

Introduction to SAW-Filters

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1 PIEZOELECTRICITY

Piezoelectricity describes the interaction between the displacement of electric charge or electric polarization on the surfaces of solids and the mechanical stress acting on them. A distinction is made between the direct piezoelectric effect—the generation of an electrical voltage through mechanical deformation—and the inverse piezoelectric effect—deformation resulting from an electrical voltage in the material. The inverse piezoelectric effect is particularly important in the generation of ultrasonic waves and therefore plays a decisive role in the development of acoustic transducers. A distinction is made between the directions of polarization in the longitudinal, transverse, and shear effects. Fig. 1 illustrates the inverse piezoelectric effect in the longitudinal and transverse directions in a simplified unit cell of alpha quartz [1].

Piezoelectric materials are dielectrics—electrical insulators that can be polarized by an external electric field—which have a crystal structure without an inversion center, allowing polarization to occur even through distortion of the crystal structure. These solids can also exhibit pyroelectric properties—the ability to generate a transient voltage in response to temperature changes—and ferroelectric properties—spontaneous electrical polarization that can be reversed by applying an electric field. Fig. 2 illustrates the relationship among these materials. Ferroelectric materials are particularly important in applications that place high demands on piezoelectricity and mechanical flexibility. Some of the most commonly used materials are quartz, lithium niobate (LiNbO_3), and lithium tantalate (LiTaO_3). The choice of the most suitable material depends on the specific application. For high precision and narrow bandwidths at variable temperatures, quartz may be the preferred choice. For applications involving high frequencies and wide bandwidths—especially when high load-bearing capacity is important—lithium niobate may be more suitable. Pure silicon,

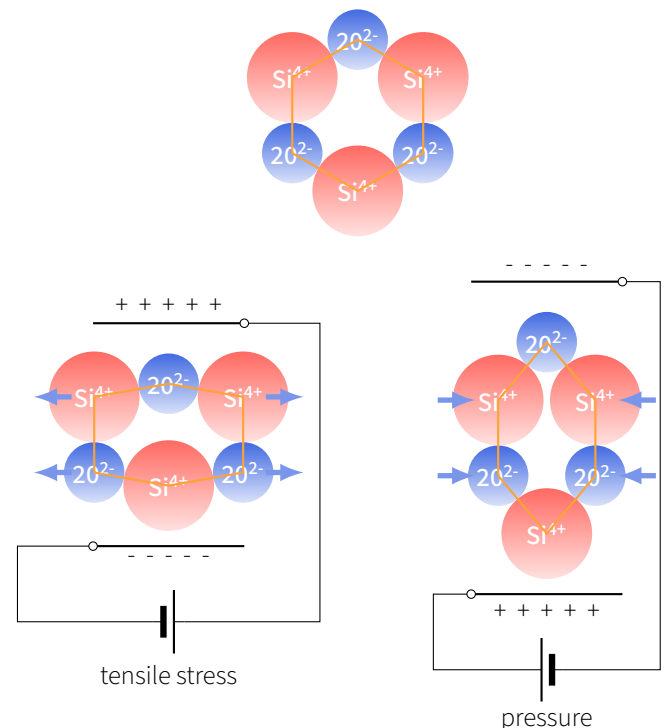


Figure 1 | Inverse piezoelectric effect in Quartz An electric field induces mechanical forces in a solid, which cause the crystal structure to deform.

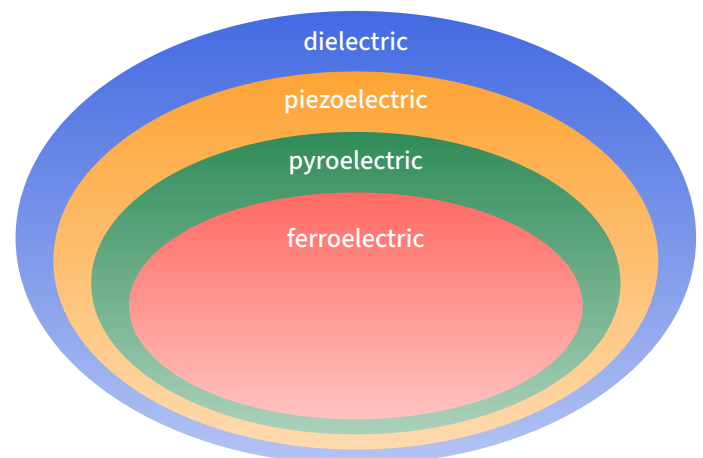


Figure 2 | The relationship between dielectric, piezoelectric, pyroelectric, and ferroelectric materials. Piezoelectric materials are a type of dielectric material. However, not all dielectrics are piezoelectric.

Symbol and SI unit	Name	Definition
$d_{ij} \left[\frac{\text{Q}}{\text{N}} \right] = \left[\frac{\text{m}}{\text{V}} \right]$	piezoelectric charge constant, piezoelectric strain constant	$\left. \frac{\partial D_i}{\partial T_j} \right _E = \left. \frac{\partial S_j}{\partial E_i} \right _T$
$g_{ij} \left[\frac{\text{Vm}}{\text{N}} \right] = \left[\frac{\text{m}^2}{\text{Q}} \right]$	piezoelectric stress constant (electrical voltage)	$-\left. \frac{\partial E_i}{\partial T_j} \right _D = \left. \frac{\partial S_j}{\partial T_i} \right _T$
$e_{ij} \left[\frac{\text{Q}}{\text{m}^2} \right]$	piezoelectric tensile stress constant	$-\left. \frac{\partial T_i}{\partial E_j} \right _S = \left. \frac{\partial D_j}{\partial S_i} \right _E$
$h_{ij} \left[\frac{\text{N}}{\text{Q}} \right] = \left[\frac{\text{V}}{\text{m}} \right]$	piezoelectric stiffness constant	$-\left. \frac{\partial E_i}{\partial S_j} \right _D = \left. \frac{\partial T_j}{\partial D_i} \right _S$
D_i : Electric displacement component $\left[\frac{\text{Q}}{\text{m}^2} \right]$ T_i : (mechanical) stress component $\left[\frac{\text{N}}{\text{m}^2} \right]$ S_i : strain component $[\]$ E_i : Electric field component $\left[\frac{\text{V}}{\text{m}} \right]$		

Table 1 | Piezoelectric constants.

on the other hand, does not fall into the category of piezoelectric materials due to its symmetrical crystallography. [2].

The applications of piezoelectric materials are extremely diverse; depending on the specific application, they can be used as actuators (e.g., fine adjustment), vibrators (e.g., ultrasonic applications), resonators (e.g., clock signal generation), or transducers (SAW filters). This therefore requires fundamental equations that can describe their properties and enable the designer to determine their behavior under operating conditions using appropriate scaling. In many cases, piezoelectric processes in materials can be described as a linear interaction between mechanical and electrical state changes. However, real solids always exhibit dissipation processes and, in some cases, can also display highly nonlinear behavior, hysteresis effects, instability, and a variety of electromagnetic effects. In linear piezoelectricity, the equations of linear elasticity are coupled via the piezoelectric constants to the charge equation of electrostatics. However, the electrical quantities are not purely static but, due to the coupling with the dynamic-mechanical equations, are only quasi-static. [3, 4].

In the linear description of piezoelectricity, so-called piezoelectric constants are used. They are listed in Table 1 along with their definitions according to [4]. The piezoelectric charge constant—also known as the piezoelectric strain constant d —which is important for the development of piezoelectric transducers, relates the electric charge generated per unit area to an applied mechanical force:

$$d = \frac{\text{strain}}{\text{electric field strength}} = \frac{\text{electric displacement}}{\text{pressure/stress}}$$

Its unit is meters per volt (equivalent to coulombs per newton). This constant is often used to evaluate the quality of a piezoelectric material. The constant d_{ij} can be related to three important material properties: k , the electromechanical coupling coefficient; k^T , the relative permittivity at a constant voltage; and s^E , the elastic modulus (unit: meters per Newton) at a constant electric field:

$$d_{ij} = k_{ij} \sqrt{\epsilon_0 k_i^T s_{ij}^E}$$

Two important constants are d_{33} and d_{31} . It is useful to keep in mind that large d -constants lead to large mechanical displacements, which are generally desired in SAW transducers. A process in which the force acts along the d_{33} direction (parallel to

the polarization axis) and acts on the same surface from which the charge is collected is described by d_{33} . If the charge is collected from the same surface as in d_{33} , but the force is applied at a right angle to the polarization axis, this behavior is determined by d_{31} . Much less common than the latter two constants is d_{15} , which describes the shear effect. The constant d is thus a tensor coupled to the electric field vector and the stress tensor. Its components d_{33} and d_{31} have the following empirical relationship [2, 5]:

$$d_{33} \approx -2.5d_{31}$$

2 TRANSFERFUNCTIONS OF SAW-FILTERS

The basic building block of an surface acoustic wave (SAW) filter is a piezoelectric substrate, typically made of lithium niobate or ST quartz, on whose surface controlled Rayleigh waves—a combination of longitudinal and transverse two-dimensional waves—propagate. The generation of such waves is effectively achieved through a clever arrangement of electrodes (which are relatively long in width). In their simplest form, these so-called interdigital transducers (IDTs) consist of two identical comb-like metal structures (electrodes), whose respective fingers are arranged on the surface in an interlaced pattern that influences one another. Fig. 3 shows such an IDT. The IDT period length p is the distance between the centers of two consecutive fingers of the same comb. The electrodes all have the same width a and the same spacing a , which is a common configuration. When an alternating voltage is applied to the IDT, the generated electric field is coupled into the piezoelectric substrate via acoustic waves that propagate in both directions along the axis perpendicular to the fingers.

In the ideal case of a reflection-free and lossless IDT, the damping frequency response of such a circuit can be calculated using the delta-function model, which dates back to Tancrell and Holland [6]. We assume here that the propagation of surface waves is not affected by the electrodes beneath which they travel. As a good approximation, there is no dispersion, so their velocity v is independent of frequency.

By representing the fingers of an IDT as delta functions (spaced p apart) in space, the signal (in time and space) that arrives at the point x_0 when a pulse is emitted simultaneously from all fingers (by applying the voltage $U_1 = \hat{u}_1 \delta(t)$) can be described by the

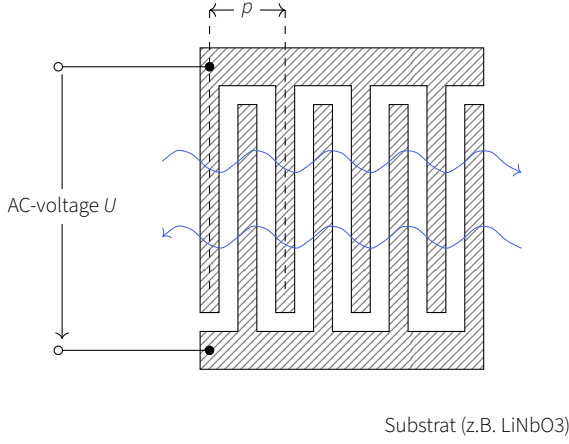


Figure 3 | Single IDT. A voltage U is applied between the lower and upper electrode combs. Both electrode combs are coupled through the piezoelectric substrate.

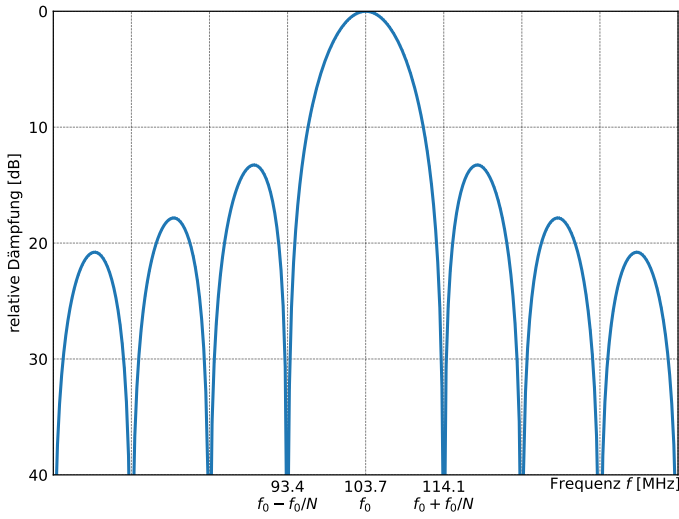


Figure 4 | Frequency dependence of a single IDT. The zeros occur at regular intervals from the center frequency. The number N of electrode fingers in this case is 10. A larger number of electrode fingers results in a narrower bandwidth.

following formula:

$$U_{\text{pulse}} = \sum_{n=0}^{N-1} a_n \delta\left(t - n \frac{p}{v}\right) = \sum_{n=0}^{N-1} a_n \delta\left(t - \frac{n}{2f_0}\right) \quad (1)$$

Here, N is the number of fingers, and $v/(2p) = f_0$ was used, where v is the propagation velocity of the surface waves. $a_n = (-1)^n \hat{a}$ represents the amplitude of the pulse, assuming that the voltages at the upper and lower crests are of the same magnitude but differ in sign.

It is known from signal theory that the impulse response of a filter is given by its transfer function $h(t)$, so that $U_{\text{impulse}} = h_1(t)$ holds. For simplicity, we set $\hat{a} = 1$. In fact, $\hat{a}$ depends on the coupling between the substrate and the electrodes. If the transient impulse response $h(t)$ is given, it can be calculated using a Fourier

transform to obtain the frequency-domain response $H(\omega)$:

$$\begin{aligned} H_1(\omega) &= \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\infty} h_1(t) e^{-i\omega t} dt \\ &= \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\infty} \sum_{n=0}^{N-1} a_n \delta\left(t - \frac{n}{2f_0}\right) \exp\{-i\omega t\} dt \\ &= \frac{1}{\sqrt{2\pi}} \sum_{n=0}^{N-1} (-1)^n \exp\left\{-i \frac{n\pi f}{f_0}\right\} \\ &= \frac{1}{\sqrt{2\pi}} \sum_{n=0}^{N-1} \exp\left\{-i\pi n \left(\frac{f}{f_0} - 1\right)\right\} \end{aligned} \quad (2)$$

Equation (2) is a geometric series, which yields:

$$\begin{aligned} H_1(f) &= \frac{1}{\sqrt{2\pi}} \sum_{n=0}^{N-1} \exp\left\{-i\pi \left(\frac{f}{f_0} - 1\right)n\right\} \\ &= \frac{1}{\sqrt{2\pi}} \frac{1 - \exp\left\{-i\pi N \left(\frac{f}{f_0} - 1\right)\right\}}{1 - \exp\left\{-i\pi \left(\frac{f}{f_0} - 1\right)\right\}} \end{aligned} \quad (3)$$

Equation (3) is a complex-valued function of f . We are primarily interested in the amplitude transfer function, which is given below:

$$\begin{aligned} |H_1(f)| &= \sqrt{H_1(f) H_1(f)^*} \\ &= \frac{1}{\sqrt{2\pi}} \left| \frac{\sin\left(\frac{N\pi}{2} \left(\frac{f}{f_0} - 1\right)\right)}{\sin\left(\frac{\pi}{2} \left(\frac{f}{f_0} - 1\right)\right)} \right| \end{aligned} \quad (4)$$

For frequencies f close to the center frequency f_0 , (4) the small-angle approximation ($\sin(x) \approx x$) can be applied. Extending this by N then yields a form of the sinc(x) function:

$$|H_1(f)| = \frac{N}{\sqrt{2\pi}} \left| \frac{\sin\left(\frac{N\pi}{2} \left(\frac{f}{f_0} - 1\right)\right)}{\frac{N\pi}{2} \left(\frac{f}{f_0} - 1\right)} \right| \quad (5)$$

When measuring electrical signals, the signal's power is usually measured instead of its amplitude. The power is proportional to the square of the amplitude. The value is normalized and expressed in dB yielding:

$$H_{\text{dB}}(f) \Rightarrow 10 \log(|H(f)|^2 / |H(f_0)|^2) \quad (6)$$

This transfer function of an IDT is plotted for representative parameters in Fig. 4.

The zeros of H_1 are the same as those of a standard $\sin(x)$ function, that is, at $x = n\pi$, with the exception of $x = 0$, which corresponds to $f = f_0$. At that point, $H_1(f)$ has its maximum. Therefore, the zeros as a function of f are given by:

$$f(n) = \frac{n2f_0}{N}, \quad n \in \mathbb{Z} \wedge n \neq 0 \quad (7)$$

The bandwidth, which extends from $f(-1)$ to $f(1)$, is therefore:

$$\Delta f = \frac{4f_0}{N} \quad (8)$$

Typical SAW circuits consist of two identical IDTs spaced d apart, as shown in Fig. 5. The bottom electrodes of both IDTs are

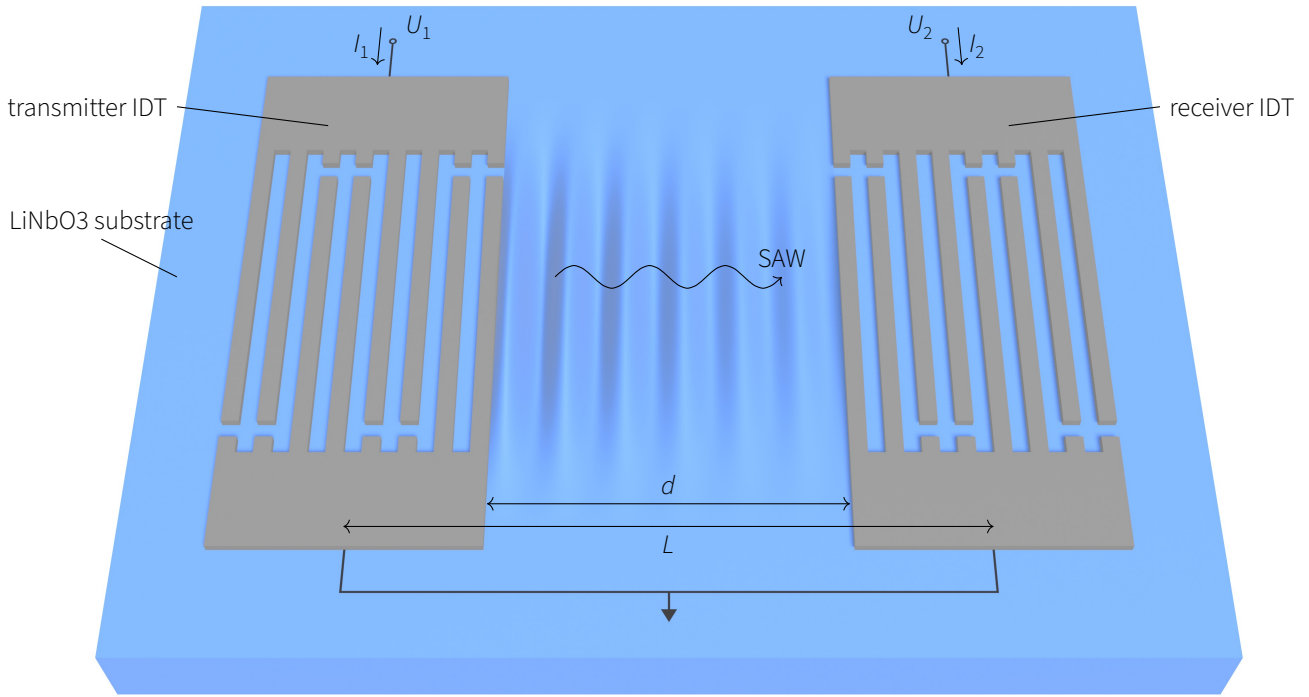


Figure 5 | 3-d Illustration of a delay-line type SAW-filter. The lower electrode comb sections of both IDTs are grounded. The upper electrode comb sections are connected to voltages U_1 and U_2 , respectively.

grounded, and both IDTs are coupled to each other through the piezoelectric substrate. In this configuration, one IDT acts as the transmitter and the other as the receiver.

The surface wave travels a distance along the substrate and arrives at the receiver IDT. In contrast to (1), where the impulse response was determined for a specific location x_0 , in the present case, the signals arriving at the individual fingers of the receiver IDT must be summed. Here, we assume that the distance between the fingers of the two IDTs for the same n is given by the length L (see Fig. 6). The impulse response can then be expressed by the following formula:

$$h_{\text{Filter}}(t) = \sum_{m=0}^{N-1} \sum_{n=0}^{N-1} a_n a_m \delta\left(t - \frac{(np + L + mp)}{v}\right) \quad (9)$$

A Fourier transform of (9), where $a_n = a_m$ and $\hat{a}_1 = \hat{a}_2 = 1$, yields:

$$\begin{aligned} H_{\text{Filter}}(\omega) &= \\ &= \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\infty} h_{\text{Filter}}(t) e^{-i\omega t} dt \\ &= \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\infty} \sum_{m=0}^{N-1} \sum_{n=0}^{N-1} a_n a_m \delta\left(t - \frac{(np + L + mp)}{v}\right) \exp\{-i\omega t\} dt \\ &= \frac{1}{\sqrt{2\pi}} \sum_{m=0}^{N-1} \sum_{n=0}^{N-1} (-1)^n \exp\left\{-i\omega \frac{np}{v}\right\} \exp\left\{-i\omega \frac{L}{v}\right\} (-1)^n \exp\left\{-i\omega \frac{mp}{v}\right\} \\ &\dots \\ &= \frac{1}{\sqrt{2\pi}} H_1(\omega) \exp\left\{-i\omega \frac{L}{v}\right\} H_1(\omega) \end{aligned} \quad (10)$$

From (10), it can be seen that the propagation path causes a phase shift of $\varphi = \omega L/v$ in the received signal, where $\tau = L/v$. As expected, this transfer function is the product of the transfer functions of the individual components, which would correspond to a

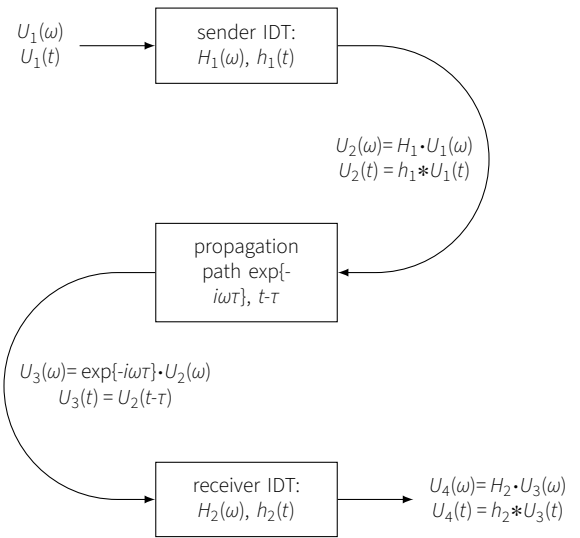


Figure 6 | Division of the entire filter into individual elements. In the frequency domain, the transfer function is the product of the individual subcomponents.

convolution in the time domain. Fig. 6 schematically illustrates the filtering of the signal and in Fig. 7 this transfer function is plotted for representative parameters.

In real-world applications, the SAWs generated on the substrate always have a component that propagates into the substrate (volume waves) and also propagate outside the IDT. Consequently, the electrical energy coupled into the substrate can never be fully recovered. These losses are called insertion loss. They can be minimized using various methods, such as reflection grids on both sides of the IDT.

The configuration of IDTs shown in Fig. 5 is commonly referred to as a delay line. Because the propagation speed of surface acoustic waves is significantly slower than that of electromagnetic waves

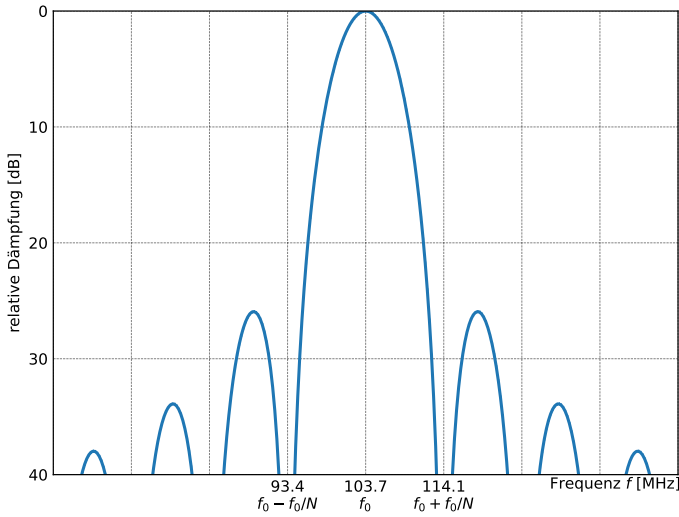


Figure 7 | Frequency dependence of the circuit's attenuation in Fig. 5. The zeros are the same as in Fig. 4, and the attenuation in the passbands is more pronounced at multiples of the fundamental frequency.

(EMWs), such circuits can delay an electrical signal by up to several tens of μs , which corresponds to an EMW propagation distance of several kilometers. A particularly undesirable effect that must be taken into account with this type of circuit is the so-called triple transit echo (TTE). At the receiving transducer, a portion of the incoming wave is reflected back to the transmitting transducer. From there, another portion is reflected back, so that these reflected components arrive at the receiving transducer after three times the propagation time.

SFW filters can be used as high-frequency reference elements in filters and oscillators. They are among the most important components in all smartphones. With 5G, the sub-6 GHz spectrum was introduced for early deployments; SAW filters can cover most bands in the lower range. Filter solutions based on SAWs provide modern mobile devices with high-performance duplexers and multiplexers for RF communication modules.

On the other hand, SAW filters can also function as sensors, in which case responses resulting from disturbances in the wave propagation properties—in particular, wave velocity and attenuation—arising from the interaction between the SAW and a surface layer are used as measures. Temperature, strain, pressure, force, and the properties of the added surface materials are examples of measured quantities. [7].

3 OPERATION PRINCIPLE OF SPLIT FINGERS

When designing SAW filters, the goal is to minimize insertion loss and unwanted interference effects, thereby maximizing performance and stability. In addition to using piezoelectric substrates with high electromechanical coupling constants, the choice of IDT type is of crucial importance. A key objective is to suppress unwanted reflections caused by so-called acoustic impedance mismatches, which significantly affect the filter's performance. This can be achieved, for example, through an angled IDT geometry and the use of double electrodes, known as split fingers. Fig. 8 illustrates the operating principle of so-called split fingers.

As shown in the figure (left), in the simplest version of the IDT, the finger widths and the spacing between them are $\lambda_0/4$, where

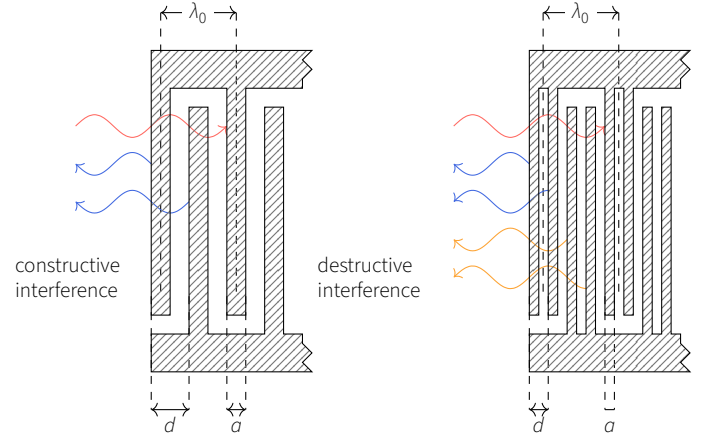


Figure 8 | Wave interference of single and double electrodes during operation at the fundamental frequency. Left: The phase difference is λ , so reflected wave pairs interfere constructively. Right: Wave pairs with a phase difference of $\lambda/2$ interfere destructively.

λ_0 is the wavelength of the fundamental mode. In contrast, an IDT with split fingers (right) for the same fundamental mode consists of a lattice structure with intervals of $\lambda_0/8$. This structure reduces the Bragg reflection, which acts opposite to the direction of propagation of the SAW and is the main cause of increased insertion loss. The Bragg wavelength λ_b , at which Bragg reflection occurs for SAWs, is given below in terms of d_{IDT} , the lattice period, and n , the order of reflection:

$$n\lambda_b = 2d_{\text{IDT}} \quad (11)$$

For an IDT with a finger width of $a = \lambda_0/4$, $\lambda/2 = d_{\text{IDT}}$, which is why the Bragg wavelength λ_b is identical to the wavelength of the fundamental frequency and thus causes high insertion loss. Bragg reflection in an IDT with split fingers occurs at twice the center frequency, since $a = \lambda_0/8$ and $\lambda_0/4 = d_{\text{IDT}}$. In this case, the reflected waves interfere maximally destructively, which is why the insertion loss at the filter's fundamental frequency is significantly lower in this IDT design. Since the triple transit echo (TTE), which arises from reflection between the transmitter and receiver IDTs, is also a result of Bragg reflection, it has the same Bragg frequency. Therefore, split fingers also minimize the TTE [8].

For filters with split fingers in which both IDTs are identical and spaced d apart, L can be determined as follows:

$$L(d, N) = a(2N - 1) + d \quad (12)$$

It is assumed here that the finger width a is equal to the distance between them.

4 MAXIMUM POSSIBLE NUMBER OF DIES PER WAFER

In the manufacture of integrated circuits, the die-per-wafer yield plays a very important role. Quite apart from production-related defects, the first step is to maximize the number of dies that can be produced from a single wafer. Algorithms have been developed to determine this number precisely, many of which are based on the Pythagorean theorem. In this process, a grid of chips is superimposed on the wafer, and the radial distance from the center of the wafer to the outermost corner of each potential chip location is calculated. Consequently, only those chips for which the radial

```

import numpy as np

def count_inside(w,h,R):
    xy0s = [(0,0),(0,h/2),(w/2,0),(w/2,h/2)]
    count = 0
    final_count = []
    for (x0,y0) in xy0s:
        count = 0
        x_all = np.concatenate((np.arange(-x0-w,-R-w,-w),np.arange(-x0+w,R+w,w)))
        y_all = np.concatenate((np.arange(-y0-h,-R-h,-h),np.arange(-y0+h,R+h,h)))
        for x in x_all:
            for y in y_all:
                if ((x**2 + y**2) <= R**2):
                    count = count + 1
        final_count.append(count)
    return np.amax(np.array(final_count))

```

Figure 9 | Python algorithm for calculating the number of dies on a wafer. The maximum number is determined from four different positions (x0, y0) in the die grid.

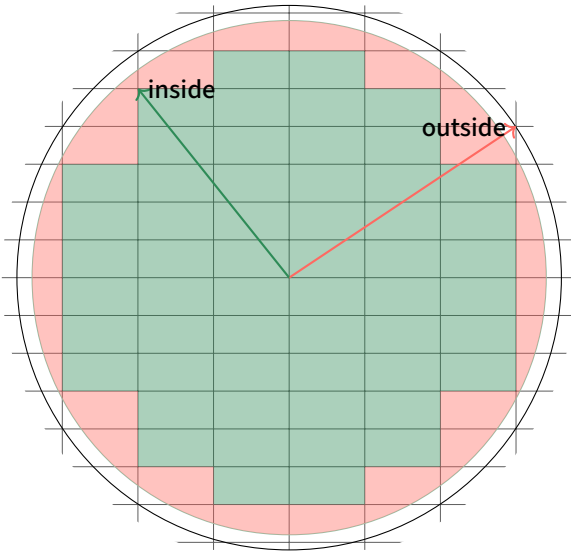


Figure 10 | Sketch of the algorithm for calculating the number of dies on a wafer.

distance is less than the effective wafer radius are included in the count. Fig. 10 illustrates this process. In the early stages of chip development, quick estimates are often sufficient, which is why one would prefer to avoid these somewhat complex numerical iterative algorithms. A simple approximation of the number of dies with a specific aspect ratio that fit on a wafer can be calculated using so-called GDW formulas (gross die per wafer). An often-cited formula goes back to Trapp [9] [10].

$$N_{GDW} = \frac{\pi}{A_D} (R_{eff} - \sqrt{A_D})^2 = \left(\frac{\pi R_{eff}^2}{A_D} \right) \left(1 - \frac{\sqrt{A_D}}{R_{eff}} \right)^2 \quad (13)$$

This formula can be interpreted as first dividing the wafer area by the chip area. Due to the roundness of the wafer and the fact that the chips must be placed on a regular grid, this number must be adjusted downward. The correction term can be illustrated by dividing the effective circumference of the wafer by the length of the diagonal of a square die of the same area. This value is thus approximately the number of dies intersected by the effective circumference. Despite the fact that this formula is very intuitive and easy to understand, it fails as die size increases. Dirk K. de Vries [11] therefore proposes a correction factor, as well as a modifica-

tion that accounts for dies with different aspect ratios, allowing the GDW value to be determined with greater accuracy:

$$N_{GDW} = \frac{\pi}{A_D} (R_{eff} - 0.29(H + W))^2 \quad (14)$$

Here, H is the height and W is the width of a die.

As an example, we compute N_{GDW} for equally sized dies of dimensions of $6405 \mu\text{m} \times 3005 \mu\text{m}$, manufactured on a 100-mm wafer, which is to be left with a 5-mm margin free of structures (therefore has an effective radius of 45 mm). Using (14), the number of dies that fit on the wafer is calculated to be $291.6 \approx 291$. To verify the accuracy of this result, we can calculate the exact number using an algorithm as shown in Fig. 10. Fig. 9 shows the Python code for this. This yields a value of 293. Using (14), the GDW can therefore be determined very accurately in a very simple manner.

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