masks.

### VII. Summary

Data and fabrication techniques for the production of low-loss distortion-free resonators on quartz have been presented.

It is shown that recessed aluminum transducers and shallow-grooved reflectors yield excellent resonator performance. Details of reflector performance are given, and it has been shown that shallower grooves permit the use of longer cavities allowing longer multicavity coupling structures and the possibility of higher \( Q \) values (when radiation losses limit the \( Q \)). Velocity data for reflectors, transducers, and the free surface have been obtained. An expression for the cavity length increase required to center the resonance in a stopband has been obtained, and
Fig. 14. Admittance characteristic about resonance frequency of one-port resonator on quartz fabricated per Fig. 1. The series resonant $Q$ is 75 300.

Fig. 15. Two-port resonator filter response for high-frequency ($F = 1.009$ GHz) device fabricated using recessed transducers and long shallow-grooved reflectors. The device $Q$ was 3200 and the etch depth $(h/\lambda_0)$ was 2%. 
values are readily calculated. Optimum transducer placement and reflector length are discussed. Fabrication procedures are outlined, and it is shown that the recessed-transducer/grooved-reflector configuration is compatible with relatively simple processes. Devices with $Q$ values exceeding 75,000 (at 71 MHz) and resonators operating in excess of 1 GHz have been fabricated.

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