

A Simple Transducer Equivalent Circuit Parameter Extraction Technique

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Abstract - This paper presents a simple technique for extracting the fundamental SAW transducer equivalent circuit parameters from measurements easily performed on a network analyzer. This technique allows for the separation and unambiguous determination of basic equivalent circuit parameters such as parasitic (thin film) resistance, transducer capacitance, acoustic susceptance, and Hilbert Transform conductance. It is applicable to any transducer geometry, but is particularly useful in model development for unconventional transducer structures, such as narrow gap or multilevel devices. In such cases, quantities such as transducer capacitance and effective coupling may not be easily calculated from previously known relationships. The basic theory for parameter extraction is presented, along with a comparison of experimental data and modeled device performance calculated using the extracted values.

INTRODUCTION

Over the past twenty-five years, numerous variations of the Mason equivalent circuit for the in-line-field and the crossed-field approximations presented by Smith et. al. [1] have been used to model the behavior of many type of SAW devices. The current work utilizes one form of this equivalent circuit model as an analytical and predictive tool for determining the effective material parameters for standard and unconventional transducer structures, and for simulating device performance.

EQUIVALENT CIRCUIT CONFIGURATIONS

The basic equivalent circuit used in this work in the parallel form consists of the conventional terms for

total electrode capacitance (C_t), two terms which represent acoustic wave excitation, the acoustic conductance $G_a(\omega)$ and the acoustic susceptance $B_a(\omega)$, and an additional series resistance which represents the parasitic thin film resistance (R_p). As usual, the electrical behavior of this circuit can be represented by either the series or the shunt version of the circuit shown in Figure 1. These two descriptions are equivalent, and throughout the equivalent parameter extraction process we will convert back and forth from the series to the parallel equivalent circuit, as necessary to facilitate our calculations.

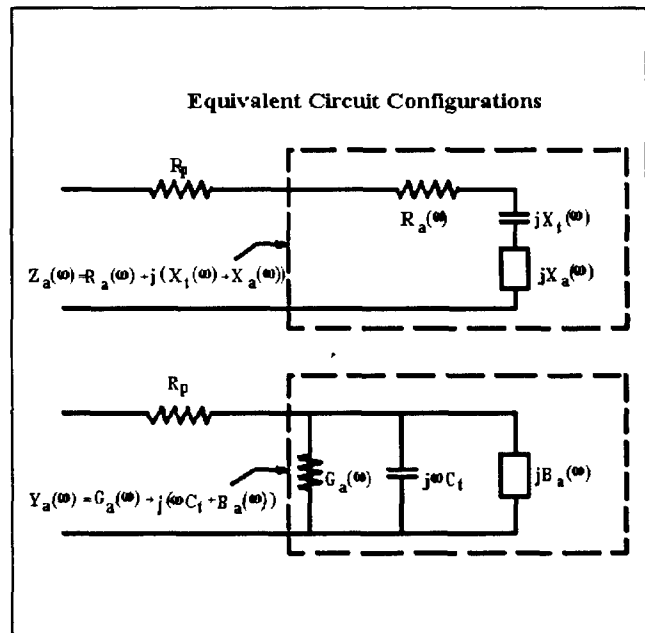


Figure 1: Series and shunt equivalent circuits.

The input impedance of the series equivalent circuit shown in Figure 1 can be written

$$Z = R_p + Z_a(\omega) \quad (1)$$

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where $Z_a(\omega)$ represents the acoustic impedance of the transducer, given by

$$Z_a = R_a(\omega) + j(X_f(\omega) + X_a(\omega)) \quad (2)$$

where $R_a(\omega)$ represents the acoustic resistance, and the total transducer reactance due to acoustic and capacitive effects is given by $(X_f(\omega) + X_a(\omega))$. The series input resistance and reactances are related to the acoustic conductance $G_a(\omega)$, the acoustic susceptance $B_a(\omega)$, and the transducer capacitance C_t by

$$R_a(\omega) = \frac{G_a(\omega)}{G_a^2(\omega) + (\omega C_t + B_a(\omega))^2} \quad (3)$$

$$X_f(\omega) = \frac{-\omega C_t}{G_a^2(\omega) + (\omega C_t + B_a(\omega))^2} \quad (4)$$

$$X_a(\omega) = \frac{-B_a(\omega)}{G_a^2(\omega) + (\omega C_t + B_a(\omega))^2} \quad (5)$$

For an unweighted transducer, the acoustic conductance can be written

$$G_a(\omega) = G_o \left[\frac{\sin(X)}{X} \right]^2 \quad (6)$$

and the susceptance is given by

$$B_a(\omega) = G_o \left[\frac{\sin(2X) - 2X}{2X^2} \right] \quad (7)$$

In the above equations,

$$X = \frac{\pi N(f - f_o)}{f_o} \quad (8)$$

and the center frequency transducer conductance is ideally

$$G_o = 8k^2 \gamma C_f f_o W N^2 = 8k^2 \gamma \frac{C_t}{\zeta} f_o N \quad (9)$$

where N represents the number of finger pairs, k^2 is the substrate piezoelectric coupling coefficient, ζ is a constant that takes into account transducer finger structure ($\zeta=1$ for non-split fingers, $\zeta=1.414$ for split fingers), γ is the ratio of normalized center frequency conductance relative to a non-split device ($\gamma=1.0836$ for a split finger transducer with 50/50 line/period ratio), W is the transducer aperture,

and C_t is the substrate equivalent dielectric constant [2,3,4]. This implies the transducer capacitance can be written

$$C_t = \zeta N W C_s \quad (10)$$

The admittance of the parallel equivalent circuit can be written as

$$Y(\omega) = \frac{1}{Z}(\omega) = \frac{1}{R_p + Z_a(\omega)} \quad (11)$$

The parasitic resistance term, however, complicates this expression relative to the acoustic admittance

$$Y_a(\omega) = \frac{1}{Z_a(\omega)} = G_a(\omega) + j(\omega C_t + B_a(\omega)) \quad (12)$$

PARAMETER EXTRACTION TECHNIQUE

The basic equivalent circuit parameters for the model presented in figure (1) can be found from simple S-parameter measurements. A full two-port calibration or two full one-port calibrations can be performed. Experimental results in the current work indicate no measurable difference in the S11 and S22 magnitude and phase measured using these two calibration techniques. Separate one-port calibrations are thus used throughout for simplicity. Magnitude and phase data are collected for S11 and S22 over a span at least two to three times as wide as the null bandwidth of the filter. This S-parameter data is then converted to port input impedances using the relationship

$$Z_j = Z_{oj} \left[\frac{1 + S_{jj}}{1 - S_{jj}} \right] \quad (13)$$

The transmission line characteristic impedance Z_{oj} used for these measurements was 50Ω. The real and imaginary portions of the input impedance can then be found. Figure (2) illustrates S11 and S22 data and the corresponding impedances for a 60 MHz unweighted split-finger transducer with aperture $W=55\lambda$ and $N=5$ built on 112° Lithium Tantalate. Although relatively accurate values were known for this substrate, this parameter extraction approach does not require any a priori knowledge of the substrate constants C_s and k^2 . These values, along with velocity, can easily be determined from measurements

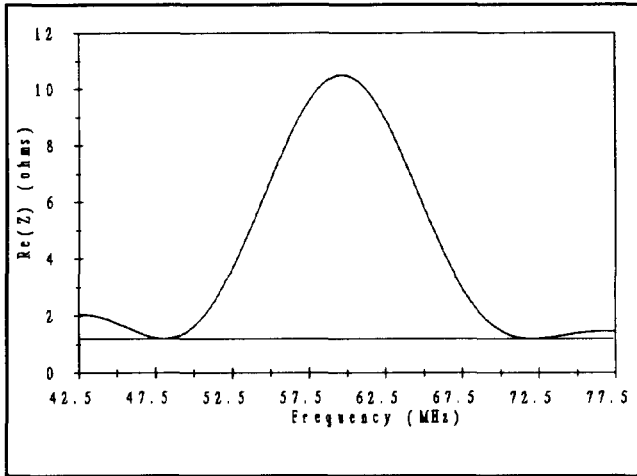


Figure 2: Real part of input impedance for 60 MHz unweighted LiTaO₃ device (calculated from S11).

provided the geometry of the device is known.

The parasitic resistance R_p in equation (1) can easily be found as the constant offset in the real part of Z , and subtracted from the transducer input impedance, yielding $Z_s(\omega)$ [5]. This acoustic impedance can then be inverted to yield the acoustic input admittance $Y_s(\omega)$. A linear fit to the values of $\text{Im}(Y_s(\omega))$ in frequency ranges away from device center frequency can be used to extract the total

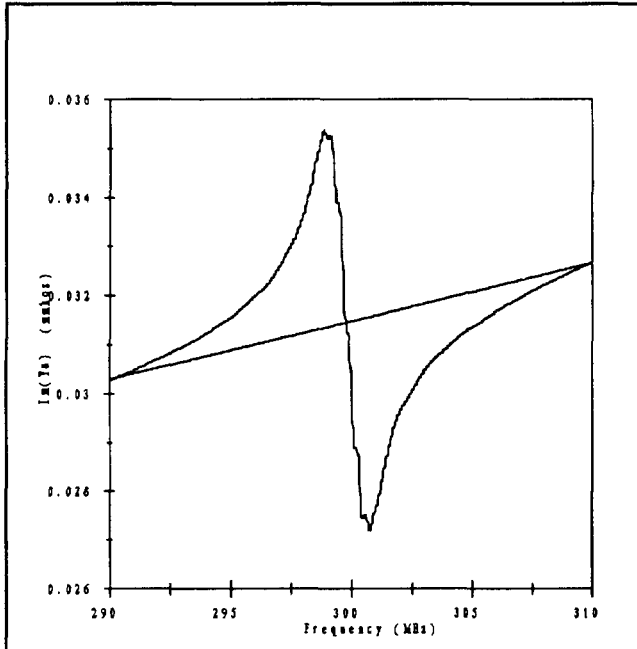


Figure 3: Extraction of total capacitance using linear fit for 300 MHz apodized device on ST-Quartz.

measured capacitance C_{meas} as shown in Figure (3). This measured capacitance is the sum of the transducer, header, and bond pad capacitances. Header capacitance can be measured directly. If the material constants are initially unknown, a value for C_s can be obtained using

$$C_s = \frac{C_{meas} - C_{header}}{\zeta NW + \left(\frac{A_{pads}}{h} \right)} \quad (14)$$

where A_{pads} is the total bond pad area on the transducer and h is the wafer thickness. Utilizing this extracted value of substrate equivalent dielectric constant, the transducer capacitance can be calculated using equation (10).

Once C_s and C_t have been found, the experimental transducer acoustic Q (Q_t) can be used to find k^2 . Q_t is related to C_t and the center frequency value of the acoustic conductance $G_s(\omega_c)$ by

$$Q_t = \frac{\omega_c C_t}{G_s(\omega_c)} = \frac{\pi}{4k^2 N} \quad (15)$$

$G_s(\omega_c)$ can be measured as the real part of the acoustic input admittance $Y_s(\omega_c)$. Solving for k^2 and substituting k^2 and C_s into equation (9) yields a value for transducer center frequency conductance. Using the extracted values of G_o and C_t and equations (3) through (8) and (12), the real and imaginary parts of the input impedance and admittance can be predicted. If desired, these values can be related to original S-parameters for direct device performance modeling.

COMPARISON WITH EXPERIMENT

Several device structures and substrates were evaluated using the parameter extraction technique described above. A 60 MHz unweighted split-finger transducer with aperture $W=55\lambda$ and $N=5$ built on 112° Lithium Tantalate was the initial test structure. Data was taken on this device over temperature from -50°C to $+85^\circ\text{C}$ in order to evaluate the temperature dependence of the material and circuit parameters extracted. Figures (4) through (6) illustrate the results of the circuit extraction over temperature for this device. The values obtained for transducer parasitic

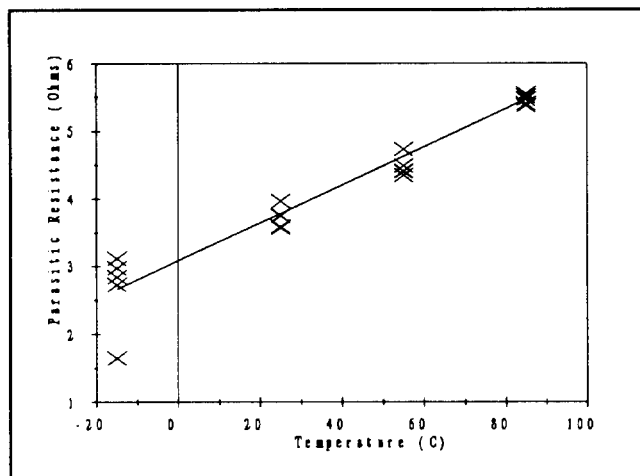


Figure 4: Temperature dependence of Parasitic Resistance (1500 Å Al on 112° Lithium Tantalate).

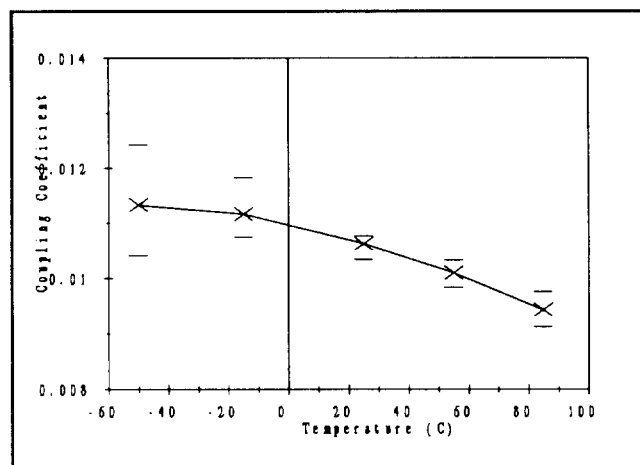


Figure 5: Temperature dependence of piezoelectric coupling coefficient k^2 (112° Lithium Tantalate).

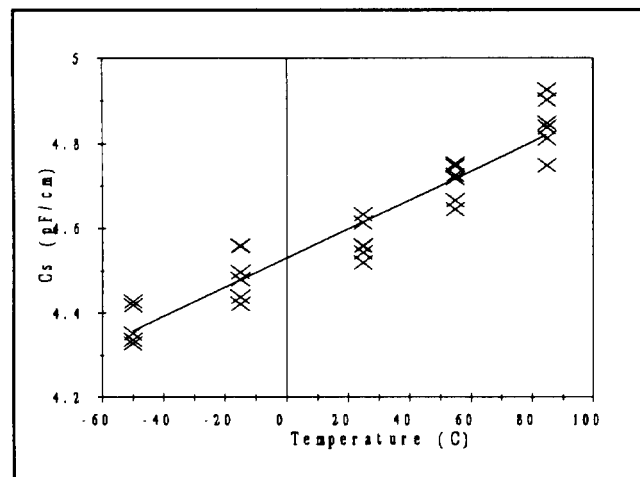


Figure 6: Temperature dependence of substrate effective dielectric constant C_s (112° Lithium Tantalate).

resistance, substrate capacitance and acoustic coupling coefficient for the Lithium Tantalate substrate used were found as the average of the extracted values for at least 5 devices. The temperature coefficient of resistivity for the 1500 Å thick aluminum film was found to be $0.00908 \Omega/^\circ\text{C}$, in good agreement with prior published values [6]. Room temperature values of extracted material constants for this cut of Lithium Tantalate are $k^2 = .009996$ and $C_s = 4.569 \text{ pF/cm}$, again close to widely used values. It is interesting to note that the standard deviation of these measurements was less than 2% of these mean values. Figure 7 shows a comparison between the measured and values of the input impedance for this transducer and values predicted using these extracted quantities. Good correlation between the responses is seen.

The same equivalent circuit parameter extraction technique can be used to obtain center frequency values for apodized transducers which are either amplitude weighted, or are overlap weighted but have no time side lobes. In this case, an effective number of transducer finger pairs N_{eff} can be found easily by matching the measured null bandwidth with the null bandwidth of an ideal unweighted transducer. N_{eff} is used in place of N in the process described above. For more complicated apodization patterns, N_{eff} can be calculated based on the actual weighting function used. A 300 MHz device using apodized and withdrawal weighted transducers fabricated on ST-Quartz was analyzed using the present technique. Once again, good comparison with known values was obtained for the substrate parameters, and the predicted device center frequency performance is in good agreement with experiment, as shown in Figure 8. Room temperature values of extracted material constants for ST-Quartz are k^2

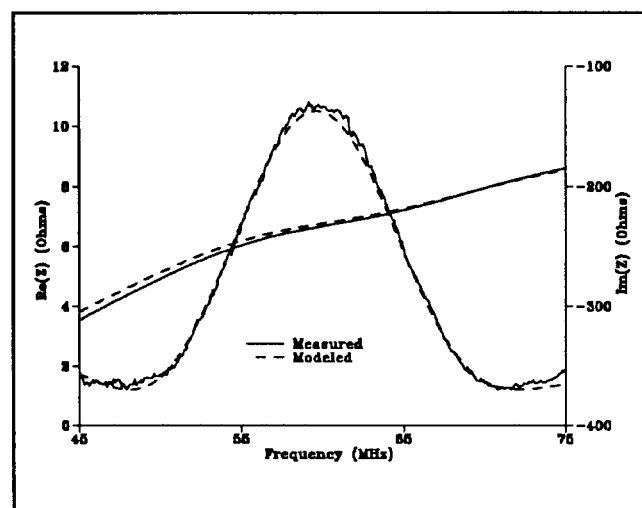


Figure 7: Measured and modeled impedance for 60 MHz Lithium Tantalate device.

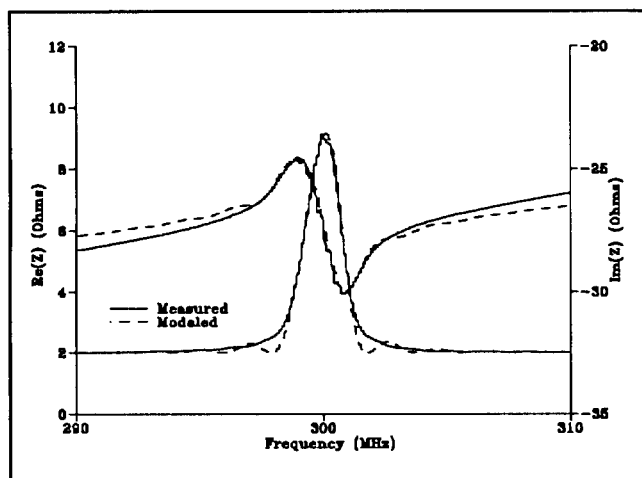


Figure 8: Performance of 300 MHz apodized transducer on ST-Quartz (measured and modeled).

$= .0011$ and $C_p = 0.522$ pF/cm, in agreement with widely used values.

APPLICATION TO UNIQUE TRANSDUCER STRUCTURES

The equivalent circuit parameter extraction technique presented above is particularly useful in model development for unconventional transducer structures, such as narrow gap or multilevel devices. In such cases, quantities such as transducer capacitance and effective coupling may not be easily calculated from previously known relationships. The parasitic thin film resistance can be found once again as the constant additive component of the real part of the input impedance. Parameter extraction can then continue as described above, provided a modified

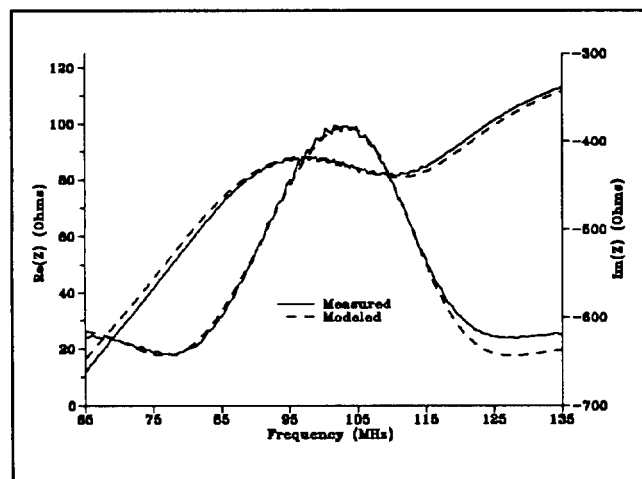


Figure 9: Input impedance of split finger multilevel transducer with Si_3N_4 layer (measured and modeled).

value of ξ is used to represent the variation of capacitance with finger separation, as described by Datta [2]. Extracted values of effective substrate capacitance can be used to calculate transducer capacitance and effective coupling coefficient can be obtained as before. These values can then be used to calculate predicted device performance and compare with experiment.

CONCLUSION

A significant amount of work will be required to quantify the dependence of these parameters on precise transducer structure, dielectric layer thickness and composition, etc. for novel transducer structures. Such a simple, versatile tool should be quite useful in the analysis of such structures, and may contribute to the investigation of less well known materials.

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