Chapter V

Development of Surface Acoustic Wave Sensors

Basic information related to surface acoustic wave is given. Difference between SAW delay lines and SAW resonators is discussed. The working mechanism of surface acoustic wave sensors is discussed. Difference between characterising SAW devices with network analyzer and oscillator is discussed. Design aspects for SAW based oscillators in three different topologies are discussed. Effect of semiconducting ZnO thin film deposited on the top of SAW delay lines and resonators is discussed. Effect of acoustoelectric interaction between UV induced charge carries in ZnO thin film and SAW propagation characteristics is discussed. Undesirable huge frequency shifts due to mode hopping in delay line based SAW oscillator is highlighted. Design of a hand held prototype for SAW sensor with measurement electronics is discussed.
5.1 Introduction

5.1.1 Surface Acoustic Wave Devices

In a solid material, an acoustic or elastic wave involves *strains* arising due to change in relative positions of atoms, and *stresses* arising out of internal forces tending to return the material to its equilibrium, unstrained state. Propagation of Acoustic waves involves *stresses* and *strains* [Morgan 2007] The existence of surface acoustic wave (SAW) was shown in 1885 by Lord Rayleigh [Rayleigh 1885]. The wave, often called a Rayleigh wave, propagates along the plane surface of an isotropic solid half-space, with amplitude decaying exponentially away from the surface. With the advent of the interdigital transducer (IDT) in 1965 [White and Volmer 1965], the surface acoustic wave device became a suitable shaped metallic thin film deposited on the surface of a piezoelectric crystal such as quartz or lithium niobate. Quartz is the most commonly used substrate because of its temperature stability for certain crystal orientations.

The transducer consists of a sequence of metal electrodes, usually of aluminium, alternately connected to bus bars [Figure 5.1]. Two IDTs are required in the basic surface acoustic wave device (SAW) configuration. A periodic electric field is produced when an *rf* source is connected to the input electrode, thus permitting piezoelectric coupling to a travelling surface wave. Direct coupling to a surface elastic wave is possible at the boundary of a piezoelectric solid if any of the components of strain at the surface is piezoelectrically active [R. M. White and F. W. Volmer (1965)]. To ensure constructive interference and in-phase stress, the distance ‘d’ between two neighbouring fingers (IDTs) is equal to half the elastic wavelength *λ*$_{R}$. The associated frequency is known as synchronous frequency. At this frequency, the transducer efficiency in converting electrical energy to acoustical is maximised. The output IDT works in a reciprocal manner, and converts mechanical SAW vibrations back into output alternating voltages. The surface outside the IDT regions need only be elastic, without being piezoelectric. The electrode width is a quarter of the center-frequency wavelength, and the minimum possible width, determined by the fabrication technology, limits the achievable frequency [Morgan 2007]. The type of acoustic wave and the wave velocity generated in a piezoelectric material depends on the substrate material properties, the crystal cut, and the structure of the electrodes used to transform the electrical energy into mechanical energy. The Rayleigh wave
generated on the piezoelectric substrate has both a surface normal component and a
surface parallel component with respect to the direction of propagation. The surface
particles move in elliptical paths having a surface normal and a surface parallel
component. The energy of the surface acoustic wave (SAW) is confined to a zone
close to the surface and is of a few wavelengths thick. Two of the most important
physical properties relating to surface wave propagation on a piezoelectric substrate
are the wave velocity ($v$) and the electromechanical coupling coefficient $K^2$ of the
piezoelectric substrate.

5.1.2 Types of Surface Acoustic Wave (SAW) Devices

5.1.2.1 Delay Line Device

The simple surface acoustic wave device is a non-dispersive delay line. Source IDT
sets up an electric field on the surface of the substrate and launches a surface acoustic
wave by means of the piezoelectric effect, and the receiving transducer converts the
acoustic wave back to an electrical signal. By reciprocity, only half of the intercepted
acoustic energy at the output is reconverted to electrical energy; hence, the inherent 6
dB loss is associated with this structure. This basic SAW device is also regarded as a
bandpass filter since the response is strongest at a particular frequency when the
surface wave wavelength is similar to the transducer pitch. The absorbing material is
applied at each end of the device to eliminate unwanted reflections from the ends of
the substrate. Numerous second-order effects, such as coupling efficiency, resistive
losses, and impedance mismatch, raise the insertion loss of practical filters to 15 - 30
dB.

Since the acoustic velocity of the surface wave is much slower than the speed of light,
an acoustic wavelength is much smaller than its electromagnetic counterpart. This
results in the SAW’s unique ability to incorporate an incredible amount of signal
processing or delay in a very small volume.

Commercially available SAW devices work in the range from 20 MHz to 5 GHz.
Below 20 MHz, the device size becomes large and inconvenient, whereas high
frequency devices above 5 GHz are difficult to fabricate with good yield due to
lithographic limitations. Throughout the world, more than 100 million SAW filters for
TV IF filtering are fabricated every year. Because of their superior performance, their
high reproducibility and small size, they have totally replaced LC filters.
Fig. 5.1: Basic Surface Acoustic Wave (SAW) Device

The principal design tool of a SAW filter is the Fourier transform, which is used to relate the time and frequency responses of the transducers and resultant filter. In general, the designer derives two impulse responses for the two transducers whose transforms can be added together (in dB) to produce the desired total frequency response characteristics. These two impulse responses are then etched onto the surface of a metalized piezoelectric substrate.

To design a SAW device or filter with a given resonant frequency $f_o$ and fractional bandwidth $B$ (measured null to null on either side of the resonant frequency), we have to take into account the following specifications:

a) Acoustic wavelength

$$\lambda = \frac{v_o}{f_o}$$

The width of each finger that results in this synchronous frequency is $\lambda/4$ and the interdigital spacing measured from center-to-center is $\lambda/2$.

b) Number of finger pairs needed to achieve fractional bandwidth specification ‘B’ is

$$N_p = \frac{2}{B}$$

c) Impedance of IDT (Z) should be matched with the impedance of the measurement system (typically 50 ohm). The IDT behaves as a capacitive system determined by the
number of finger pairs, their spacing, as well as the degree of overlap. Total capacitance required is:

$$C_t = \frac{1}{2\pi f_o Z}$$

d) Acoustic aperture (overlap between fingers, in length units) is

$$W = \frac{C_t}{C_o N_p}$$

where $C_o$ is capacitance per finger pair per unit length.

e) Frequency magnitude response of IDT approximates as incoherent addition of contributions from individual fingers:

$$\phi_1(f) = \sin \left( \frac{X}{X} \right)$$

where

$$X = \frac{N_p \pi (f - f_o)}{f_o}$$

f) Cumulative frequency magnitude response for the SAW is simply the dot product:

$$S_{21}(f) = \phi_1(f) \cdot \phi_2(f)$$

where $\phi_2(f)$ is frequency response from output IDT

g) Other important parameter in SAW device design is the metallization ratio, $\eta = a/b$ where $a$ is the metal width of an IDT finger and $b$, the combined metal cum finger spacing. The normalized thickness $h/\lambda$, where $h$ is the IDT metallization thickness, also needs to be taken into account. Film thickness ratios are normally quoted for aluminum metallization.

5.1.2.2 Resonators

The role of SAW resonators emerged when it was realized that they could provide the low insertion losses being demanded for bandpass filtering in mobile telephones. Surface waves can be reflected strongly by a reflector grating consisting of regular array of strips of shorted metal electrodes. Although any surface perturbation, such as deposited strip will reflect surface waves, it will in general produce bulk waves as well, leading to unexpected losses. At the Bragg frequency, the periodicity equals half the wavelength, and reflections from individual strips have the same phase so they
add coherently. At other frequencies, the waves get reflected out-of-phase and consequently die out resulting in a highly frequency selective resonator. In a grating each strip reflects weakly, but if there are many strips the total reflection coefficient can be close to unity at the center frequency. A SAW reflection grating may employ up to several hundred reflecting strips to achieve near total SAW reflection [Campbell 1998]. Surface acoustic wave resonator structures are realized by employing either one or two IDTs bounded by such reflection gratings. One port SAW resonators [Figure 5.2] employ a single IDT for input and output, in conjunction with two reflection gratings, while two port SAW resonators [Figure 5.3] are formed using separate IDTs for input and output signals, which are contained between SAW reflection gratings. One-port and two-port SAW resonators have become popular due to their use in oscillatory circuitry, for low power unlicensed wireless applications (e.g. remote control security, automobile keyless entry etc.) typically transmitting less than 1mW of power, and operating over distance of from 5 to 500 metres [Campbell 1998].

![Fig. 5.2: One Port Surface Acoustic Wave (SAW) Resonator.](image)

In a one-port SAW resonator, surface waves are emitted from both sides of the excited IDT and they are constructively being reflected by the SAW reflection gratings at the center frequency. This gives rise to SAW standing waves within the IDT. As the SAW gratings reflect surface waves emanating from both sides of an excited bidirectional IDT, the device insertion loss can be less than 6 dB.
All the acoustic resonators operate at the resonant frequency. Resonance occurs when the input impedance is at a minimum and anti-resonance occurs when impedance is at a maximum. The resonant frequency and the anti-resonant frequency are referred to as the *series* frequency and the *parallel* frequency respectively. A *series* resonant circuit allows a maximum current flow at resonant frequency, whereas a *parallel* resonant circuit allows a minimum. At these frequencies, the response of SAW resonators is completely real and does not have any imaginary component.

The lumped-element equivalent circuit of one port SAW resonator, depicted in figure 5.4, consists of a series inductance-capacitance-resistance (LCR) branch shunted by a capacitor. Shunt capacitance $C_T$ represents the IDT capacitance, while the elements $L_r$, $C_r$, and $R_r$ relate to equivalent motional parameters for the series resonance condition.

The capacitance ratio $C_T/C_r$ is an important parameter in one-port SAW resonator, as it is often desirable that the parallel and series resonance frequency be offset by a specific amount for optimum performance. One-port SAW resonator has a large static-to-motional capacitance ratio $C_T/C_r$, with small frequency separation between the series- and parallel-resonance modes. This property of one-port resonator limits it to fixed frequency designs for oscillator operation.

If IDT finger reflections are neglected, the equivalent series resistance $R_r$ may be approximated by

$$R_r \approx \frac{1}{G_a(f_o) (1+|\rho|)} \text{ ohms}$$

Where $G_a(f_o)=8K^2 f_o C_s N^2 \rho_r$ = unperturbed radiation conductance at IDT center frequency $f_o$. Parameter $\rho$ ( $|\rho| < 1$) is a dimensionless reflection coefficient, relating the ratio of reflected-to-incident surface waves entering the reflection grating. $K^2$ is the electromechanical coupling coefficient.

The series inductance $L_r$ may be approximated by

$$L_r \approx \frac{d_e}{\lambda_o} \left( \frac{1}{4f_o G_a(f_o)} \right)$$

where $d_e$ is the effective resonant cavity length.
The equivalent series capacitance $C_r$ is obtained as

$$C_r = \frac{1}{4\pi^2 f_0^2 L_r}$$

For high $Q$-resonators, the frequency separation between parallel and series resonant frequencies is

$$f_p - f_s = \frac{f_0 C_r}{2 C_T}$$

For grating reflection coefficient magnitudes $|\rho|$ close to unity, the resonant cavity $Q$ is approximated by

$$Q \approx \frac{d_e}{\lambda_c} \frac{2\pi}{1 - |\rho|^2}$$

**Fig. 5.3:** *Two Port Surface Acoustic Wave (SAW) Resonator.*

The two port SAW resonator have equal number of fingers in both the input and output IDTs. The voltage transfer-function response between input and output IDTs may be considered to be a composite of two contributions. In the absence of reflection gratings, the response would be similar to that of a SAW filter with uniform, and equal, input and output IDTs, giving a $(\sin X) / X^2$ amplitude distribution. With the inclusion of reflection gratings, the resonator response around center frequency will be superimposed.
The operation of SAW resonators is crucially dependent on the separation between IDTs and adjacent reflection gratings, as this controls the standing wave pattern and optimum performance of the resonant structure.

Fig. 5.4: Lumped equivalent circuit model of (a) SAW one-port, and (b) SAW two-port resonator.

5.1.3 SAW Device Characterization

The amplitude and phase characteristics of Surface Acoustic wave devices/sensors are measured using a two port Network Analyzer. The network analyzer functions as a combined source and vector voltmeter, measuring the amplitude and phase as a function of input frequency. The network analyzer consists of a synthesized sweeper (0.3MHz – 1.3 GHz in case of Agilent 8712ES), test setup and a digital processor. The sweeper provides the stimulus and the test setup provides signal separation. The display panel of network analyzer is used to define and conduct various measurements. The device under test (DUT) is connected between PORT 1 and PORT 2. The port at which device is connected to the test setup is called the reference plane. All measurements are made with respect to this reference plane. Figure 5.5 (a) and (b) shows the experimental set for characterizing SAW devices in frequency domain. While the network analyzer has utility in measuring the gain phase response of SAW devices, the spectrum analyzer and frequency counter helps in the design and characterization of SAW oscillators. Special shielded connectors are required for rf measurements as shown in figure 5.5(b).
Scattering Parameters and the Smith Chart

Scattering parameters (S-parameters) have earned a prominent position in Radio Frequency circuit design, analysis and measurement. S-parameters are defined and measured with ports terminated in reference impedance [Rhea 1995].

Two port S-parameters are defined by considering a set of voltage waves. When a voltage wave from a source is incident on a network, a portion of the voltage wave is transmitted through the network, and a portion is reflected back towards the source. Incident and reflected voltage waves may also be present at the output of the network. Power waves are extracted by dividing the square of the magnitude of these voltage waves with the reference impedance.

\[
|a_1|^2 = \text{incident power wave at the network input} \\
|b_1|^2 = \text{reflected power wave at the network input} \\
|a_2|^2 = \text{incident power wave at the network output} \\
|b_2|^2 = \text{reflected power wave at the network output}
\]

S-parameters are related to these variables by the following expressions:

\[
b_1 = a_1S_{11} + a_2S_{12} \\
b_2 = a_1S_{21} + a_2S_{22}
\]

Terminating the network with a load equal to the reference impedance forces \(a_2=0\)

\[
S_{11} = \frac{b_1}{a_1} \quad \text{and} \quad S_{21} = \frac{b_2}{a_1}
\]
S_{11} is defined as the input reflection coefficient and S_{21} is the forward voltage transmission coefficient of the network.

Terminating the network at the input with a load equal to the reference impedance and driving the network from the output port forces \( a_1 = 0 \)

\[
S_{22} = \frac{b_2}{a_2} \quad \text{and} \quad S_{12} = \frac{b_1}{a_2}
\]

S_{22} is defined as the output reflection coefficient and S_{12} is the reverse transmission coefficient of the network.

The S-parameters expressed in decibel units are:

\[
|S_{11}| = \text{input reflection gain (dB)} = 20 \log |S_{11}|
\]

\[
|S_{22}| = \text{output reflection gain (dB)} = 20 \log |S_{11}|
\]

\[
|S_{21}| = \text{forward gain (dB)} = 20 \log |S_{21}|
\]

\[
|S_{12}| = \text{reverse gain (dB)} = 20 \log |S_{12}|
\]

S_{11} and S_{22} are referred to as return losses. The complex input impedance is related to the input reflection by the expression.

\[
Z_{\text{input}} = Z_0 \frac{1 + S_{11}}{1 - S_{11}}
\]

Input VSWR and S_{11} are related by VSWR = \( \frac{1 + |S_{11}|}{1 - |S_{11}|} \)

where VSWR is the Voltage Standing Wave Ratio.

Measurement of scattering parameters (S-parameters: S_{11}, S_{21}, S_{12}, S_{22}) allows determination of both phase velocity and attenuation of the SAW device over a given frequency domain. Useful information such as device admittance, impedance and transmission loss can be yielded at any chosen frequency.

The impedance Smith chart is an insightful display overlay for mapping the impedance plane and the reflection coefficient. The polar form of a reflection coefficient plotted on a Smith chart provides the corresponding impedance. All values on the Smith chart are normalized to the reference impedance such as 50 ohms. The radius of the standard Smith Chart is unity. The magnitude of the reflection
coefficient is plotted as the distance from the center of the Smith chart. A perfect match plotted on a Smith chart is a vector of zero length (the reflection coefficient is zero) and is therefore located at the center of the chart which is 1+j0, or 50 ohms.

Purely resistive impedances map to the only straight line of the chart with zero ohms on the left and infinite resistance on the right. Pure reactance is on the circumference. Arcs rising upwards are constant normalized inductive reactance and descending arcs are constant normalized capacitive reactance. High impedances are located on the right portion of the chart, low impedances on the left portion, inductive reactance in the upper half, and capacitive reactance in the lower half. The angle of the reflection coefficient is measured with respect to the real axis, with zero degrees to the right of the center, 90° straight up, and -90° straight down.

Figure 5.6 shows the impedance Smith chart for an input side of a SAW delay line filter operating at 36 MHz and a two port resonator operating at 434 MHz.

![Impedance Smith Chart](image)

**Fig. 5.6:** Impedance Smith Chart in $S_{11}$ mode for (a) 36 MHz SAW delay line, and (b) 434 MHz two port resonator.

The smith chart obtained for $S_{11}$ parameter from the network analyzer is useful for measuring the capacitance ($C_T$) and the acoustic conductance ($G_a$) of the IDT at the resonant frequency $\omega_o$. The electromechanical coupling factor can also be evaluated from the smith chart using the following relation [Parmanand 2002].

$$K^2 = \frac{\pi G_a}{4(N - 1/2)\omega_o C_T}$$
5.1.4 Surface Acoustic Wave Sensors

The principal means of detection of a change of physical property follows from the transduction mechanism of a SAW device, i.e. involving the conversion of signals from the physical (acoustic wave) domain to the electrical domain. Acoustic wave sensors rely on the change in the acoustic velocity of a bulk or a surface wave within or on the surface of the piezoelectric substrate. Measurement of attenuation of acoustic wave shows crystal amplitude damping effects due to perturbation of the electric field above the propagation path length.

Small perturbations affecting the acoustic wave manifest themselves as large scale changes when converted to the electromagnetic domain because of the enormous difference in their velocities.

Acoustic wave sensors are inherently rugged, very sensitive, intrinsically reliable and competitively priced. Many of these sensors are also capable of being passively and wirelessly interrogated, i.e. they can work without any power source [Drafts 2001]. The applications of acoustic wave devices as sensors include industrial sensors (vapour, humidity, temperature, mass sensors), automotive sensors (torque and tire pressure sensors), and recently in medical biosensors. If the cut of the piezoelectric substrate material is rotated appropriately, the wave propagation mode changes from a vertical shear SAW sensor to a SH-SAW sensor. This is advantageous in situations when there is a need to work the sensor in a liquid medium, allowing the SH-SAW sensor to operate as a biosensor.

The spectrum of physical and chemical phenomena that can be sensed by SAW devices is achieved by coating the devices with materials that undergo sheet conductivity changes, mass loading, visco-elastic perturbations. By proper selection of a coating that absorbs only specific chemical vapors, a vapour sensor can be materialised. If a coating is applied that absorbs specific biological chemicals in liquid medium, a biosensor can be realised. A photoconducting thin film coating on top of a SAW device can also be utilised to verify the photodetector properties of SAW sensor. An interesting property of quartz substrate is that it is possible to select the temperature dependence of the material by the cut angle and the wave propagation direction. With proper selection, the first-order temperature effect can be minimized. An acoustic wave temperature sensor may be designed by maximizing this effect.
Temperature sensors based on SAW delay line oscillators have millidegree resolution and good linearity.

When an IDT sensor is directly connected to an antenna, electromagnetic (em) waves received by wireless transmission can execute surface acoustic waves in the piezoelectric material. This property can be used to functionalize passive and wireless (remotely operable) SAW sensors. These wireless sensors are an attractive proposition when low power sensors are needed and even more attractive for use in remote, inaccessible locations like sensors buried in concrete/ground or fixed in automobile tyres.

The important properties of piezoelectric transducers that justify their use for sensing applications are:

1. There is an ideal coupling mechanism between the electric circuit and the mechanical properties of the crystal, ensuring that the frequency of the mechanical acoustic wave is identically equal to the electrical frequency, i.e., a distortion free interface having extremely low dissipation.
2. The anisotropic properties of piezoelectric crystals allows for different angles of cut with respect to the crystallographic axis, which, therefore allows the use of crystals with a wide range of frequencies.
3. The model of the acoustic wave can be changed from compressional to shear, or the angle of propagation can be varied.

Change in acoustic velocity of a SAW microsensor is a combination of many material parameters. Several different properties of film coatings that can affect the acoustic waves are – mass, density, conductivity, electrical permittivity, strain and viscoelasticity. Change in acoustic velocity can be reflected by the total differential theorem to the change in any or all of the properties.

\[
\frac{\Delta v}{v} \approx \frac{1}{v} \left( \frac{\delta v}{\delta \text{mass}} \Delta \text{mass} + \frac{\delta v}{\delta \text{elec}} \Delta \text{elec} + \frac{\delta v}{\delta \text{mech}} \Delta \text{mech} + \frac{\delta v}{\delta \text{env}} \Delta \text{env} + \cdots \right)
\]

Care must be taken in the choice of IDT design and signal processing techniques so that only changes in the desired parameter are measured and not the cross interfering signals e.g. mechanical strain or environmental temperature. The PCB on which the SAW sensor is mounted is also significant and care must be taken to avoid any strain on PCB since such strain could lead to relaxation of PCB and consequently frequency drift of SAW oscillator (Nimal et al. 2009).
To functionalize a SAW sensor, the SAW device (delay line filter or resonator) is used as the frequency control element in a closed loop oscillator circuit.

Analyzing a Surface acoustic wave device in oscillator mode is more advantageous than compared to characterization in a Network analyzer are:

1. The measured frequency resolution of an oscillator is limited by the accuracy of the frequency counter, which is much higher than that of any network analyser. The minimum resolution possible in the sensing of a dynamic signal is limited by the minimum time taken by the network analyzer to sweep across a range of frequencies. The network analyzer has to be used in conjunction with real time dynamic data acquisition (DAQ) system like Labview to obtain dynamic data.

2. The network analyzer is far too sophisticated for general purpose routine laboratory use and required highly qualified personnel for its operation. The oscillator technique, on the contrary, is less expensive than network analyzer. Microcontroller based Frequency Counter can be integrated with the SAW Oscillator to develop a miniaturized prototype for SAW based sensing applications with an estimated cost of less than $50 as done in the present study.

3. Wireless SAW sensors are possible, and are applicable for remote monitoring of any physical or chemical property, where the rf signal from the wireless sensors is retrieved using a spectrum analyzer.

4. Only one acoustic sensor can be monitored at a time using a network analyzer. Whereas multiple SAW oscillators based sensors can be integrated on a single platform to develop an electronic nose along with principal component analysis.

5. The measurement of frequency has the greatest dynamic range $10^7:1$, whereas amplitude and phase measurements have a much reduced dynamic range between 10000:1 and 1000:1.

5.1.4.1 Acoustoelectric Interaction Mechanism

The perturbation of the electrical boundary condition based on the acoustoelectric interaction in a semiconducting/piezoelectric multilayer structure is a promising approach to realize practical SAW sensors. There is a longitudinal electric field associating with the surface acoustic waves, originating from the materials polarization under mechanical deformation. When a semiconducting layer is deposited on the surface of a piezoelectric material, on which the SAW is
propagating, an interaction occurs between the SAW associated longitudinal electric field and free carriers in the semiconductor mesa.

In the absence of semiconducting overlayer, the energy generated by the wave gets stored in the evanescent electric field. The power flow in this case is given by

\[ P_{T1} = \frac{I_o^2}{2 j \omega k (\varepsilon_o + \varepsilon_s)} \]

Where \( I_o \) is the current generated per unit area of surface, \( 'k' \) is acoustic wave number, \( \varepsilon_o \) and \( \varepsilon_s \) are air and substrate dielectric permittivities.

In the presence of semiconducting film, the power flow becomes

\[ P_{T2} = \frac{I_o^2}{2 [k^2 \sigma_s + j \omega k (\varepsilon_o + \varepsilon_s)]} \]

\( k^2 \sigma_s \) is the shunt conductance due to conduction currents in the film overlay.

The total acousto-electric response is nothing but the difference between the two power flows, which entirely depends on the film conductivity

\[ P_T = P_{T2} - P_{T1} = -\frac{I_o^2}{2} \left[ \frac{k^2 \sigma_s}{j \omega kc_s [k^2 \sigma_s + j \omega kc_s]} \right] \]

This gives the general relationship between power transferred from the wave and the resulting changes in the wave propagation characteristics.

Due to interaction of the electric potential wave associated with the SAW and the mobile charge carriers in the semiconducting overlayer, both a velocity shift and increase in attenuation of the acoustic wave is observed. The fractional velocity shift and the acousto-electric attenuation per wave number \( \alpha/k \) due to the interaction of the charge carriers with the SAW is given by [Ricco et al. 1986]

\[ \frac{\Delta v}{v_o} = -\frac{K^2}{2} \frac{\sigma_{sh} z}{\sigma_{sh} z + v_o^2 C_s^2} \]

\[ \frac{\alpha}{k} = \frac{K^2}{2} \frac{v_o C_s \sigma_{sh}}{\sigma_{sh} z + v_o^2 C_s^2} \]
where $K^2$, the electromechanical coupling coefficient, is a measure of the piezoelectric strength of the substrate, $\sigma_{sh}$ is the sheet conductivity of the film and $Cs$ is the capacitance per unit length of the surface.

The maximum acousto-electric attenuation and the greatest rate of change in velocity both occur when the sheet conductivity is given by (figure 5.7)

$$\sigma_{sh} = v_0 C_s$$

where the capacitance per unit length, $Cs = \varepsilon_0 + \varepsilon_1$, is the sum of the permittivities of the substrate and the region above it.

![Diagram](image)

**Fig. 5.7:** Plot of fractional wave velocity shift $\Delta v/v_o$ and acousto-electric attenuation $\alpha/k$ as a function of sheet conductivity $\sigma_{sh}$ of a thin film overlay on a piezoelectric SAW substrate with coupling coefficient [Ricco et al. 1986].

The SAW velocity decreases as the attenuation goes up. Thus acoustoelectric interaction will influence the SAW propagation properties by resulting in attenuation due to ohmic loss and velocity change due to the piezoelectric stiffening of the material. The magnitude of the acousto-electric response is proportional to $K^2$, and thus is substrate dependent. The acoustoelectric effect can be many times greater than other dominating effects like mass loading effect. The “work point” of such a structure must be shifted to the high sensitivity region, where small variations in conductivity (under the influence of UV illumination) cause remarkable changes in
the wave velocity. Thus, to take full advantage of the high sensitivity offered by the SAW sensor, the sheet conductivity of sensing film must be tailored to a particular range.

The acoustoelectric interaction can be exploited with many different effects, such as field effect controlled or photon induced carrier changes, etc., leading to a family of novel multifunctional SAW devices.

5.1.5 Surface Acoustic Wave Ultra Violet Detector – A Review

The interaction of the SAW with the UV induced charge carriers in the photoconducting layer results in a phase shift and an insertion loss change, as functions of the incident light power and wavelength. Several authors have reported on the effect of UV radiation on the frequency response of SAW oscillator. The first report on SAW UV photodetectors was based upon nitride materials (GaN). Ciplys et al. 2001 reported the effect of UV radiation on the frequency response of a GaN/Sapphire based SAW oscillator operating at 220.9 MHz. A large frequency change of 100 kHz under UV light illumination was observed when the entire active area of the SAW device, including the transducers was illuminated by a mercury lamp, which later dropped by one order when only the active SAW path was illuminated while shielding the transducer, indicating the complexity of UV-intensity-to-frequency conversion mechanism. However the intensity of the UV light was not mentioned. A later study by Ciplys et al. (in 2002) reported a frequency downshift of 60 kHz in a GaN-based SAW oscillator device operating at 221.34 MHz under UV illumination. The spectral characteristics of the SAW oscillator response in the range from 330 to 600 nm was studied for remote visible–blind UV sensing. Sharma & Sreenivas (2003) reported a downshift of 170 kHz in a ZnO/LiNbO$_3$ SAW device operating at 36 MHz, when illuminated with UV light at 40 mW/cm$^2$. The hybrid ZnO/LiNbO$_3$ SAW device structure exhibited a relatively large shift in frequency at a much lower operating frequency of 37 MHz in comparison to the earlier reported GaN nitride-based SAW oscillator operating at high frequency. That study indicated the significance of large electromechanical coupling coefficient for maximum acoustoelectric interaction, and later sparked the interest of the scientific community in ZnO based UV photodetectors. Emanetoglu et al. (2004) fabricated a SAW ultra violet photo-detector using epitaxial multilayers of ZnO/MgZnO/piezoelectric ZnO on
sapphire, and have shown a significant change in insertion loss and phase shift with varying light intensity in the ultra-violet wavelength range. ZnO nanostructures grown by hydrothermal technique on 128°YX-LiNbO₃ have also been explored for making SAW UV sensors, and a maximum frequency shift of over 40 kHz has been reported in a dual delay-line SAW oscillator system. Chivuku et al. (2010) deposited chemically derived ZnO nanoparticles on LiNbO₃ substrate and observed significant phase changes over a wide range of UV wavelengths (280-370 nm) and reported a peak response at 345 nm for the fractional acousto-electronic SAW velocity change per unit power density of the order of 2.8 ppm/µW/cm². Recently Phan et al. 2012 reported a ZnO/Si two port resonator structure for UV detection by using third harmonic mode. The sensor showed a frequency shift of 400 kHz when illuminated with UV light at an wavelength of 380 nm and intensity 3 mW/cm², compared to a frequency shift of 10 kHz in the fundamental mode, indicating a 40-fold increase in sensitivity using third mode.

Table 5.1 presents a summary on various device structures and performance details of the devices reported so far. Efforts are still continuing to understand the effects of acousto-electric interaction with semi-conducting ZnO films/nanostructure.

**Table 5.1: Comparison of SAW sensors used for Ultraviolet (UV) light detection.**

<table>
<thead>
<tr>
<th>Photo Conducting Layer</th>
<th>Device Structure Substrate/Film and IDT Location</th>
<th>Operating Frequency MHz</th>
<th>Wavelength and Intensity</th>
<th>Observations</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>GaN</td>
<td>IDT/GaN/Sapphire</td>
<td>220.9</td>
<td>-</td>
<td>100 KHz Frequency Upshift</td>
<td>Ciplys et al. 2001</td>
</tr>
<tr>
<td>GaN</td>
<td>IDT/GaN/Sapphire Delay Line</td>
<td>221.34</td>
<td>330-400nm</td>
<td>60kHz Frequency Downshift</td>
<td>Ciplys et al. 2002</td>
</tr>
<tr>
<td>ZnO</td>
<td>ZnO/IDT/LiNbO₃ SAW Filter</td>
<td>37.0</td>
<td>365 nm 40mW/cm²</td>
<td>170 kHz Frequency Downshift</td>
<td>Sharma &amp; Sreenivas 2003</td>
</tr>
<tr>
<td>ZnO</td>
<td>ZnO/Mg₀.₂Zn₀.₈O/piezoelectric ZnO/r-Al₂O₃ Delay Line</td>
<td>711.3</td>
<td>365nm 2.32mW/cm²</td>
<td>107° phase shift and -22.8 dB losses</td>
<td>Emanetoglu et al. 2004</td>
</tr>
<tr>
<td>ZnO</td>
<td>ZnO/IDT/128°YX-LiNbO₃ Delay Line</td>
<td>37.0</td>
<td>365nm 450nW/cm² – 34µW/cm²</td>
<td>28kHz (34µW/cm²) 11mV attenuation (450nW/cm²)</td>
<td>Kumar et al. 2005</td>
</tr>
</tbody>
</table>
### 5.2 Objectives of the Present Work

Conventional MSM photodetector, discussed in the previous chapter, uses a power supply for biasing and needs to be electrically connected to its associated electronic measurement equipments. In a SAW UV detector, the UV induced photoconductivity in the ZnO overlayer is transmuted in the frequency domain of the piezoelectric interface. In the present study, the main focus is to integrate the optimised MSM photodetector structure with surface acoustic wave technology and develop oscillators in various topologies to measure the frequency and attenuation caused by acousto-electric interaction.

- Develop Precision Surface Acoustic Wave Oscillators based on delay line devices working at lower frequency and higher fractional bandwidth, and resonators – both one-port and two-port – operating at higher frequency and sharp resonance (high Q factor).
- Understand the design aspects for SAW delay line oscillators using open-loop oscillator configuration and analyzing the gain phase response.
• Understand RF Circuit design fundamentals to avoid wrong oscillator design.
• Integrate the ZnO thin film based MSM photodetector (investigated in the earlier chapter), with the SAW oscillator technology, and understand the impact of ZnO thin film deposition on delay line and resonator performance.
• Study the photoresponse property of SAW resonator using network analyzer method when illuminated with varying intensity of UV light.
• Study the photoresponse properties of Delay line based oscillators and understand their limitations at higher incident optical power.
• Understand the mechanism of frequency hooping in SAW delay line oscillators and their correlation with acoustoelectric interaction.
• Develop a hand held SAW sensor prototype with an on-board microcontroller based frequency counter with 1 Hz resolution.

5.3 Development of Surface Acoustic Wave Oscillator

Oscillator design is one of the least understood practices in rf engineering, and is considered to be the most complex one also [Sayre 2008]. For an amateur in the field of rf engineering, it looks quite easy to design an oscillator- just design a poor amplifier and turn on the power – and it will probably begin to oscillate, without any resonating element in the feedback network, or without any visible feedback circuit. At rf frequencies, a large difficulty in transistor based oscillators is a positive feedback originating due to the internal feedback capacitance between the transistor’s collector and its base. At a specific frequency, this capacitance will send an in-phase signal back into the base input from the collector’s output, which will create spurious oscillations. At higher rf frequencies, the distances between components and circuits can be a significant portion of a wavelength, and any simplistic methodology will adversely affect performance of amplifier, matching network and the entire oscillator circuit. Impedance matching is another significant issue in rf circuit design. Impedance mismatch between various sections of the entire circuit including the measuring equipment gives rise to standing waves formed by the reflected rf waves bouncing off the mismatched load, and then interacting with the forward wave, creating fixed peaks and valleys of voltage and current at every half wavelength along the trace. As a rule of thumb, whenever a trace or component dimension reaches more
than 1/20th of wavelength in separation, the designer has to be extremely careful of all 
rf-related issues, otherwise the design will not turn out as expected.

The working mechanism of a resonator based oscillator with CE transistor can be 
explained simply as: Whenever a pulse is applied to a tank circuit, it will ring at the 
tank’s resonant frequency, and will decay soon sinusoidally. An active device like a 
CE transistor amplifier is used to amplify and sustain this output. Thus oscillators use 
a small part of their output signal from the active device in order to send a 
regenerative or in-phase feedback signal back into their own input. An electronic 
oscillator is self starting. Whenever the sinusoids corresponding to the resonant 
frequency of the tank circuit build to a very high level, saturation of the active device 
occurs and surplus loop gain is then dissipated. The other point missing so far is the 
phase shift of the feedback network. The oscillator functions by feeding an 180° out-
of-phase signal back to its input, with this phase shift caused by the common-emitter 
configuration of the oscillator’s own amplifier. A method is required to shift this out-
of-phase output signal back to 0°, in order to obtain the necessary regenerative 
feedback.

Oscillator design and analysis is performed using two different methods – one method 
involves analyzing the open-loop gain and phase response versus frequency (Bode 
Response). The second method considers the oscillator as a one-port with a negative 
real impedance to which a resonator is attached [Rhea 1995]. The loop method 
provides a more complete and intuitive analysis as has been done in the present study 
while designing oscillators from delay-line based devices with high insertion loss and 
large bandwidth. The negative resistance method is more suitable for broad tuning 
oscillators operating above several hundred megahertz.

Surface acoustic wave (SAW) based oscillators have been extensively used for 
numerous sensing applications like pressure sensors, torque sensors, gas sensors, and 
biosensors. The basic block diagram of a SAW oscillator is shown in figure 5.8.

SAW oscillator works whenever there is a regenerative (positive) feedback so that the 
net gain of the closed network comprising the amplifier stage and the SAW delay 
line/resonator in the feedback stage is greater than unity and the total phase in the 
oscillator loop is n* 360. The Barkhausen criterion for sustained oscillations is:
The net gain of the open loop comprising the amplifier and the frequency determining element placed in the feedback route is greater than unity \( (> 0 \text{ dB}) \), and

The open loop has a total transmission phase of 0° or 360°.

\[
2\pi f\frac{L}{s} + \phi_A(n) + \phi_T(n) = 2\pi n
\]

where \( f \) is the oscillation frequency, \( L \) is the effective length of the acoustic delay line, \( s \) is the velocity of the acoustic surface wave, \( n \) is an integer (mode number) and \( \phi_A \) and \( \phi_T \) are the phase shifts of the amplifier and the (input and output) transducers.

\[\text{Fig. 5.8: General Structure of a SAW RF Oscillator.}\]

The amplifier should be designed in such a way so that the gain of the amplifier is greater than the attenuation of the feedback and power dissipation in the RF load.

In the case of resonators based oscillators, the phase slope of the resonator is very steep in the area of resonance frequency. This steep phase slope is necessary for better frequency stability.

To initiate the working of an oscillator, the inherent thermal noise is amplified in the active stage. The low broadband noise continuum is fed to the amplifier’s input by the feedback loop. In the noise continuum, there are frequency components at the desired oscillating frequency. These frequency components pass the frequency sensitive feedback stage without strong attenuation. Once the phase condition is met, the feedback signal is added to the input signal. This procedure is repeated cycle by cycle, and the output signal increases. The maximum oscillating level is limited by nonlinear
effects of the amplifier, or by some automatic level control. When the signal amplitude and the oscillation frequency are fixed, the oscillator is in the steady state mode. At steady state, the closed loop gain is unity.

The transient time (start-up time) depends primarily on the SAW resonator’s loaded Q-factor and the characteristic of the active circuitry.

5.3.1 RF Circuit Design Guidelines

(a) A solid ground plane should always be used when designing a PCB containing rf components. The purpose is to create an efficient 0V-reference node in the circuit that everything can be de-coupled to. The negative terminal of the power supply should always be connected to this ground plane. If this is not done properly, obscure circuit behavior might occur. Due to the low impedance of the ground plane, there will be no signal coupling between two nodes that are de-coupled to it. This is very crucial as there are signals on the board with a very large difference in amplitudes.

(b) At rf frequencies, even a short line will work as an inductor. As a rule of thumb, the inductance will be about 1nH per mm of length. The ground plane is therefore used, otherwise most ground lines will be longer than this and the rf circuit board will not function properly. All connections to the ground plane must be made as short as possible. A via should be placed close to every pad that is to be grounded. Also sharing of one via by two ground pads should be avoided, otherwise this can lead to cross talk between the two pads due to the impedance of the ‘via’ itself.

(c) To avoid the inductances of the lead terminals of discrete active and passive devices like transistor, diode, resistor, capacitors etc., surface mount devices (SMDs) should be used. With surface mounted PCBs all signal routing is done on the same side as where the components are mounted, and the ground plane will be on the opposite side.

(d) De-coupling capacitors should be placed as close as possible to the pins that are to be de-coupled. Only one de-coupling capacitor for each node should be used that is to be de-coupled. The value of the decoupling capacitors should be chosen so that their series resonance frequency is equal to the signal frequency they are to de-couple.
(e) Analog and digital circuitry should not be put on the same circuit board, as it creates a lot of complications. Digital signal lines often swing between ground and positive power supply, which for most of the applications, implies a peak to peak amplitude of \(~ 3V\). Switching time of digital signals is fast, normally in the nanosecond range. Due to the large amplitude and fast switching time the signal will contain a considerable amount of high frequency components, independent of the switching frequency. On the contrary, analog rf signals have comparably less signal amplitudes. The magnitude difference between the digital and RF signal can be significant to deteriorate the performance of rf circuits.

A double-sided PCB design was used for assembling the different oscillator circuits. The gain and phase characterization of SAW devices and the open loop amplifier was done using Agilent 8712ES RF Network Analyzer. Oscillator frequency was measured using Agilent 53132A universal frequency counter. The Fourier spectrum of the SAW oscillator comprising the fundamental frequency and its harmonics were examined using a Tektronix SA2600 spectrum Analyzer. Gold plated SMA edge connectors were used for connection between network analyzer, frequency counter ports and test cell.

### 5.3.2 Types of SAW Oscillator Circuits

Two different kind of SAW devices were used in the present study – SAW resonators (both one port and two-port) with higher operation frequency and smaller finger width/spacing, and SAW delay line filters with lower operation frequency (36 MHz) and larger finger width/spacing.

SAW resonators are popular as frequency stabilizing components in VHF and UHF oscillators because of their superior performance relative to LC resonators and low cost compared to bulk-wave crystal resonators. When connected in the feedback loop of an amplifier, 2-port resonators determine the oscillator frequency and guarantee an excellent signal-to-noise ratio. When used as one-port devices, they serve as narrowband frequency variable resistors and are generally connected between the base of a transistor and the ground to stabilize the current consumption of the transistor. HB434, a two port, 180° SAW resonator, available in a metal TO-39 case was used in the present study. It provides reliable, fundamental-mode, quartz frequency stabilization i.e. in transmitters or local oscillators operating at 433.92 MHz. One port
SAW resonator based study was carried out using device HR433.92A, operating at the same high frequency (434 MHz), and was procured from the same manufacturer (www.Hoperf.com). Figure 5.9 shows the S-parameters of two-port SAW resonator, with center frequency = 434 MHz, measured using a network analyzer.

A commercial SAW band-pass filter of center frequency 36 MHz, used for VIF stages in television was used in this work, and is named as device ‘A’. The 36 MHz SAW delay line was procured from two different SAW manufacturers, one provided us with Quartz based SAW filter, while the other SAW filter was based on piezoelectric ZnO thin film on glass substrate. The commercially available SAW IF filter (working frequency 36 MHz) works in inline configuration with one overlap weighted IDT in the input side and a uniform IDT in the output side [Ruppel et. al. 1993]. The SAW filter have features that are hard to beat by any other filter type – they have as close to a brick wall filter response as can be obtained. In the IF stage of the television receiver, the SAW filter selects the channel and forms the spectrum by decreasing the level of the sound carrier by 20 dB, and its group delay characteristic helps to counteract the distortion of the system. Figure 5.11 shows the S-parameters of SAW delay line filter, with center frequency = 36 MHz.

![Fig. 5.9: S-parameters of 2- Port SAW Resonator with center frequency = 434 MHz.](image)
5.3.2.1 One Port SAW Resonator based Oscillator

The oscillator is designed based on the common base Colpitt oscillator configuration. The Colpitt oscillator is stabilized with a SAW-resonator and allows for a simple design with only a few components besides the transistor and SAW-resonator [Figure 5.10(a)].

The transition frequency \( f_T \) of the transistor should be several Gigahertz \( (f_T > 5 f_r) \), where \( f_r \) is oscillation frequency) in order to ensure oscillator start-up. Transistors with a high transition frequency not only maintain the desired 180 phase shift but also provide a higher feedback gain. However, using a transistor with too high of a \( f_T \) will also increase the harmonic levels, and therefore it is not recommended to use state-of-the-art transistors with transition frequencies far beyond 10 GHz. In the present study, BFR93A—a wideband \( npn \) transistor available in the SOT23 package, having \( f_T = 6 \) GHz—was used in a common emitter (CE) arrangement with emitter degeneration.

Neglecting the internal transistor capacitance, the load and the parasitic capacitance of the PCB, the frequency of oscillation is determined by the resonance frequency of the parallel resonant circuit consisting of \( L_1 \) and the serial connection of \( C_1 \) and \( C_2 \), thus giving the resonance frequency as follows:

\[
f_p = \frac{1}{(2\pi\sqrt{L_1 C})}
\]

Where \( C \) is the combined capacitance of \( C_1 \) and \( C_2 \) and is expressed as

\[
C = \frac{C_1 \times C_2}{C_1 + C_2}
\]

The serial connection of \( C_1 \) and \( C_2 \) acts as a voltage divider, which ensures that not the entire output power of the transistor is fed back to the input. By doing so, the harmonics will be kept low. The ratio \( C_1/C_2 \) should be low so that there is a higher voltage drop across \( C_1 \) which ensures that the power fed back to the input is low. The figure of merit for the selectivity of the parallel resonance circuit is decided by the ratio \( L_1/C \). The higher the ratio \( L_1/C \), the lower the selectivity and therefore the higher the second harmonic. On the other hand, to get the maximum power out of the transistor, power matching must occur and therefore the output power of the oscillator changes with the inductance value of \( L_1 \) for a fixed ratio of \( L_1/C \) and \( C_1/C_2 \).
The oscillator starts up with an oscillation at the LC circuit. Since the Q-factor of a LC-resonator is limited and the resonance frequency can change by several percent due to tolerances, a SAW-resonator (SAWR) for frequency stabilization is required. The unstabilized frequency of oscillation of the LC-resonator circuit should be close to the desired one of the SAW resonator. The one port SAW-resonator has a high Q-factor, which will result in a long settling time that gives the oscillator enough time to start oscillation at the unstabilized frequency. Furthermore, the oscillator will not oscillate exactly at the resonant frequency of the SAW-resonator, but the frequency of oscillation will be shifted towards the unstabilized one.

5.3.2.2 Two-port SAW Resonator based Oscillator

The oscillation circuit of a two port SAW resonator is almost like a RF amplifier with a feedback loop. The two-port resonator based oscillator circuit is based on the feedback /Pierce oscillator configuration [Figure 5.10(b)].

The Pierce oscillator is an unconditional stable amplifier with a signal feedback from the output to the input. The feedback consists of two matching networks and the two port SAW resonator. For this oscillator a SAWR with 180° degree phase shift between input and output is used.

The load (often 50 Ohm) is connected to the collector by a small capacitor. This capacitor transforms the high impedance level of the collector down to the desired load impedance. The advantage of a Pierce Oscillator is the stable design and less parasitic effects. Disadvantage of this design is the higher amount of components (inductances).

For the matching networks a “PI” structure with two capacitors and one inductor is used ([Cp1,L1,Cp2],[Cp2,L2,Cp4]). The matching network between the collector and the transistor transforms the high collector impedance to lower source impedance for the SAW Resonator. The second matching network transforms the base impedance to the desired SAW Resonator load impedance. Additionally, the two matching networks create a certain phase shift to fulfill the phase condition on the oscillator loop.

The internal $C_T$ of the SAW resonator is a part of $Cp_1$ and $Cp_3$. The phase loop condition is satisfied by the two-port SAW device and the two tuning inductors $L_1$ and $L_2$. Other shunt elements ($Cp_2$ and $Cp_4$) can be eliminated by other parasitic
capacitors. For the feedback network to operate successfully it is required that the reactance of series elements and shunt elements have opposite signs. The SAW resonator is forced to operate in the positive reactance condition assuring the maximum phase slope.

The best working of the two port SAW Resonator (SAWR) based oscillator can be realized with an oscillating frequency close to the Resonator’s serial resonance frequency. The serial resonance frequency of the SAW Resonator (centre frequency) is very close to the minimum insertion loss point.

![Fig. 5.10:](a) Colpitt oscillator using one-port SAW resonator, and (b) Pierce Oscillator using two-port SAW resonator.

At the serial resonance frequency, the loaded Q-factor of the SAWR is at his maximum to get the best frequency stability.

### 5.3.2.3 Delay Line based Oscillator Circuits - Challenges

Designing an oscillator with a SAW delay line is normally tedious since the inherent insertion loss in the passband of SAW filter is high (>20 dB) and requires cascading of a few amplification stages so that the net gain is greater than unity (>0 dB). The final oscillator must have a large gain margin to take into account losses arising from different sensing phenomena like mass loading, acoustoelectric interaction, viscoelasticity effects etc.

The various scattering parameters of a SAW delay line operating at a center frequency of 36 MHz is shown in figure 5.11. As evident from the $S_{21}$ plot in figure, the SAW delay line has a high insertion loss of 22 dB in the pass band and it is necessary for
the amplifier to have a gain greater than 22 dB. But an open loop gain greater than unity or 0dB is not sufficient to design a SAW Oscillator based sensor for Ultraviolet detection due to the following reasons:

1. Any thin film overlayer on top of SAW device results in mass loading on the surface wave propagation of SAW. This mass loading is a function of the frequency of operation of the SAW device. Thickness monitors use this concept of mass loading in crystal oscillators to measure the thickness of thin film while in-situ deposition. Mass loading leads to change in acoustic velocity of the SAW device, along with attenuation of the wave.

2. As mentioned earlier, due to acoustoelectric interaction there is an attenuation of the acoustic wave. This attenuation in the amplitude of acoustic wave is reflected in the $S_{21}$ (or forward voltage transfer ratio/gain) parameter of the SAW device.

3. All the impedance matching between various sections of the electronic oscillator circuit is not exactly perfect. Impedance mismatches do occur and maximum power is not usually transferred to the load. The low input impedance of the measuring equipment can also bring down the gain of the amplifier resulting in destroying the oscillations.

Fig. 5.11: S-parameters of delay line based SAW Filter with center frequency = 36 MHz.
Keeping all these factors into consideration, it is imperative to design an oscillator circuit with a high gain margin. In the present work, the gain along with a passive impedance matching network was theoretically worked out to reduce the number of amplification stages. In the present work, a gain margin of around 10 dB was targeted.

5.3.3 SAW Delay Line Oscillator Circuit Design

Two of the requirements of Barkhausen criterion for sustained oscillations can be validated by using open-loop gain-phase technique, or the Bode plot. A reference is inserted into the input of the circuit, and the gain and phase of the signal is displayed as it passes though the oscillator’s open-loop circuit, while frequency is swept over a range (20-50 MHz) around the center frequency (36 MHz). The reference signal is considered to be at zero gain and zero phase shift. Any gain, either positive or negative, or any phase shift that occurs to this input frequency after it passes through the circuit, will be read in the Network analyzer display window, and displayed as gain versus frequency and phase shift versus frequency, in decibels and degrees. Figure 5.12 shows in a black box approach, the entire open loop oscillator, comprising the SAW filter, the amplifier and the impedance matching network.

![Diagram of SAW Delay Line Oscillator Circuit Design](image)

**Fig. 5.12:** Open loop configuration for measuring gain and phase response.
5.3.3.1 RF Amplifier Design

The targeted gain margin in the present study is 10 dB. The gain in excess of zero decibels at the phase zero crossing is referred to as the gain margin. Too high a gain margin is also undesirable as it constrains the amplifier to operate in the non-linear region thereby enhancing higher harmonic energy. We were successfully able to work out a design that resulted in reducing the number of cascading stages of amplifiers to unity. The various parameters essential for a successful CE transistor amplifier are: collector current, stability factor, voltage-divider biasing, $h_{fe}(\beta)$, transconductance etc. The amplifier design details for a desired gain are mentioned in Table 5.2.

Once the first Barkhausen criterion of unity loop gain is satisfied, the next step is focussed on the total phase shift of the network. If the total phase shift of the open-loop oscillator is not 0° or 360°, then a phase shift network is required to compensate for the remaining/ excess phase. Depending on the direction of the phase shift needed, either a Butterworth highpass or a Butterworth lowpass filter can be used, with the component values and number of poles optimized within a linear rf simulator software for the ideal phase shift.

5.3.3.2 Impedance Matching

For accurate gain and phase responses under open-loop configuration, the input and output impedances of the circuit should be at a common value of 50 ohm. If the terminating port impedances of the oscillator circuit is not taken into account, the gain and phase margins measure on the Network Analyzer would be incorrect. It is not always possible to obtain common input/output impedance at each port of the open loop oscillator. New generation SAW filters are been manufactured with input/output standard impedances of 50 ohm. Also standard bipolar transistors are being replaced with monolithic microwave integrated circuits (MMICs) which have 50 ohm input/output impedances. Some SAW filter need no input/output matching at all, and any external matching, as suggested by manufacturer may force the SAW into a desired mismatched condition, instead of a perfect conjugate match, to minimize phase and amplitude ripples within their passband. We have refrained from any input/output impedance matching of the 36 MHz SAW filter.
Impedance matching allows the maximum power transfer and the attenuation of harmonics to be achieved between stages.

![Diagram of impedance mismatch and matching](image)

**Fig. 5.13:** (a) Impedance mismatch between source and load, and (b) Impedance matching using L-network.

In the present study, considering the poor voltage driving ability of the CE transistor amplifier due to large output resistance, it was necessary to design an appropriate impedance matching network. The simple L network, so named due to its L shape, is the matching topology used for narrowband impedance matching. Figure 5.13(a) shows a two stage circuit without any matching network between the source and load, while figure 5.13(b) demonstrates the same two stages with a simple L network inserted, which is capable of matching a higher output impedance source to a lower input impedance load.

The L matching network has the disadvantage that the loaded $Q$ of the circuit cannot be chosen at the start of our calculations. To design a basic resistive – matching – only L network, used for matching higher output impedance source (amplifier/oscillator output) to a lower input impedance load (standard 50 $\Omega$), the network topology must be chosen as listed in Table 5.3.

The smith chart tool is an effective tool to confirm that both the input($S_{11}$) and the output impedances ($S_{22}$) of the open loop are matched at the frequency of interest, which is critical to the accuracy of oscillator open-loop design.
Table 5.2: Parameters for Amplifier Design at $V_{cc} = 9V$, $I_c = 10mA$, $\beta=100$, $V_{CE}=V_{cc}/2$.

<table>
<thead>
<tr>
<th>S.No.</th>
<th>Particular Parameters involved in Steps</th>
<th>Used Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>Trans-conductance of amplifier $g_m = 1/r_e' = \frac{\Delta I_c}{\Delta V_{be}} = \frac{I_c}{0.025}$ at 300°K</td>
<td>0.4</td>
</tr>
<tr>
<td>2.</td>
<td>Input resistance of amplifier, $r_n = \frac{\Delta V_{be}}{\Delta i_b} = \frac{\beta}{g_m}$</td>
<td>250 Ω</td>
</tr>
<tr>
<td>3.</td>
<td>Emitter Resistance $R_E = \frac{V_{cc}}{I_c \times 10}$</td>
<td>90 Ω</td>
</tr>
<tr>
<td>4.</td>
<td>Output Resistance $R_L = \frac{V_{cc} - V_{ce}}{I_c}$</td>
<td>450 Ω</td>
</tr>
<tr>
<td>5.</td>
<td>Ideal Voltage Gain of Amplifier $= \frac{\Delta V_{ce}}{\Delta V_{be}} = -g_m R_L$</td>
<td>180 (or 45dB)</td>
</tr>
<tr>
<td>6.</td>
<td>Voltage Divider Bias Parameters</td>
<td></td>
</tr>
<tr>
<td></td>
<td>i) $R_{Th} = \frac{(S-1)(1+\beta)}{(1+\beta-S)} \times Re$ (S → Stability Factor)</td>
<td>$R_{Th} = 899 , \Omega$</td>
</tr>
<tr>
<td></td>
<td>ii) $V_{Th} = I_b R_{Th} + V_{be} + I_e R_e$</td>
<td>$V_{Th} = 1.64V$</td>
</tr>
<tr>
<td></td>
<td>iii) $R_1 = \frac{R_{Th} V_{cc}}{V_{th}}$</td>
<td>$R_1 = 4.93 , \text{KΩ}$</td>
</tr>
<tr>
<td></td>
<td>iv) $R_2 = \frac{R_{Th} R_1}{R_1 - R_{Th}}$</td>
<td>$R_2 = 1.1 , \text{KΩ}$</td>
</tr>
<tr>
<td>7.</td>
<td>Total Input Resistance $= R_{Th} \</td>
<td></td>
</tr>
</tbody>
</table>

Table 5.3: LC Matching network considerations.

| 1.    | Find the natural Q of the Circuit | $Q = \sqrt{\frac{R_{High}}{R_{Low}} - 1}$ |
| 2.    | Find the reactance $X_P$ of the L network Connected in parallel | $X_P = \frac{R_{High}}{Q}$ |
| 3.    | Find the reactance $X_S$ of the L network Connected in Series | $X_S = Q \times R_{Low}$ |
| 4.    | Convert the calculated $X_P$ reactance into capacitor value | $C_P = \frac{1}{2\pi f X_P}$ |
| 5.    | Convert the calculated $X_S$ reactance into inductor value | $L_S = \frac{X_S}{2\pi f}$ |
5.3.3.3 Bode Plot Analysis

It is noted from the Bode-plot (Fig. 5.14) of the open loop configuration of the SAW delay line filter along with the RF amplifier and impedance matching network that the Barkhausen criterion is satisfied at three frequencies (35.6 MHz, 36.7 MHz and 37.9 MHz shown by three vertically drawn dotted lines in Fig. 5.14) showing the insertion loss to be minimum. The gain margin at these points is found to be around 8.4 dB which is found to be less than what predicted from theoretical calculations due to limitations imposed by various impedance mismatches. As we had mentioned earlier that the total phase shift of the open loop needs to be equal to 0° or 360°, and the remaining or excess phase has to be compensated using some filter network. Fortunately, we found that the phase response of the SAW filter varies in periodic manner. The total phase shift of the open loop oscillator comprises of three points in the pass band of the SAW filter having values of zero degrees. Also these three points were found to coincide with the points at which maximum gain was found (35.6 MHz, 36.7 MHz and 37.9 MHz). Therefore, the SAW filter if connected in the feedback loop of closed loop oscillator has equal probability to oscillate at the three frequencies - 35.6 MHz, 36.7 MHz and 37.9 MHz.

The loaded $Q$ of an oscillator governs its phase noise and frequency drift. The higher the $Q$ is, the more stable the oscillator is over temperature, along with possessing lower phase noise figures. The loaded $Q$ must never be permitted to degrade below 5 or 10, and should preferably be much higher to lower its phase noise. The loaded $Q$ is measured from the open-loop Bode Plot using the formula below:

$$Q = \frac{f_o}{\text{3dB BW}}$$

where $3dB BW$ is the bandwidth (Hz) of the oscillator’s gain (S21) at its half power points.

From figure 5.14, we can estimate that the loaded $Q$ of the delay line based open-loop oscillator is very low (~10) due to the low operating frequency and the wide bandwidth of the filter. This is one of the biggest disadvantages of SAW delay line compared to resonators while developing oscillators.
5.3.3.4 Final Circuit Design and Oscillator Analysis

The final circuit diagram of the 36 MHz SAW oscillator along with the values of various electronic components is shown in Fig 5.15. Since the open loop gain is maximum at the centre frequency of the SAW filter i.e. 36.7 MHz, the oscillator exhibited very stable and sustained oscillations at 36.7 MHz. A complete spectra of the SAW Oscillator in the frequency domain as obtained from a spectrum analyzer is shown in Fig. 5.16(a) and (b). The output power of the SAW oscillator is 4 dBm at 36.7 MHz, which was significantly higher over the power levels of the first and second harmonics at 73.4 and 110.1 MHz respectively. The quality factor (f/Δf) of the fundamental frequency was estimated to be 4587, which is very high as compared to conventional LC filters. This is the beauty of the circuit design workflow we followed so far. The loaded $Q$ of the open loop oscillator was found out to be ~10 due to the inherent property of the delay line. But the $Q$ of the final oscillator is very high (4587) and this is mainly due to the phase frequency relationship of the SAW filter. The strange periodic phase frequency relationship as shown in figure 5.14 helps in locking only few particular frequencies in the pass band where the first Barkhausen criterion (gain >= 0dB) is satisfied. This frequency selective behavior of the SAW delay line

Fig. 5.14: Gain phase response (Bode-plot) of open-loop oscillator circuit measured using Network Analyzer.
helps in enhancing the loaded $Q$ of the final closed loop oscillator as evident in the present study.

**Fig. 5.15:** Circuit diagram of 36 MHz delay line based SAW oscillator along with impedance matching network.

**Fig. 5.16:** Frequency spectra of the SAW oscillator (a) showing the high Quality factor ($Q=4587$) of fundamental mode, and (b) comparison of fundamental mode and harmonics as captured by a spectrum analyzer.
5.4 Effect of Acoustoelectric Interaction in SAW UV Sensors - Results

A comprehensive study on ZnO MSM photodetectors operating in the Ultraviolet range of the spectrum was done in chapter, and all the device performance like Photoresponsivity, gain mechanisms, frequency characteristics, persistence photoconductivity, wavelength dependence etc were discussed in detail. It was established that 200nm thick ZnO thin film deposited at 30mTorr and 60% O\textsubscript{2} showed the highest photocurrent with fast rise and fall temporal characteristics. In this section, the same MSM photodetectors with different IDT specifications, denoted earlier with Device ‘A’ and Device ‘B’ were utilized for developing surface acoustic wave oscillators in different configurations – delay line oscillator, one-port and two-port resonator oscillator. Figure 5.17 shows the various SAW devices used in the present study. The header of the commercial SAW devices were cut open using a diamond saw, and while cutting the debris falling onto the SAW device was washed away in isopropyl alcohol followed by soft baking at 80°C. The un-packaged devices were loaded inside the sputtering chamber for depositing a 200 nm thick photo conducting ZnO overlay at room temperature using rf magnetron sputtering technique.

![Fig. 5.17](image)

Fig. 5.17: (a) Uncut Commercial SAW devices –delay line filter, one-port and two-port SAW resonators, along with final devices optimized for ultra violet photoconductivity, and (b) Top view of ZnO thin film deposited on SAW devices. Film deposited over the entire active area of the SAW, including the transducers.

5.4.1 Effect of ZnO Deposition on SAW Sensor Performance

The mass sensitivity of SAW sensors is highly dependent on the working frequency of the SAW. The resonators with a center frequency of 434 MHz showed a tremendous downshift of frequency when ZnO thin film was grown on top of the
IDT/SAW structure. Table 5.4 shows the effect of ZnO thin film thickness on the frequency shift from the center frequency of uncoated SAW. Losses are also observed as thickness of ZnO is increased from 50-200nm. Figure 5.18 and 5.19 shows the effect of ZnO thin film thickness on the $S_{11}$ and $S_{21}$ parameters of one-port SAW and two port SAW resonators. The SAW resonators are more sensitive to the ZnO thin film overlayer, although both the depositions of same thickness were done at the same time.

On the contrary, ZnO thin films of varying thickness from 50nm to 1µm showed negligible change in the center frequency, and loss induced was also very low (Figure 20(a) and (b)). This proves the utility of the device for functionalizing diverse range of sensitive thin films that can be grown on top of the SAW.

**Fig. 5.18:** Change in the $S_{11}$ and $S_{21}$ scattering parameter of 2-port SAW resonator upon deposition of 50-200nm photoconducting ZnO thin film.

**Fig. 5.19:** Change in the $S_{11}$ and $S_{21}$ scattering parameter of 1-port SAW resonator upon deposition of 50-200nm photoconducting ZnO thin film.
Table 5.4: Effect of ZnO deposition on the attenuation and frequency (or phase velocity change) of one-port and two-port SAW resonators (operation frequency = 434 MHz)

<table>
<thead>
<tr>
<th></th>
<th>One Port SAW Resonator</th>
<th>Two Port SAW Resonator</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>50 nm</td>
<td>200 nm</td>
</tr>
<tr>
<td>Loss (dB)</td>
<td>0.485 dB</td>
<td>2.519 dB</td>
</tr>
<tr>
<td>Frequency Shift (∆f)</td>
<td>2.4 MHz</td>
<td>9 MHz</td>
</tr>
<tr>
<td></td>
<td>50 nm</td>
<td>200 nm</td>
</tr>
<tr>
<td>Loss (dB)</td>
<td>2.506 dB</td>
<td>4.617 dB</td>
</tr>
<tr>
<td>Frequency Shift (∆f)</td>
<td>2.4 MHz</td>
<td>8.2 MHz</td>
</tr>
</tbody>
</table>

Fig. 5.20: Changes in the (a) $S_{11}$, and (b) $S_{21}$ scattering parameter of 36 MHz SAW delay-line upon deposition of 100nm-1μm photoconducting ZnO thin film.

5.4.2 Ultra Violet Photoresponse Properties of ZnO Thin Film based SAW Sensors

In the following section, the effects of photoconducting ZnO thin film on the response characteristics of SAW sensors upon UV illuminations of varying intensity are studied. There is an interest to know the transmutation of photoconductivity in the frequency domain and study the effect of acoustoelectric interaction on the frequency and amplitude of surface waves in devices of similar piezoelectric substrate but different IDT patterning. The SAW resonators and delay line operating at different frequencies show attenuation upon UV exposure, accompanied by frequency downshifts.

5.4.2.1 SAW Resonators

The response of the SAW resonator was monitored by measuring the $S_{21}$ and $S_{11}$ parameter with a Network Analyzer. The measured $S_{21}$ value clearly shows that there is an attenuation of the acoustic wave, accompanied by frequency downshifts (Figure
5.21(a) and (b)). The loss in the amplitude of two port SAW resonator shows a linear relationship with increase of UV illumination, as it rises to almost -1.9dB at 238µW/cm² (figure 5.22).

![Two port SAW resonator](image)

**Fig. 5.21:** Effect of UV illumination on two port SAW resonator when measured in (a) $S_{21}$ mode, and (b) $S_{11}$ mode.

![Two port SAW resonator](image)

**Fig. 5.22:** (a) Zoomed portion of the resonant peak of two-port resonator showing attenuation and frequency change, and (b) linear variation of attenuation with incident UV illumination.

The Colpitt and Pierce oscillators based on one-port and two-port resonators were found to be ineffective for their implementation as Sensors. The problem does not lie with the design of the oscillators, but with SAW resonators itself. As evident from table 5.4 the 200nm thick photoconducting ZnO deposition is responsible for almost 9 MHz shift from the center frequency and hence the frequency selectivity of the SAW
oscillators is hampered badly. Additional LC tuning and phase shift network needs to be designed which has to be integrated with the oscillator module itself which will compensate for the ZnO related loading of the resonators. The resonator based oscillators are designed to work with a precise frequency and hence have wide range of applications, but for sensor purposes, the design of the oscillators need to be altered after measuring the gain-phase response of resonators when thin films are coated on top of it. This limits the use of pre-fabricated oscillators based on one-port (Colpitt) and two-port (Pierce) resonators, as the design itself requires the estimation of various SAW related changes beforehand.

5.4.2.2 SAW Delay Line Based Oscillator

The presence of a thin over-layer of 200 nm thick photo conducting ZnO film showed a significant downshift in the SAW oscillator frequency, which increased with increasing UV light intensity as shown. The real time data measured using frequency counter Agilent 53132 in Fig. 5.23 and Fig. 5.24 shows a linear downshift in frequency (2.2 to 19.0 kHz) with varying UV intensity (20-100 \( \mu \text{W/cm}^2 \)) arising due to fractional velocity change caused by acoustoelectric interaction.

Two distinct stages in the frequency downshift are noted from figure 5.23, and could be distinguished as a relatively fast initial drop in the frequency followed by a characteristic slow drift to a lower value which depended strongly on UV light intensity. When the UV illumination is switched off, the recovery is again very fast, but is unable to reach the initial value especially at levels of UV intensity (40 to 80 \( \mu \text{W/cm}^2 \)).

The amplitude of the SAW oscillator was also measured using an Oscilloscope and it was found that the SAW amplitude is damped due to damping of the acoustic wave by photogenerated charge carriers. Figure 5.25 and 5.26 shows the amplitude decay with the variation of UV light intensity from 0-80\( \mu \text{W/cm}^2 \). The variation of the attenuation is found to be quite linear with UV intensity. The data also reflects the fact that higher incident optical power has the potential to destroy SAW oscillation in the electronic circuit. This problem can be dealt with some alternate amplifier design with higher gain, or an automatic gain control (AGC) mechanism to adjust the gain automatically with the increasing UV induced losses. Another point that can be inferred from comparing both the frequency measurement and the amplitude measurement is the
dynamic range of the individual measurements. An illumination level of <100 µW/cm² produces a frequency downshift of around 18kHz, and we can measure the frequency shifts up to a very high precision using good frequency counters, whereas the amplitude changes by only 0.6 V. The amplitude of any electronic oscillator is not a very stable value with superior precision (ppm), whereas frequency of a resonator based oscillator is very stable (Q factor of delay-line/resonator based oscillator varies from $10^3$-$10^6$).

**Fig. 5.23:** Frequency downshift at different levels of UV illumination, 20–80 µW/cm². Inset: Two step frequency downshift due to increased attenuation at 100 µW/cm².

**Fig. 5.24:** Frequency downshift $\Delta f_1$ relating to fractional velocity change, measured at different levels of UV illumination, 20–100µW/cm².
Fig. 5.25: Attenuation of voltage output of SAW delay line oscillator upon various levels of illumination (0-80 µW/cm²), as seen on an oscilloscope.

Fig. 5.26: Linear variation of attenuation of SAW oscillator amplitude with incident UV illumination (0-80 µW/cm²).

A further increase in UV light intensity to 100 µW/cm² and above, led to a distinctive observation with repeated on/off exposure cycles to UV light, and the fine changes are shown in the inset for each cycle in Fig. 5.27. In the first cycle-I an initial downshift of 19 KHz is seen, followed by a constant region, and then a sudden drop by 43 KHz resulting in a total frequency downshift of 62 KHz. When the UV light is switched off the oscillator frequency quickly recovers back to its initial value (dark value). A similar trend was observed for every repeated cycle of measurement. It is interesting
to note that in subsequent exposure cycles, the initial downshift remains unchanged (17 to 19 KHz), and the constant region continues to extend longer in subsequent cycles. The total frequency downshift in cycle –II is 711 KHz, and thereafter from cycle-III and beyond the frequency downshift remains constant ~ 981 KHz.

**Fig. 5.27:** SAW oscillator frequency response for repeated on/off cycles of UV illumination at 100 $\mu W/cm^2$.

### 5.4.2.3 Frequency Hopping

The minute change in the fractional velocity of acoustic wave due to acousto-electric interaction is reflected as a change in the center frequency of SAW oscillator, whereas minute change in attenuation is reflected as a decrease in the amplitude of the SAW oscillator output. The output of the SAW oscillator when observed with a frequency counter shows the effect of acousto-electric interaction on the velocity. But a significant attenuation of the acoustic wave upon UV illumination can affect the oscillation criteria, and the combined effect of velocity changes and attenuation can give a complex frequency response as seen in the present study. In order to see the effects of attenuation, the $S_{21}$ parameter was analyzed using a Network Analyzer to understand the large frequency downshift observed in cycle-III and above. The normal band pass response (Fig 5.28) in dark consists of 3 amplitude ripples in the pass band centering at 35.6, 36.7, and 37.9 MHz marked as 1, 2 and 3 in Fig. 5.28.
These are the frequencies where the criteria (maximum gain and zero phase) are satisfied for stable SAW oscillation. Stable oscillations were seen at 36.7 MHz (point 2) where the insertion loss (IL) was minimum (\(I_{L2} < I_{L1} < I_{L3}\)). However, the previous state (\(I_{L2} < I_{L1} < I_{L3}\)) at these 3 points of the SAW response have changed after illumination at 100 \(\mu\text{W/cm}^2\) due to attenuation induced by acoustoelectric interaction. After exposure to UV for 3 seconds the induced loss is maximum (-4.21 dB) at the centre frequency (36.7 MHz). This leads to a reduction in the ultimate stop-band rejection level and shows up as two more pronounced amplitude ripples. It may be noted from Fig. 5.28 that after UV exposure, the insertion loss at point 1 is minimum as compared to points 2 and 3 (\(I'_{L1} < I'_{L2} < I'_{L3}\)). As a result, now the Barkhausen criteria is satisfied at a new point (centre of ripple 1) and frequency hopping to 35.6 MHz is observed (Table 5.5). When the UV light is withdrawn, the original state (\(I_{L2} < I_{L1} < I_{L3}\)) is restored and the sustained oscillations revert back to the dark level (36.727 MHz). Thus a frequency hopping behavior is seen due to the simultaneous interplay between the change in attenuation and a change in acoustic velocity, resulting in a huge frequency downshift (\(\Delta f = \Delta f_1 + \Delta f_2\)) due to acousto-electric interaction.

Fig. 5.28: (Color online) Variation in S21 transmission spectra before and after UV illumination at 100 \(\mu\text{W/cm}^2\), and the recovery after illumination.
Table 5.5: Insertion loss at the center point of the ripples of a SAW pass band before and after UV illumination at an intensity of 100µW/cm².

<table>
<thead>
<tr>
<th>UV On/Off State</th>
<th>Insertion Loss at Ripple-1</th>
<th>Insertion Loss at Ripple-2</th>
<th>Insertion Loss at Ripple-3</th>
</tr>
</thead>
<tbody>
<tr>
<td>Before UV exposure</td>
<td>-23.52 dB (I_L1)</td>
<td>-23.06 dB (I_L2)</td>
<td>-24.17 dB (I_L3)</td>
</tr>
<tr>
<td>3 sec after UV exposure</td>
<td>-26.73 dB (I'_L1)</td>
<td>-27.21 dB (I'_L2)</td>
<td>-28.27 dB (I'_L3)</td>
</tr>
<tr>
<td>Recovery after UV ‘Off’</td>
<td>-24.78 dB</td>
<td>-24.47 dB</td>
<td>-25.65 dB</td>
</tr>
<tr>
<td>UV induced Losses (δI_L=I_L-I'_L)</td>
<td>-3.21 dB</td>
<td>-4.15 dB</td>
<td>-4.10 dB</td>
</tr>
</tbody>
</table>

5.5 Development of Hand-held SAW Sensor Prototype

Typical sensing effects constitute relatively small frequency deviations from their unperturbed resonance, from several kHz to a few MHz. Measuring such high frequency (typically of SAW resonators working at hundreds of Megahertz) in the very high frequency (VHF) and ultra high frequency (UHF) bands require very expensive RF instrumentation. Normal Frequency Counters have ports to measure frequencies accurately till 250 MHz. And cost of including an additional channel for measuring high frequency is very high. There is a need to move the RF spectrum down to the audio spectrum, and developing low cost read-out counters so that the final sensors are portable and can give a frequency resolution of 1 Hz for monitoring frequency changes in thin film based SAW sensors.

5.5.1 Design Workflow

The following steps were implemented for the development of the hand-held SAW sensor prototype, capable of displaying the sensor signal, without the aid of any expensive measurement equipments.

5.5.1.1 Differential Frequency and Mixer Design

In order to avoid other physical and environmental variables such as temperature, humidity and other strain/stress due to pressure fluctuations in the ambient of the SAW surface, to affect the performance of SAW sensor and picking spurious frequency changes, and extract only the relevant signal without any interference, researchers have employed dual device mode sensing. In this configuration, an
identical SAW device (called reference device) is placed next to the sensor device and exposed to identical conditions. The dual configuration provides a common mode rejection for such external influences such as temperature, drift etc. [Penza and Cassano Sens B 68]. It also helps to bring down the sensor output frequency to kHz range eliminating the need to carry expensive frequency counters.

Mixer circuits are designed to yield both a sum and a difference frequency at a single output port when two distinct input frequencies are inserted into the other two ports. The lower received difference frequency available at the mixer’s output is called the intermediate frequency (IF).

The mixing process involves heterodyning the reference (or local oscillator LO) and sensing oscillator (SO) frequencies. Unbalanced mixer have an output consisting of $f_{LO}$, $f_{SO}$, $f_{SO}-f_{LO}$, $f_{SO}+f_{LO}$, and other spurious harmonics. A double balanced mixer (DBM) is required in this case, as it supplies superior IF-LO-SO inter-port isolation, while outputting only the sum and difference frequencies of the sensing oscillator (SO) and the reference oscillator (LO), and attenuating both the LO and SO signals.

SA602A, a low-power VHF monolithic double-balanced mixer with input amplifier, on-board oscillator, and voltage regulator has been used in the present study. This IC is available in an 8-lead SO (surface-mount miniature package). A low pass filter is designed to ensure that only the difference frequency is passed on. The differential frequency $\Delta f$ from the output of low-pass filter is

$$\Delta f = (f_{ref} - f_{sample})$$

Although the mixing circuit technique will significantly reduce the effects of common mode interference, there is always the possibility that interference could compound, and, therefore increase measurement errors.

5.5.1.2 Frequency Counter Design

The design of the frequency counter is implemented using the microcontroller PIC 16F84A. Working principle of the counter is quite simple. The counter will measure the time difference between two peaks, most likely by measuring the number of peaks
in a given period of time and calculating the time for one cycle. The frequency is then obtained by finding the inverse of that value.

The frequency output of the mixer circuit is passed through a low pass filter to filter the differential frequency or the sensing frequency. This is amplified to increase the voltage level of the sinusoidal IF frequency necessary for the next stages in the frequency counter design. There is a need to convert the sinusoidal voltage signals to appropriate levels to serve as digital signals for the NAND gates. To achieve this, the signal is passed through a DC blocker and a limiter. An N channel Junction FET and a PNP BJT are used to amplify the signal to 0-5 levels. Finally, the digital pulses (0V – low, and 5V = high) are generated with the same frequency as the input signal. This will allow for a greater tolerance of input signals with varying DC offset levels as well as larger amplitudes.

The adjusted input signal is then passed through a series of NAND gates (74HC132) before being fed into the microcontroller (PIC 16F84A). One of the outputs of the PIC microcontroller is used as feedback to the series of NAND gates (74HC132) to turn the signal on or off. The Schmitt trigger (74HC132) is used due to its better tolerance to signal noise. The final signal is received by the microcontroller, which counts the number of pulses in a specific time period. This digital data is simultaneously been fed to the LCD display (Hitachi 20x4 character LCD with HD44780 Controller Chip). The microcontroller PIC16F84A begins with a gate time of 0.1 seconds. For higher frequencies, lower gate time is used in order to obtain a better average value. It executes a subroutine to test if the measured frequency is in the MHz range. If so, it calculates the appropriate position of the decimal and sends that value as well as units “MHz” to the LCD. If the input frequency is lower, it executes a different subroutine to test if the measurement is in the kHz range. In a similar fashion as aforementioned, if the frequency is in kHz range, it determines the location of the decimal and appends the units “kHz” to the readings to the reading before sending it to the LCD. If the frequency is even lower, it executes yet another similar subroutine but for the “Hz” range instead.
5.5.1.3 Printed Circuit Board Design

The first step involves collecting all the datasheets for all the components that will be used. The PCB fabrication is a complex process and involves a lot of steps. Computer-aided Design (CAD) software like ORCAD or NI Multisim/Ultiboard is used to create a schematic. This intuitive and easy-to-use software platform combines schematic capture and industry-standard SPICE simulation into a single integrated environment. The Schematic capture tool is used to create schematic using available symbols from the symbol libraries and interconnecting them with the ‘wire’ tool. This means that we are adding different components on to our board and connecting them with wires. Passive components like resistors, capacitances, inductors, and active components like transistors, FETs along with connectors, headers, electromechanical components, DIP ICs are all available in the library of any standard PCB CAD software. Multisim offers multiple ways to analyze the circuit using virtual instruments like multimeter, oscilloscopes etc. Figure 5.29 shows the final circuit diagram for the hand-held SAW sensor prototype (SAW Oscillator circuit is not shown in the figure here as they are developed on separate PCBs)

Once the schematic is ready, the next step involves the choice of footprints of the components that will be used in the final design. Footprint is a packaging view of the component that includes the holes through our board or pads for surface mount devices (SMDs). All the standard components are nowadays available in both through hole and SMD packages, and the footprints need to be selected to meet the mechanical requirements for the components. Inside the CAD software, the name of the selected footprint has to be attached to the components symbol by editing the properties for each component. Any final interconnection between the components on the printed wiring board (or PCB) can be made using the ‘place wire’ tool.

After completing the schematic and the footprint selection, we generate the ‘netlist’ and import it to PCB layout software to complete the board layout. Layout software like Ultiboard will automatically insert footprints into the board based on the information given earlier. A board outline is placed on the workplace with the
components (above the board outline) ready to be placed. The next step of PCB design is layout the components. Placing the components with proper routing takes practice to perfect. The power and ground planes are also defined during this stage. Once the board layout is completed and routing is done, technology files called ‘Gerber’ files are created. These files are required by professional PCB manufacturers to develop PCBs in large scale production. However, we developed the PCB in the lab using simple laser toner transfer technique to transfer the pattern on the double sided copper plated FR4 board. Then the board was etched using ferric chloride solution, followed by removal of toner using acetone. The SMD components were finally soldered on the etched PCB using temperature controlled soldering station. The soldering is a very careful procedure as excess heat can damage the SMD components. Figure 5.30 shows all the stages of SAW sensor prototype design.

Fig. 5.29: Final schematic of the hand held SAW sensor prototype built using microcontroller PIC16F84A.
Fig. 5.30: (a) Layout of the circuit with manual routing as done on ‘Ultiboard’, (b) & (c) Top and bottom 3-D view of the final prototype as seen on the Ultiboard virtual environment, (d) & (e) etched PCB developed using in-house laser toner transfer method, (f) Final photograph of the hand-held SAW sensor Prototype (only the mixer and counter circuits assembled here, RF oscillators on separate PCBs to avoid rf-digital interference).
5.6 Conclusions

A stable oscillator has been designed using a low cost commercial SAW Television TV-IF filter with apodized IDT operating at 36 MHz, and the importance of various steps involved in the design have been highlighted. A linear change in the frequency downshift (2.2 to 19.0 kHz) is observed under low level UV illumination (20–100 \( \mu \text{W/cm}^2 \)) and relates to the fractional velocity change due to acoustoelectric interaction. However, at 100 \( \mu \text{W/cm}^2 \) the observable change is masked by a huge frequency downshift that occurs due to increased attenuation and leads to a frequency hopping behavior. The frequency hopping observed in the present study appears to be specific to a delay line based SAW oscillator, because a SAW bandpass filter has a tendency to pick any frequency from the passband where the Barkausen criterion is satisfied. Alternate oscillator designs based on SAW resonator with a sharp spectral response can overcome the difficulty faced due to increased attenuation upon UV illumination. The huge frequency downshifts as noted in the present study with a delay line based SAW oscillator can lead to a misinterpretation of the acousto-electric interaction affecting the fractional velocity change.

Frequency measurement of oscillators along with real time data acquisition is a slow process due to finite gate time required by the measurement electronics. This is disadvantageous for monitoring the transient photoresponse behavior of SAW sensors. Whereas photocurrent measurement using MSM photodetectors are fast as modern oscilloscopes have huge sampling rates (in giga-samples per second).

The development of SAW delay line based oscillators, using either commercially available devices or lab-made devices is tricky, laborious and not as effective as many researchers worldwide report to be. Spurious frequency close to the center frequency may be achieved, as we have found in our preliminary studies(not discussed here), and can misled us while using the devices for sensor applications, as they will not respond directly to the excitation (UV or any vapor analyte etc), and this can undermine the role of SAW technology for sensor applications. However careful rf circuit design and proper knowledge can help a lot in pitfall prevention.

Rf circuit design takes time to perfect and researchers should be well acquainted with the small nuances that can lead to a completely disastrous performance of the SAW sensors.