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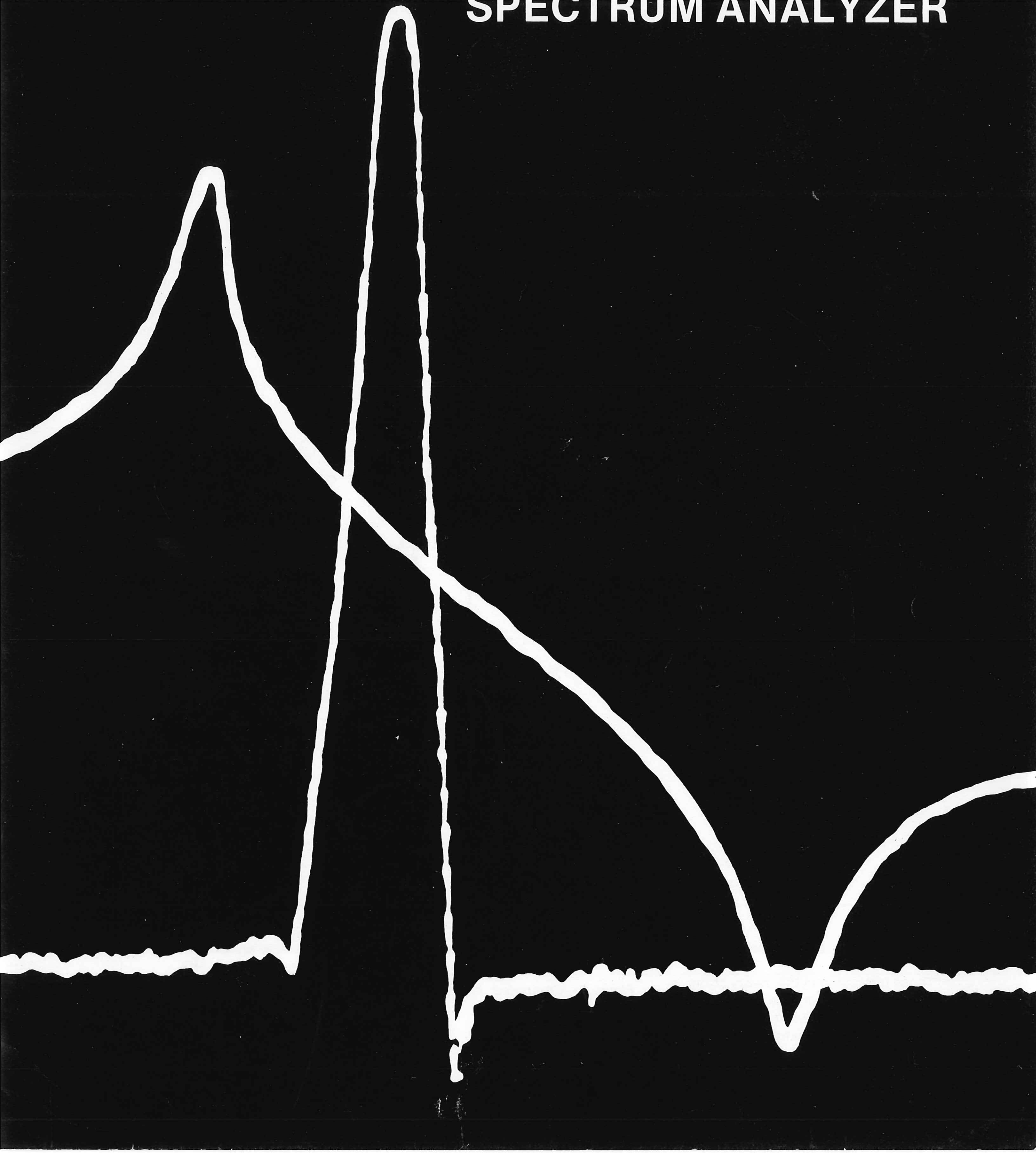
CRYSTAL DEVICE MEASUREMENTS USING THE SPECTRUM ANALYZER



CRYSTAL DEVICE MEASUREMENTS USING THE SPECTRUM ANALYZER

By Morris Engelson

Quartz crystals, crystal filters, and surface acoustic wave (SAW) devices share many common characteristics. In particular, they are all high Q and relatively narrow band devices. A high stability spectrum analyzer/tracking generator (Sa/Tg)* system is ideally suited to such narrow band device measurements. These instruments can easily characterize devices with Q's of over one million.

Accordingly, the intent of this note is to show how to make narrowband crystal device measurements using the Sa/Tg system. Let us begin with the quartz crystal.

I. The Quartz Crystal

The Equivalent Circuit

Figure 1 shows the basic equivalent circuit of a quartz crystal. The resonance conditions of this circuit are:

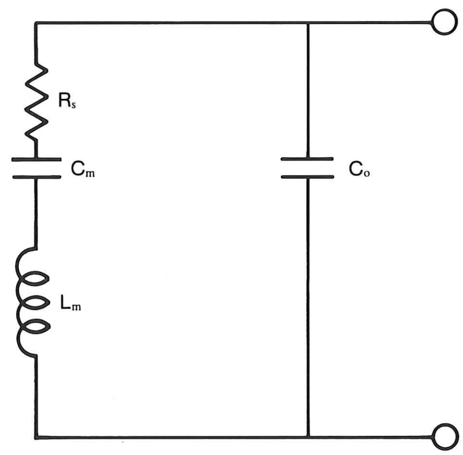
— Minimum impedance (maximum transmission) at the series resonant frequency of

$$f_s = \frac{1}{2\pi\sqrt{L_m C_m}}$$

— Maximum impedance (minimum transmission) at the parallel resonant frequency of

$$f_p = \frac{1}{2\pi\sqrt{\dfrac{L_m C_m C_o}{C_m + C_o}}}$$

A typical 5 MHz crystal resonator having a motional capacitance C_m of 0.02 pf, and a shunt capacitance C_o of 5 pf would be series resonant at 5 MHz and parallel resonant at 5.010 MHz as shown by the computer generated plot in figure 2. The Sa/Tg can be used to generate a similar graph. The test setup is shown in figure 3. This setup will accurately measure the f_s point, but the parallel resonance frequency will be in error because of stray capacitance added to C_o.



L_m = motional inductance
C_m = motional capacitance
R_s = series resistance
C_o = shunt capacitance

Figure 1—Crystal resonator, equivalent circuit.

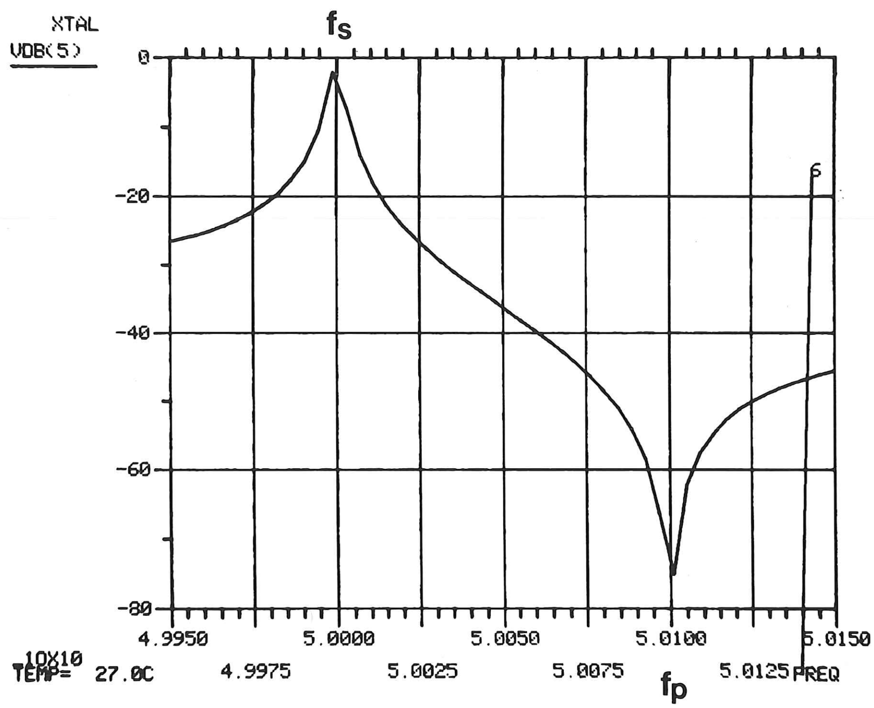


Figure 2—Computed crystal resonator transmission response.

*Those interested in the theory of operation of these instruments are referred to "The Tracking Generator/Spectrum Analyzer System," Tektronix AX-3281.

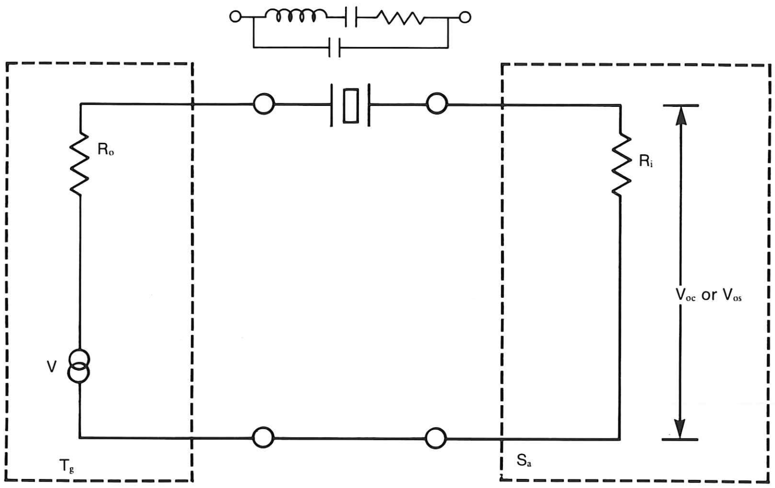


Figure 3—Sa/Tg test set up.

It is possible, with great care, to minimize external fixturing capacitance and get an accurate measurement. However, neutralizing the external circuit capacitance will produce a more accurate measurement. Figure 4 shows a simple neutralizing circuit. The variable capacitor is adjusted for minimum transmission with the crystal out of the circuit. This effectively cancels the reactive current due to stray circuit capacitance.

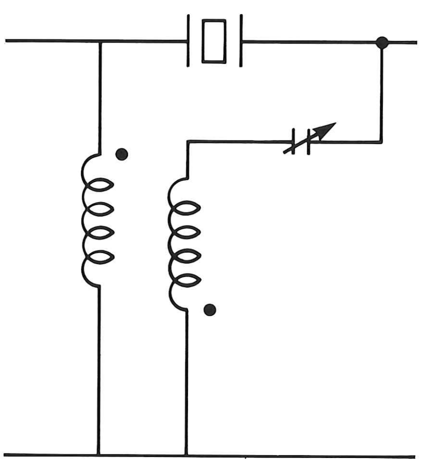


Figure 4—Capacitance neutralizing circuit.

Resonant Frequency Measurements
Figure 5 shows the transmission response of a 4.18 MHz crystal resonator as measured with a TEKTRONIX 7L5 Option 25 Sa/Tg. The series resonant frequency (f_s) aligned over the 7L5 frequency dot, is 4.18025 MHz. (The Sa/Tg displays the frequency in the upper right hand corner of the screen.) This is 250 Hz (0.006%) higher than the frequency stamped on the crystal. Determining f_s this accurately is important when matching crystal resonators in applications such as narrow band filters.

The dual trace display, in figure 5, shows two parallel resonance points. The position 5 kHz above f_s is the response in an unneutralized socket. After neutralizing stray capacitance, the true parallel resonance frequency is shown to be 7 kHz above f_s.

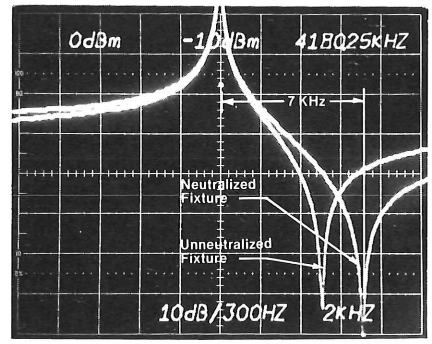


Figure 5—Transmission response of 4.18 MHz crystal resonator.

Checking R_s
One of the important parameters is the value of R_s. This can be computed from the transmission loss at the series resonant frequency, f_s. Clearly, at series resonance it is possible to replace the crystal with R_s if the reactance of C_o is large enough, or if C_o is neutralized. Thus, from figure 3, the output voltage is

$$V_{oc} = \frac{R_i}{R_i + R_o + R_s} V$$

with the crystal, and

$$V_{os} = \frac{R_i}{R_i + R_o} V$$

when a short circuit replaces the crystal. The increased transmission loss is therefore

$$\frac{V_{oc}}{V_{os}} = \frac{R_i + R_o}{R_i + R_o + R_s}$$

The smaller the values of R_i and R_o, the finer is the ability of the Sa/Tg to measure small changes in R_s. This is illustrated by figures 6 and 7.

R_i and R_o are 50 Ω each for figure 6. This is a dual trace display showing the effects of input drive level on R_s. The larger response was taken at a level of − 10 dBm while the smaller response corresponds to a signal level of − 40 dBm. The zero loss position is, in both cases, the full screen reference level. Thus, at − 10 dBm drive level the transmission loss is about 1.6 dB. This corresponds to

$$1.6\text{ dB} \rightarrow 1.2 = \frac{100 + R_s}{100};\ R_s = 20\,\Omega$$

The insertion loss goes up to about 2.2 dB at a drive level of − 40 dBm. This corresponds to an R_s of 29 Ω, a substantial increase. So, this crystal resonator is not a good choice for an application requiring a constant insertion loss with drive level.

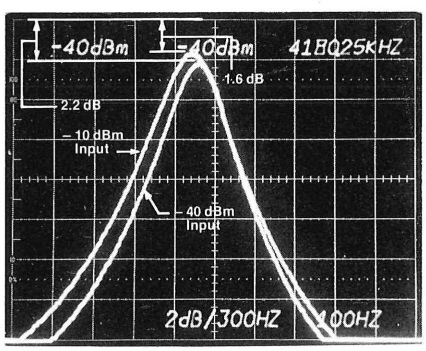


Figure 6—Measuring R_s in a 100 Ω loop ($R_i + R_o = 100\,\Omega$).

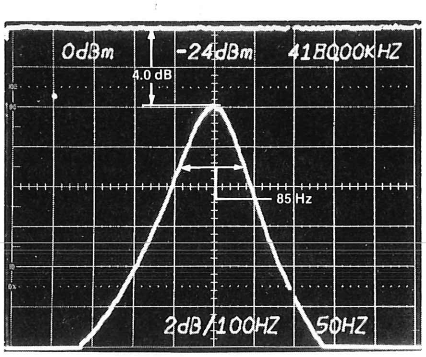


Figure 7—Measuring R_s in a 38.2 Ω loop ($R_i + R_o = 38.2\,\Omega$).

As shown by the above calculations, R_s changes rapidly for small changes in insertion loss where $R_s < (R_o + R_i)$. Reducing the value of R_i and R_o increases measurement accuracy. This is easily accomplished by use of two min-loss L pads, or with somewhat more complicated matching pi pads. Figure 7 was taken with the pi pad configuration shown in Figure 8. The 19.1 Ω of impedance nearly matches the roughly 20 Ω of the crystal, while the spectrum analyzer and tracking generator work into a nearly matched 48 Ω. Any value of crystal or measurement equipment impedance can be easily accommodated by changing the resistor values of the matching pads.

Each of the matching pads in figure 8 has a loss of 12.1 dB for a combined difference of 24.2 dB when the crystal is replaced by a short circuit. Figure 7 shows the loss of the matching pads alone as a full screen display at − 24 dBm for a 0 dBm input. Drive level to the crystal is about − 12 dBm and crystal insertion loss is 4 dB. This corresponds to

$$4\text{dB} \rightarrow 1.58 = \frac{38.2 + R_s}{38.2};\ R_s = 22.3\,\Omega$$

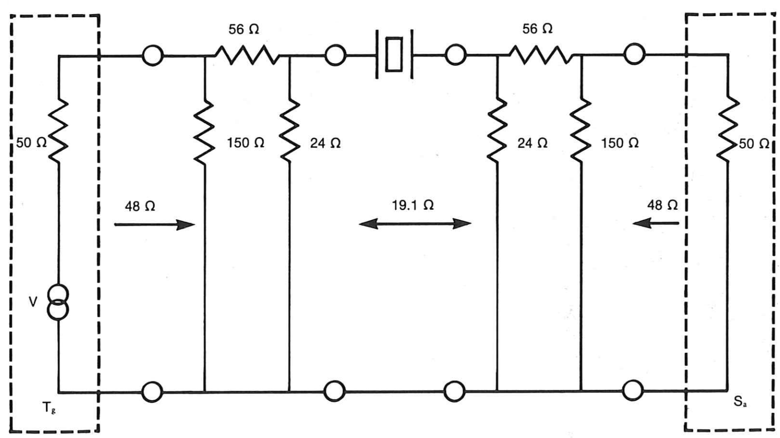


Figure 8—Matching pi pad configuration.

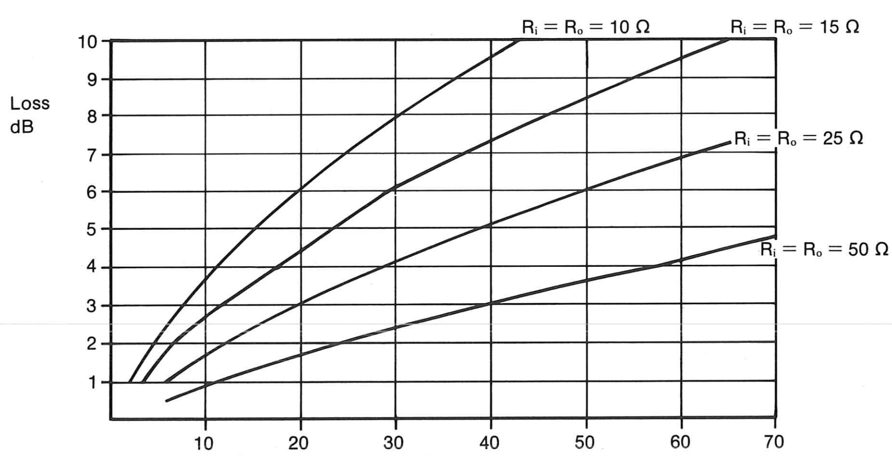


Figure 9—Crystal R_s as a function of insertion loss.

Figure 9 graphs the computed R_s as a function of the increase in insertion loss when a crystal replaces a short circuit. Various values of matching resistance are used.

Caution

A word of caution. Use of isolation pads does not necessarily remove reactances. Indeed, things can actually get worse as shown by figure 10, the crystal response in the 19.1 Ω fixture. Here the anti-resonant point has shifted from over 6 kHz to within 2.2 kHz of f_s. Any C_o calculation or filter design based on this spacing would be totally wrong. Neutralizing the fixture restores the $f_p - \text{to} - f_s$ spacing to the correct value.

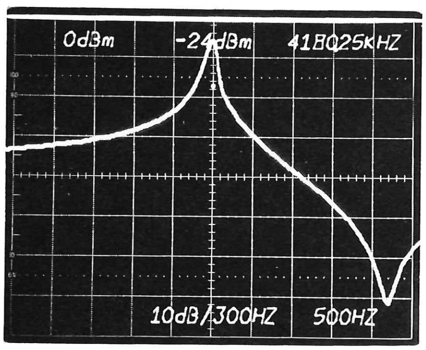


Figure 10—Transmission response showing f_s to f_p spacing in unneutralized pi pad fixture.

Determining the Reactances

The resonance relationships

$$f_s = \frac{1}{2\pi\sqrt{L_m C_m}}$$

and

$$f_p = \frac{1}{2\pi\sqrt{\dfrac{L_m C_m C_o}{C_m + C_o}}}$$

can be combined with other information for computing the value of reactive elements C_m, L_m, and C_o. Thus, the Q of a series resonant circuit is given by

$$Q = \frac{1}{2\pi f_s C_m R} = \frac{f_s}{f_3}$$

where f_3 is the 3 dB bandwidth and R is total circuit resistance including R_s, R_o and R_i.

It is important to not allow the parallel resonance at f_p to affect the measurement of Q. This means that f_p should be separated from f_s by minimizing C_o or by neutralizing C_o by means of a circuit such as that shown in figure 4. Figure 11 compares the neutralized crystal resonator response to the unneutralized response.

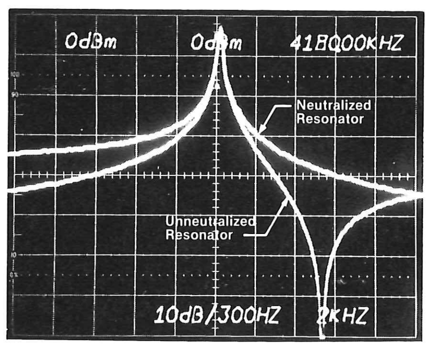


Figure 11—Transmission response, dual trace—unneutralized and with neutralized C_o.

Let us now compute C_m from a measurement of Q. Figure 7 shows an expanded series response. The vertical analyzer setting is 2 dB/div, so 3 dB down is 1.5 divisions. Careful measurement of the response width 1.5 divisions down gives a 3dB bandwidth of 85 Hz. If need be, we can get a more accurate measurement by running the analyzer in the manual sweep mode and counting the two 3 dB frequencies with an external counter. Circuit Q is

$$f_s/f_3 = \frac{4.18 \times 10^6}{85} = \frac{1}{2\pi \times 60.5 \times 4.18 \times 10^6\, C_m};$$

where $R_s = 22.3$ Ω and $R_i = R_o = 19.1$ Ω for a total $R = 60.5$ Ω.

Therefore,

$$C_m = \frac{85}{2\pi \times 60.5 \times (4.18 \times 10^6)^2} = 0.0128 \text{ pf.}$$

The motional inductance is

$$L_m = \frac{1}{\omega_s^2 C_m} = \frac{1}{(2\pi \times 4.18 \times 10^6)^2 \times 0.013 \times 10^{-12}} = 112 \text{ mH.}$$

From the resonance relationships for f_s and f_p it can be shown that:

$$\frac{f_p - f_s}{f_s} \approx \frac{C_m}{2C_o}.$$

Hence

$$C_o = \frac{C_m f_s}{2(f_p - f_s)} = \frac{0.013 \times 10^{-12} \times 4.18 \times 10^6}{2(7 \times 10^3)} = 3.9 \text{ pf.}$$

The data for $f_p - f_s$ was taken from the neutralized measurement shown in figure 5. The unneutralized display (which includes fixture capacitance) shows a $f_p - f_s$ shift of 5 kHz for a total shunt capacitance of about 5.5 pf. Of course, using a spectrum analyzer to determine C_o is something of an overkill since C_o can be easily determined by use of an appropriate L – C meter.

There is an alternative. The addition of a series capacitance (C_s) shifts the series resonance point:

$$\Delta f = f_s\left\{\sqrt{\frac{C_m + C_s + C_o}{C_s + C_o}} - 1\right\} \approx \frac{C_m f_s}{2(C_o + C_s)}.$$

We can use this relationship to compute C_m and hence L_m when C_o is known, or we can compute C_o if C_m is known. For example, with $C_m =$ 0.0128 pf a 12.8 pf capacitor shifts the series resonant point by 1450 Hz (see figure 12.) This corresponds to

$$C_o + C_s = \frac{0.0128 \times 4.18 \times 10^6}{2 \times 1.45 \times 10^3} =$$

$$18.45 \text{ pf.}$$

Hence, the unneutralized $C_o = 18.45 - 12.8 = 5.65$. That agrees with the 5.5 pf we got before.

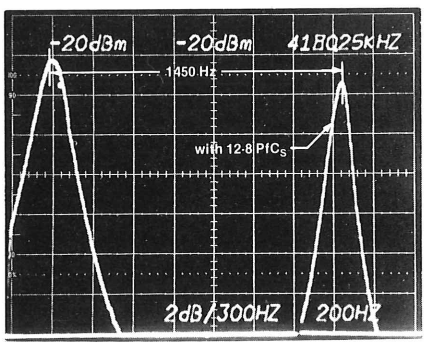


Figure 12—Shift of crystal resonator f_s with the addition of 12.8 pf in series.

Overtone Crystals

We may be in for some surprises if we use high overtone crystal resonators that are insufficiently characterized. Consider the 100 MHz, 5th-overtone crystal whose transmission response is shown in figure 13. We can easily determine the crystal series resonant frequency by using the 7L13/TR502 Sa/Tg in the dot counter mode with a DC502 Opt 7 counter measuring a frequency of 100.013 MHz at the bright dot position on the display.

We can determine insertion loss, R_s, C_m and C_o for the overtone crystals the way we did before. One point of concern is the numerous spurious resonances in the vicinity of the main resonance. These should be characterized in terms of R_s and frequency to avoid spurious modes in oscillator design or spurious passbands in filters. Indeed, this resonator has numerous responses, including a major response at 20 MHz (figure 14), and a less prominent one at 220 MHz (figure 15).

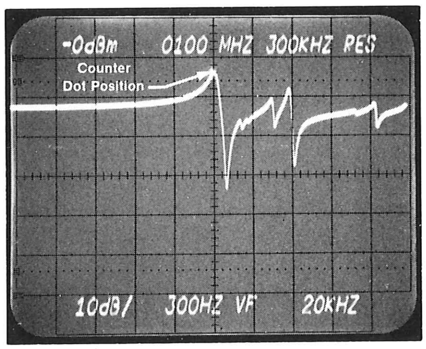


Figure 13—Transmission response of overtone crystal resonator.

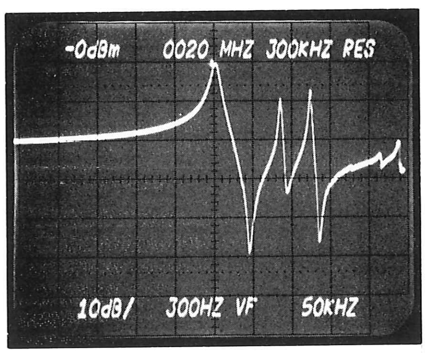


Figure 14—20 MHz response of fifth overtone 100 MHz crystal resonator.

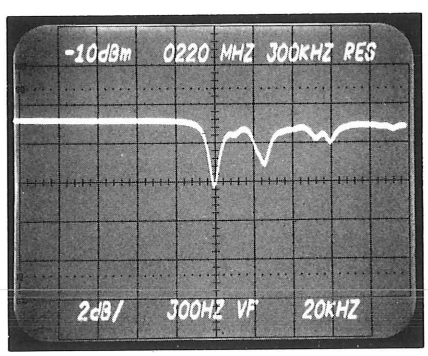


Figure 15—220 MHz response of 100 MHz overtone crystal resonator.

II. Crystal Filters

Crystal filters are usually narrowband devices constructed from crystal resonators. In fact, one of the reasons for accurately characterizing crystal resonators is to provide data for filter design. Some basic filter measurements are insertion loss, bandwidth, frequency, and ultimate attenuation. These are illustrated in figures 16 and 17.

Figure 16 shows an insertion loss measurement. The upper horizontal line shows the tracking generator at 0 dBm output level as displayed on a 7L5 Option 25 Sa/Tg with 0 dBm reference level setting. Filter insertion loss is 3 dB. Filter center frequency is 3.5785 MHz as indicated in the upper right hand crt readout position.

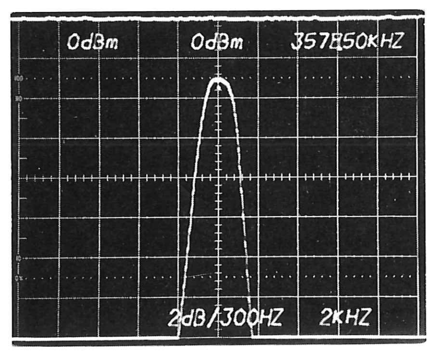


Figure 16—Filter response measurement, fine detail at 2dB/div.

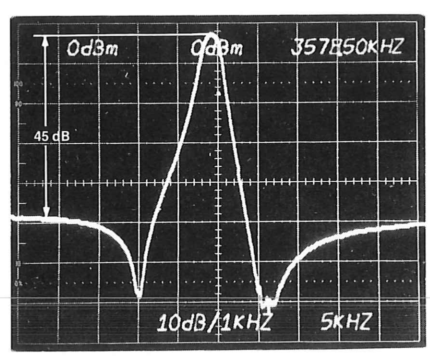


Figure 17—Filter response measurement, at 10dB/div.

Filter bandwidth can likewise be determined by measuring the frequency width at the desired response positions. Here, the 3 dB width is one horizontal division, or 2kHz. Figure 17 shows the response of the same filter in the 10 dB/div. display mode. Close to the passband, out-of-band rejection is 65 dB, degrading to 45 dB at more than 10 kHz from the passband. Ultimate attenuation of this filter is, therefore, 45 dB.

One way of improving the ultimate attenuation of a filter is to cascade several sections. This is illustrated in the composite of two photos, figure 18. This filter shows an ultimate attenuation of over 100 dB. Insertion loss is 8 dB, and bandwidth at the 100 dB down points is 8 kHz for a 100 dB/3 dB shape factor of only four to one.

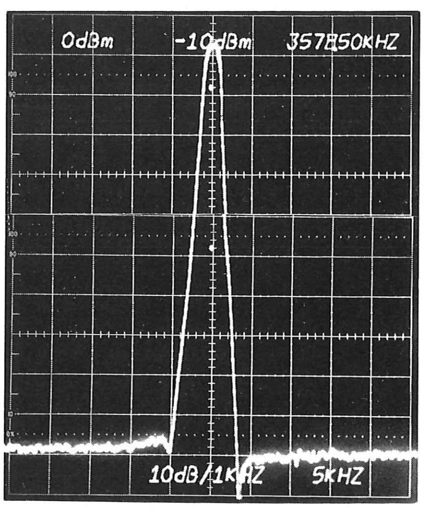


Figure 18—High ultimate attenuation measurement, composite of two photos.

Return Loss and Intermodulation
Shape factor, loss and ultimate attenuation are just a few of the measurements of interest. Figure 19 shows a return loss measurement using a directional bridge. The return loss at bandpass center is over 20 dB for a swr of less than 1.3:1.

Another measurement is that of spurious responses (figure 20). Here the single section filter shows two large spurious responses of 80 kHz and 120 kHz above the main response.

Note that the spurious responses are eliminated in the more elaborate design whose transmission characteristic is illustrated in figure 21. A useful capability of the 7L5 Opt. 25 is storing one display in memory while updating a second one. This permits storing a desired filter shape in memory so that other filters may be tuned to match the sample (figure 22).

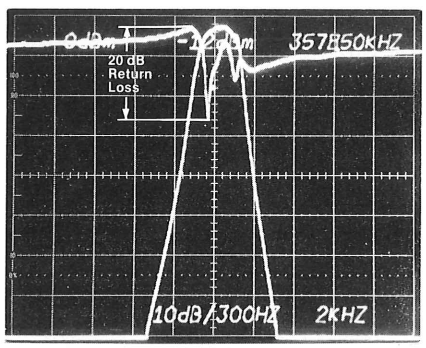


Figure 19—Filter return loss measurement.

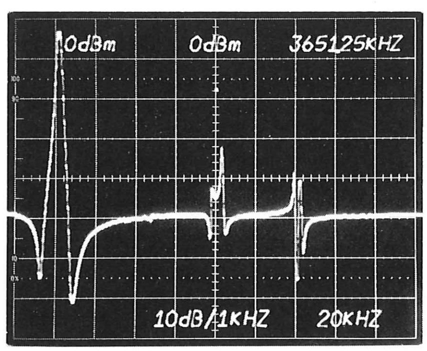


Figure 20—Spurious responses of simple filter.

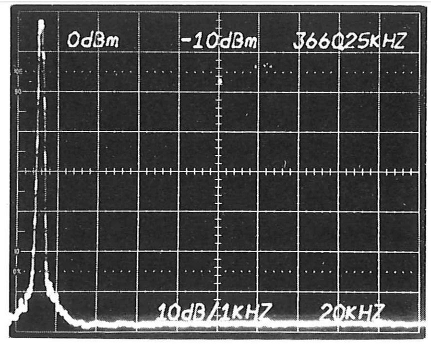


Figure 21—Multi-section filter eliminates spurious responses.

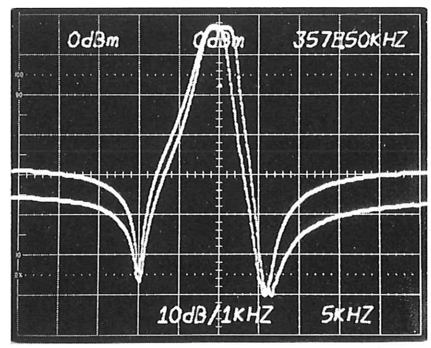


Figure 22—Tuning one filter to match the stored response of another.

It is becoming more important to determine crystal filter intermodulation. Crystal resonators may not be perfectly linear, as we saw in figure 6. This nonlinearity causes filters to produce intermodulation. Such a measurement is illustrated in figures 23 and 24.

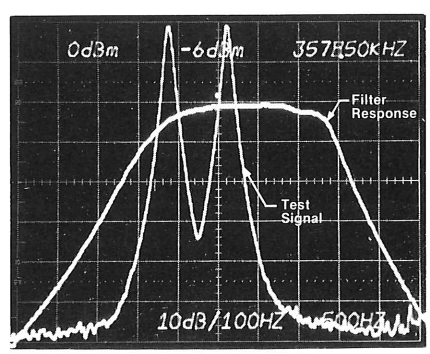


Figure 23—Filter transmission response and two tone test signal input.

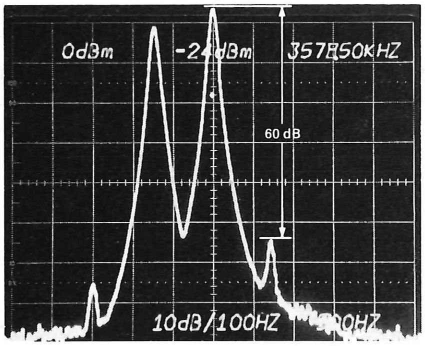


Figure 24—Filter response of two tone test signal showing intermodulation components.

Figure 23 shows the filter passband response and the two tone test signal superimposed on it. Note that we only have two tones (third-order products are below the noise level).

Figure 24 shows the output of the filter. The upper signal which is centered in the filter passband is larger than the lower frequency signal which is down on the filter skirt. In addition, we now have two third-order intermodulation products on either side of the two tone test signal. The left intermodulation product is on the filter skirt and the right product is within the passband. Thus, with one test signal and one intermodulation product within the passband we can directly measure the ratio of unwanted to wanted components. Figure 24 shows that the third order response is 60 dB below the desired signal. The test signal spacing with respect to the filter response is not accidental. At least one of the test signals (some filters need both test signals) and one undesired product must be within the filter passband to get a true indication of filter intermodulation.

SAW Devices
Surface Acoustic Wave Devices have stabilities comparable to crystal resonator filters, but operate at much higher frequencies. Let's characterize a low frequency version of such a device.

Figure 25 shows the response of a SAW filter as measured with a 7L13/TR502 Sa/Tg unit. Center frequency is 71 MHz, bandwidth is 5 MHz at the −3 dB level, and 7.5 MHz at the −60 dB level. Shape factor 60/3 dB is a sharp 1.5 to 1. Ultimate attenuation is 60 dB. The tracking generator drive level is 0 dBm showing an insertion loss of 20 dB for a −20 dBm output level. The relatively high insertion loss and the ripple response in the upper stop band are typical of such a filter.

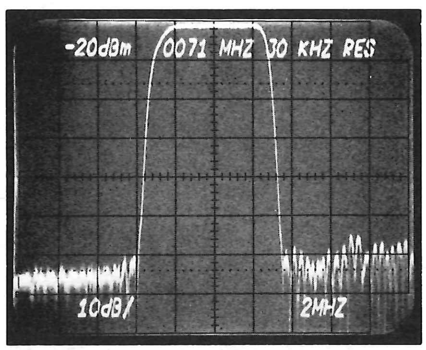


Figure 25—Transmission response of SAW filter.

Input impedance characteristics are defined by the return loss display, figure 26. Return loss is less than 4 dB for a swr of about 4 to 1.

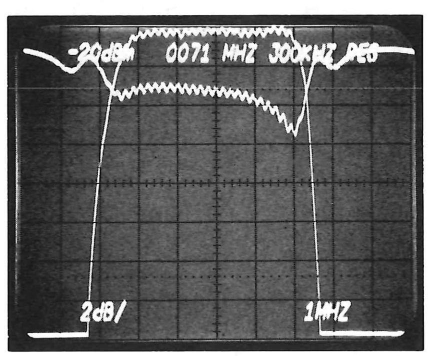


Figure 26—SAW filter return loss measurement.

High frequency characteristics of the filter show numerous extraneous passbands at multiples of the primary one. Figure 27 shows a response at three times 71 MHz. The insertion loss has increased by another 10 dB, but otherwise it is almost a usable filter.

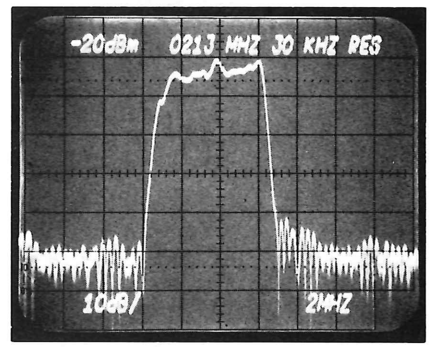


Figure 27—SAW filter transmission response at three times design frequency.

III. Summary
Spectrum analyzer/tracking generator systems like the TEKTRONIX 7L13/TR 502 (which has less than 10 Hz fm'ing up to 1800 MHz) and the TEKTRONIX 7L5 Option 25 (which has less than 2 Hz fm'ing up to 5 MHz) provide a powerful tool in characterizing a wide variety of narrowband, high Q devices.

Formulas

$$f_s = \frac{1}{2\pi\sqrt{L_m C_m}} \; ; \; f_p = \frac{1}{2\pi\sqrt{\dfrac{L_m C_m C_o}{C_m + C_o}}}$$

$$L_m = \frac{1}{(2\pi f_s)^2 C_m} \; ; \; Q = \frac{f_s}{f_3} = \frac{1}{2\pi f_s C_m R}$$

$$\frac{f_p - f_s}{f_s} = \sqrt{1 + \frac{C_m}{C_o}} - 1 \approx \frac{C_m}{2C_o}$$

$$\frac{\Delta f}{f_s} = \sqrt{1 + \frac{C_m}{C_s + C_o}} - 1 \approx \frac{C_m}{2(C_o + C_s)}$$

$$\text{Crystal insertion loss} = \frac{R_i + R_o}{R_i + R_o + R_s}$$

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