OPTICAL CHARGE INJECTION INTO A GALLIUM ARSENIDE
ACOUSTIC CHARGE TRANSPORT DEVICE

By

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ABSTRACT

There is a need for monolithic devices capable of spatial resolution in imaging and ionizing radiation detection. In this thesis, a GaAs acoustic charge transport device (ACT) was studied for this purpose. A new method of charge injection has been demonstrated for the ACT. Using near-infrared optical pulses incident through thin semi-transparent chromium windows, electron-hole-pairs were separated by the electric field in a depleted n-type channel region of the device. For light penetration less than the depth of the electron potential minimum, and for small injection levels, calculations indicated that electrons and holes were separated at their saturation velocities. Holes moving toward the surface of the substrate could recombine with electrons at an evaporated Schottky metal plate. Electrons moving toward the channel centre were bunched and transported by the electric field coupled to a $\langle 110 \rangle$ propagating surface acoustic wave (SAW) on (100) cut GaAs. Quantum efficiency, defined as the number of electrons collected at the output per incident photon on the GaAs surface, was greater than 9% at an optical wavelength of 730 nm. When compensation was made for the loss and reflection due to the chromium windows, the quantum efficiency was in excess of 24%. Charge transfer efficiency was greater than 0.992 with the ACT clocked at 360 MHz. The demonstrated optical injection technique may be of use in future ACT imaging devices.

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LIST OF SYMBOLS

a_k	-	eigenvector component (Appendix A)
A	-	partial wave matrix (Appendix A)
b	-	SAW beamwidth
b_k	-	eigenvector component (Appendix A)
c	-	speed of light in vacuum
c_{11}, c_{12}, c_{44}	-	stiffness constants (abbreviated subscripts)
c_{ijkl}	-	stiffness constants (full tensor form)
C_s	-	low frequency capacitance of a single IDT finger pair per unit SAW beamwidth
C_T	-	low frequency IDT capacitance
d	-	thickness of chromium window layer
D	-	boundary conditions matrix (Appendix A)
D_i	-	electric displacement
e_{14}, e_{25}, e_{36}	-	piezoelectric constants (abbreviated subscripts)
e_{ijk}	-	piezoelectric constants (full tensor form)
E, E_k	-	electric field
E_c	-	conduction band energy level
E_{cp}	-	electric field in ACT channel
E_{fm}	-	metal Fermi level
E_s	-	synchronous electric field
E_v	-	valence band energy level
f	-	frequency variable

LIST OF SYMBOLS (continued)

f_0	-	centre frequency of IDT
f_{BURST}	-	RF burst frequency applied to ACT input
g	-	electron-hole-pair generation rate
g_k	-	generalized immittances used in calculation of optical loss through chromium film
$G(f)$	-	frequency response of ACT to RF burst signals
h	-	Planck's constant
$H(f)$	-	frequency response of SAW delay line
I_{cp}	-	channel plate current
I_{D}	-	drain current
I_{OUT}	-	ACT output ohmic current
J_{AE}	-	acoustoelectric current density (A/m)
k	-	acoustic wavenumber
K_{S}^2	-	SAW electromechanical coupling coefficient
l_{c}	-	transfer length (ohmic contact)
L	-	ohmic contact test structure gap width (Figure 3.8)
L_{Cr}	-	fraction of optical power lost in chromium film
L_{x}	-	gap-width axis intercept of ohmic contact resistance graph (Figure 3.8)
n_0	-	optical EHP density at device surface
N	-	number of IDT finger pairs
$N_{\text{A}}, N_{\text{D}}$	-	doping densities
N_{e}	-	number of electrons per centimeter channel width

LIST OF SYMBOLS (continued)

N_k	-	optical constants
N_t	-	number of equivalent charge transfers
P_{AI}	-	acoustic intensity (Watts per λ_s beamwidth)
P_{CONV}	-	power lost in conversion between electrical and acoustic energy
P_{IN}	-	electrical power applied to IDT
P_{OPT}	-	optical power incident on ACT
P_{REF}	-	electrical power reflected from IDT
P_{SAW+}	-	component of SAW power in direction of ACT channel
P_{SAW-}	-	component of SAW power in direction opposite to ACT channel
q	-	electronic charge
Q_L/Q_T	-	fraction of charge lost from leading edge of the ACT output current pulse
r_2	-	reflection coefficient for air-chromium interface
r_c	-	ohmic contact resistivity
R	-	reflectance of air-chromium interface
R_c	-	resistance derived from ohmic contact graph (Figure 3.8)
R_{in}	-	real part of the ACT input impedance
RL	-	return loss (in decibels)
R_s	-	sheet resistance derived from ohmic contact graph (Figure 3.8)

LIST OF SYMBOLS (continued)

S	-	acoustic power density (W/m)
S_0	-	initial acoustic power density at $x_1=0$
S_{ij}	-	strain tensor
t	-	time variable
t_2	-	transmission coefficient for chromium-GaAs interface
t_r	-	rise time of ACT output current pulse used in calculation of charge transfer efficiency
T	-	transmittance of chromium-GaAs interface
T_{ij}	-	stress tensor
u_1, u_3	-	sagittal plane displacement components
u_2	-	displacement normal to sagittal plane
v_i	-	components of particle velocity
v_s	-	SAW phase velocity
v_{SAT}	-	carrier saturation velocity for GaAs
v_{so}	-	SAW velocity with unmetallized surface
v_{ss}	-	SAW velocity with metallized surface
V_{SR}	-	source voltage
V_D	-	drain voltage
V_{DS}	-	drain-source voltage
V_G	-	gate voltage
V_{IDT}	-	IDT bias voltage
V_m	-	Schottky channel plate voltage
V_s	-	applied substrate voltage
V'_s	-	effective substrate voltage
W	-	total depletion layer thickness

LIST OF SYMBOLS (continued)

W_n	-	geometric thickness of n-epitaxial layer
W_p	-	width of p-type region at substrate/epitaxial layer junction
W_T	-	width of gap in ohmic contact test structure (Figure 3.8)
W_k	-	weighting factors (Appendix A)
x_1, x_2, x_3	-	spatial coordinates
x_{3m}	-	position of electron potential minimum in channel
Y_{dB}	-	magnitude of IDT amplitude response relative to centre frequency value
Z_{A0}	-	acoustic characteristic impedance
$Z_{in}(f_0)$	-	IDT centre-frequency impedance
α	-	acoustoelectric attenuation coefficient
α_0	-	optical absorption coefficient
β	-	anisotropy parameter used in diffraction theory
γ_k	-	angles as defined in Figure B.1
δZ_A	-	change in acoustic characteristic impedance from a free to a mass-loaded surface
ϵ	-	dielectric constant
ϵ_0	-	permittivity of free space
ϵ_{ik}	-	permittivity (tensor form)
ϵ_R	-	relative permittivity
ϵ_t	-	charge transfer inefficiency
η	-	quantum efficiency

LIST OF SYMBOLS (continued)

θ	-	propagation angle relative to $\langle 110 \rangle$ on (100) surface
λ_p	-	wavelength of incident light pulse
λ_s	-	SAW wavelength
μ	-	carrier mobility
ν	-	optical frequency
ρ_l	-	phase constant of chromium film
ρ_D	-	space-charge density due to ionized donors in ACT channel
ρ_M	-	mass density
ρ_W	-	piezoelectric space-charge density
σ, σ_k	-	decay constant (Appendix A)
ϕ	-	beam-steering angle
ϕ_c	-	channel potential components with spatial dependence
ϕ_d	-	depletion potential in ACT channel
ϕ_{dm}	-	electron potential minimum in channel
ϕ_{ml}	-	Schottky metal work function
ϕ_o, ϕ_w	-	SAW quasistatic potential
ϕ_p	-	GaAs work function
ϕ_s	-	plane wave form of SAW wave potential
ω	-	radian frequency of SAW

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LIST OF ACRONYMS

ACT	-	acoustic charge transport (device)
CCD	-	charge-coupled device
CGCCD	-	capacitive-gate charge-coupled device
CTE	-	charge transfer efficiency
CTI	-	charge transfer inefficiency
CW	-	continuous wave
EHP	-	electron-hole-pair
IDT	-	interdigital transducer
ITO	-	indium-tin oxide
LED	-	light-emitting diode
LSAW	-	leaky surface acoustic wave
MESFET	-	metal-semiconductor field-effect transistor
MBE	-	molecular-beam epitaxy
MOCVD	-	metal-organic chemical vapour deposition
PIN	-	p-type/intrinsic/n-type (diode)
RF	-	radio frequency
RGCCD	-	resistive-gate charge-coupled device
SAW	-	surface acoustic wave
SEM	-	scanning electron microscope
SI	-	semi-insulating
TE	-	transverse electric
TM	-	transverse magnetic
UHF	-	ultra-high frequency
VHF	-	very-high frequency
VPE	-	vapour-phase epitaxy

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1.0 INTRODUCTION

This thesis describes electron injection into a GaAs acoustic charge transport (ACT) device using near-infrared optical pulses ($\lambda_p=730$ nm). Frontside illumination was achieved using a focussed light source incident within the surface acoustic wave (SAW) beamwidth through semi-transparent metallic windows. It will be shown that the electron-hole pairs (EHP's) created by the optical pulse were quickly separated by a strong electrostatic field in the depleted n-epitaxial channel of the ACT. The holes drifted toward an evaporated Schottky contact on the surface where they could recombine with electrons in the metal. Electrons moved toward the channel centre where they were bunched and transported by the $\langle 110 \rangle$ propagating SAW on (100) cut GaAs.

The optical injection experiments were motivated by a need to spatially resolve radiation patterns. For an ACT, the fundamental resolution limit is determined by the SAW wavelength: At 360 MHz the wavelength is 8 μm . As the drive frequency increases, the wavelength-limited resolution of the ACT improves. An imaging device based on charge-coupled device (CCD) technology has spatial resolution determined by the gate dimensions and number of phases. For a two-phase CCD with 5 μm gate widths and 1 μm intergate gaps, the best resolution would be 12 μm , independent of clock frequency.

In a radiation detection environment spatial resolution is only one measure of performance. For an optical imaging device, parameters such as quantum efficiency, spectral

response, linearity, response time, and responsivity are important for many applications. With the exception of spectral response, these parameters have been considered in this thesis for an experimental ACT device.

As a nuclear particle detector, device degradation and eventual destruction caused by radiation induced substrate damage must be assessed. A recent study of GaAs CCD devices indicated an increased charge transfer inefficiency and leakage current due to radiation induced traps resulting from total doses of $\approx 10^6$ rad. [1]. One might expect similar results to apply to ACT devices given a comparable substrate configuration.

Transient effects have been reported for GaAs MESFET's exposed to low-level (≈ 100 rad.) ionizing radiation [2]. Noted was a sudden drop in MESFET drain-source current immediately following the radiation pulse followed by a very slow recovery. The response was attributed to electron trapping in the semi-insulating (SI) substrate for devices built on epitaxial material. The trap lifetimes were several seconds at room temperature, resulting in a slow release of electrons. Similar transient phenomena have been observed in GaAs MESFET photodetectors [3]. These effects could impact on ACT performance for optical or ionizing radiation penetration depths extending into the SI substrate. The use of a buried p-layer to shield substrate trapping effects from the n-type transport channel has been proposed to reduce this effect [4].

Optical charge injection has been reported in the past for GaAs CCD's [5-6]. Two separate approaches have been proposed depending upon whether the light was incident on, or opposite, the gating side of the device. A backside illuminated structure reported by Liu, Anderson, Deyhimy, and Milano [5] is shown in Figure 1.1a. A p-type AlGaAs "window" was grown on a transparent substrate to passivate the surface states of a p^+ -GaAs absorber layer and provide a long wavelength cut-off for the incident light. Electrons generated in the p^+ -GaAs diffused to the junction of the wide bandgap n-AlGaAs layer and were collected in a depletion region charge well, created by appropriately biased metal electrodes on the surface.

A frontside illuminated device described by Sahai, Pierson, Anderson, Martin, Sovero, and Higgons [6] is shown in Figure 1.1b. In this simpler structure, a semi-transparent indium-tin-oxide (ITO) film was sputtered on the surface of the device to form the gating electrode array. Light incident through the ITO electrodes created EHP's in the channel of the device. The EHP's were separated by the depletion field under the gate structure.

The devices described by Liu et al. [5] and Sahai et al. [6] used AlGaAs and GaAs active transport regions, respectively. The use of wide bandgap AlGaAs is preferred for reduction of dark current. To date, no ACT devices have been reported that use an AlGaAs transport channel. Very few studies have been concerned with the SAW properties of AlGaAs

Figure 1.1a: A backside illuminated AlGaAs/GaAs CCD imager

Figure 1.1b: A frontside illuminated GaAs CCD imager

with some results reported by Adachi [7] and Sapriel, Michel, Toledano, Vacher, Kervarec, and Regreny [8]. As a result, ACT devices on AlGaAs heterojunction structures remain an area requiring future research. Devices built for this thesis have been based on n-GaAs transport channels.

The ACT device concept first reported in 1982 by Hoskins, Morkoc, and Hunsinger [9] utilized both the semiconducting and piezoelectric nature of GaAs. In its simplest form, the ACT device operated as an analog delay line. Electrons injected into the ACT were bunched and transported as discrete charge packets at the speed of sound by the electric field coupled to a propagating surface acoustic wave. Physically, the ACT device was considered as a GaAs MESFET with a very long gate in tandem with a SAW interdigital transducer. The MESFET portion of the device was biased to depletion, providing an electric field distribution that tended to confine electrons to the centre of the channel. The mechanism responsible for the bunching and transport of electrons at sonic velocity was the large signal acoustoelectric effect [10, 11].

The idea of transporting charge by means of the electric field associated with a surface acoustic wave in a piezoelectric material is not new. Work in the past concentrated on devices built with silicon, a non-piezoelectric semiconductor, using a thin overlay of piezoelectric material such as ZnO in a ZnO:SiO₂:Si configuration [12-15]. Transduction of the SAW took place in

the ZnO and the electric field was directly coupled into the Si, where interactions with minority charge carriers occurred. Another method involved placing a piezoelectric material, such as LiNbO_3 , in close proximity to a silicon substrate so that the external SAW electric field generated in the LiNbO_3 interacted with charge carriers in the Si through an air gap [16, 17]. Fabrication difficulties and performance problems reduce the appeal of these devices. Three such limitations on charge transfer efficiency and device length are [15]: surface charge trapping effects; the difficulty of obtaining a uniform and reproducible air gap in the LiNbO_3 device and the attenuation of the SAW in sputtered ZnO films.

Recently, GaAs charge coupled devices have been developed [4, 18-26]. In the originally proposed capacitive gate CCD structure (CGCCD), charge was carried in wells defined by electrically clocking a multiphase array of Schottky gates on the surface of the device. Good charge transfer efficiency was achieved by keeping the gaps between individual gates very small [20]. Generation of the multiphase clocking waveforms and subsequent filtering of clock interference at the output are substantial technical problems, especially at high clock frequencies.

The emerging resistive-gate GaAs CCD technology (RGCCD) [25] reduced the photolithography problems associated with the originally proposed capacitive gate GaAs CCD while providing stronger fringing fields at reduced transport

depths. This allowed epitaxial layer thickness to decrease to a value compatible with other GaAs active devices such as microwave MESFET's, a significant advantage. Unfortunately, multiphase clocking circuitry was still required. Increased leakage currents due to the non-ideal Schottky behaviour of the "Cermet" resistive gates [25] was one disadvantage of the RGCCD.

The ACT device provides an elegant solution to several CCD device shortcomings. In particular, the ACT drive signal is an easily-generated single-phase sinusoid. Effective decoupling of the clock and input signals is achieved since the SAW transducer is remote from the charge transport channel, not directly integrated with the channel as in a CCD. The clock waveform can be eliminated at the output of the ACT with a narrowband notch filter.

ACT charge transport occurs continuously in time and space at the sound velocity under the influence of a constant electric field. Electrons follow straight line paths from the device input to output. The slower process of diffusion is not involved. As a result, ACT charge transfer efficiency (CTE) should be superior to that achieved with GaAs CCD devices, particularly at high frequencies. This seems to be the case, with CTE in excess of 0.99994 reported in the literature for an ACT device operated at 360 MHz [27].

Some applications of CCD technology, such as time expansion and compression of pulses cannot effectively be performed by the ACT. This is because the drive frequency is

fixed with the latter device by the narrowband interdigital transducer (IDT). Since GaAs is weakly piezoelectric, efficient generation of a large amplitude SAW requires a long, and hence narrowband, IDT. Typically 100 finger pairs with fine dimensions are necessary. Finger widths of 2 μm or 1 μm are used depending upon the type of transducer design, at 360 MHz. Higher frequencies necessitate either smaller dimensions or the launching of SAW on a harmonic of a lower frequency IDT.

Device performance is critically dependent on the surface morphology of the GaAs substrate. Attenuation of high frequency SAW on GaAs is sensitive to topological irregularities, as may be produced during a polishing or epitaxial layer growth process [28]. The SAW attenuation characteristics have limited the length of reported devices to several millimeters at 360 MHz. The latter limitation is likely to improve as GaAs epitaxial wafer growth techniques advance.

The first ACT devices were delay lines. A signal was electrically injected through an input diode and was destructively sensed by an output diode. Since electron transport occurred near the surface of the substrate, an array of electrodes placed along the propagation direction could sense the passage of injected charge by means of image currents [29, 30]. By varying the apodization of these sense electrodes and summing their output, signal processing devices such as transversal filters have been formed. At

present, this type of analog matched filter operation has been achieved predominantly with CCD or SAW devices.

A first step toward extension of ACT device concepts into the areas of optical or ionizing radiation imaging is reported in this work. By studying the optical charge injection characteristics of the device using near infrared pulses, experiments have indicated a charge transfer efficiency in excess of 0.992 and net quantum efficiency greater than 9%. When compensation was made for the optical signal loss through the chromium windows, a quantum efficiency greater than 24% was calculated. The results to date should provide impetus for future research aimed at applying ACT technology in an actual imaging device design.

This thesis contains five chapters. Chapter 2 deals with theoretical considerations of ACT design and operation. Chapter 3 discusses the experimental device, outlines the fabrication process, and describes test structures added to the mask set for process control. Problems which had to be overcome while fabricating these devices are also discussed. Chapter 4 provides measured performance data including the optical charge injection results. Finally, Chapter 5 provides a summary of experimental measurements, a discussion of the device performance and offers some possible areas for future study.

2.0 THEORY

2.1 Characteristics of <110> Propagating SAW on (100) cut GaAs

The properties of SAW on GaAs have been extensively studied [31-39]. This research has determined that the <110> propagating SAW on (100) cut GaAs has the highest electromechanical coupling coefficient ($K_S^2 \approx 6.4 \times 10^{-4}$) defined by [40]:

$$K_S^2 = 2(v_{ss} - v_{so})/v_{so} \quad (2.1)$$

v_{ss} = SAW velocity with metallized surface

v_{so} = SAW velocity with unmetallized surface

Fortunately, the (100) wafer orientation possessing the highest value of K_S^2 is commonly used for fabrication of present GaAs active devices.

The electromechanical coupling coefficient is important in the design of any SAW device since it relates transduction efficiency and bandwidth. High efficiency and large bandwidth cannot be simultaneously achieved with weak piezoelectrics such as GaAs using an interdigital transducer.

A long, and hence narrowband, IDT is required to launch a SAW on GaAs with sufficient wave potential to drive the ACT charge transport process. However, in contrast with most SAW devices, the ACT bandwidth is not determined by the acoustic transducer. The charge injection/detection technique and finite charge transfer efficiency limit overall bandwidth to a value with its upper bound predicted by the Nyquist sampling theorem.

In devices where greater coupling than that afforded by GaAs alone is required, for example in wide bandwidth SAW filters, it is possible to deposit a piezoelectric ZnO layered structure over the transducer to improve efficiency [41]. Fortunately, the design of the ACT decouples the IDT bandwidth from the signal processing bandwidth. As such, a narrowband transducer design suffices.

The $\langle 110 \rangle$ propagating SAW on (100) cut GaAs is sometimes referred to as a pseudo-surface acoustic wave or the limit of a leaky surface acoustic wave (LSAW). This is because the surface wave has a higher velocity than a bulk shear wave whose displacement is perpendicular to the sagittal plane, ≈ 2860 m/s for the former compared to ≈ 2468 m/s for the latter [38]. The sagittal plane is defined by the direction of propagation (phase velocity) and the substrate surface normal. A leaky surface acoustic wave has finite energy coupling between itself and a bulk wave. In fact, it has been shown [32] that as the wave propagation direction varies from 0 to 45 degrees from $\langle 100 \rangle$ towards $\langle 110 \rangle$ on (100) cut GaAs, the surface wave starts out as a pure Rayleigh mode with no coupling to the electric field, becomes coupled to the electric field and a shear bulk mode, and finally ends up coupled to the electric field but decoupled from the bulk shear mode along $\langle 110 \rangle$. Hence, for propagation exactly along $\langle 110 \rangle$, the wave is not "leaky" and behaviour is similar to a "normal" Rayleigh mode.

Substrate orientation, particle motion in the sagittal

plane, and definition of axes used throughout this work are as given in Figure 2.1. Material "particles" trace out retrograde elliptical paths in the sagittal plane for propagation in the direction indicated. The magnitudes of the displacement decay rapidly into the substrate with most of the energy confined to within approximately one acoustic wavelength of the surface.

Building devices whose SAW propagation is aligned with $\langle 110 \rangle$ is aided by the natural GaAs cleavage planes along these directions. The loss due to coupling between the SAW and bulk shear mode is a strong function of propagation direction as shown in Figure 2.2 [36]. For this reason, and beam steering considerations to be discussed later, precise orientation along $\langle 110 \rangle$ is highly desirable.

An important characteristic of the $\langle 110 \rangle$ SAW is the variation of wave potential as a function of distance into the substrate. SAW displacement components and quasistatic potential magnitudes have been calculated in the present work using the method developed by Tseng and White [42] and Campbell and Jones [43]. Details of the calculation are given in Appendix A. An alternative method of solution has recently been presented by Datta and Hunsinger [44].

The calculations were made assuming no free carriers in the substrate. The surface of the crystal was assumed to be electrically shorted by a massless, perfectly conducting sheet. The material constants which were used are given in

U_i = Displacement Components
 $i = 1, 2, 3$

Rayleigh Wave Displacement

Figure 2.1: Definition of axes and particle displacements
 for $\langle 110 \rangle$ propagating SAW on (100) cut GaAs

Figure 2.2: SAW attenuation due to coupling to bulk shear wave near the $\langle 110 \rangle$ propagation direction

Table 2.1 [36]:

Table 2.1: GaAs material constants [36]

Constant	Type	Value	Units
Stiffness	c_{11}	11.9×10^{10}	N/m^2
	c_{12}	5.38×10^{10}	N/m^2
	c_{44}	5.95×10^{10}	N/m^2
Piezoelectric	e_{14}	-0.16	C/m^2
	e_{25}	-0.16	C/m^2
	e_{36}	-0.16	C/m^2
Relative Permittivity	ϵ_R	11.0	-
Mass Density	ρ_M	5.3×10^3	kg/m^3

The results of numerical calculations are shown in Figures 2.3a and 2.3b normalized to an acoustic intensity of 1 mW per wavelength beamwidth at 360 MHz. The SAW velocity calculated was 2865.6 m/s. Values in the literature range from 2856 m/s [38] to 2866 m/s [36] for an electrically unshorted surface. Shorting the surface lowers the SAW velocity by $\approx 0.03\%$. Calculating the SAW velocity for both a free and shorted surface allows the electromechanical coupling coefficient to be derived using equation (2.1).

The disagreements appearing in the literature regarding the calculated SAW velocity are presumably due to the use of slightly different GaAs material constants in the analysis. For example, when material constants recently given by Webster and Carr [38] were used, a SAW velocity of 2854.4 m/s

Figure 2.3a: SAW displacement components

Figure 2.3b: SAW quasistatic potential

was found for an electrically shorted surface.

The calculated quasistatic potential magnitude (ϕ_0) using the constants in Table 2.1 is $\approx 6\%$ higher than the curve given by Hoskins [45] page 8. When the material constants of Webster and Carr [38] were used, the quasistatic potential calculated was $\approx 16\%$ lower than the curve previously calculated using the material constants given in Table 2.1. The displacement component magnitudes were virtually identical for both calculations. The stiffness, piezoelectric and mass density given by Webster and Carr do not differ greatly from the values used by other authors [35-37]. However, the value of relative permittivity used by the former was 13.1 while the latter authors quote 11.0; a difference of $\approx 16\%$. It is therefore not surprising that the quasistatic potential values differ by approximately this amount. To the present author's knowledge, the appropriate value of ϵ_R to be used in the calculation has not been established. In keeping with the majority of publications describing SAW on GaAs, the relative permittivity of 11.0 was assumed for calculations presented in this thesis.

To use the curves shown in Figure 2.3 for other acoustic intensities, denormalization based on the proportionality below must be applied:

$$|\phi_0|, |U_1|, |U_3| \propto \sqrt{P_{AI}/2\pi v_S} \quad (2.2)$$

where: P_{AI} = Acoustic intensity in Watts/ λ_S beamwidth
 v_S = SAW velocity

An important property of the SAW potential curve shown

in Figure 2.3b, is the peak in wave potential that occurs $\approx 0.45 \lambda$ below the substrate surface. Using a clock frequency of 360 MHz, maximum wave potential is achieved at $\approx 3.6 \mu\text{m}$ into the substrate. Section 2.2 discusses the method used to confine transported charge so that the depth of transport coincides with the depth of greatest wave potential.

Consideration of the SAW potential profile into the substrate supports the choice of a 360 MHz drive frequency. Operation at much lower clock rates is impractical as the optimum depth of charge transport moves further from the device surface. This reduces the bandwidth of surface electrodes designed to sense the charge carried by the SAW. Epitaxial growth of GaAs layers more than several microns in thickness with good surface morphology and depletion of such layers create additional complications. Choice of higher clock frequency, while it alleviates these difficulties, increases the problems associated with SAW transduction since the finger widths of quarter and eighth wavelength transducers quickly approach limits achievable using standard optical patterning techniques.

Careful consideration must be given to acoustic diffraction and beam steering in SAW device design. For the ACT, these effects place fundamental limits on the practical channel length due to the spreading and steering of the SAW beam profile. Szabo and Slobodnik [46] have reviewed SAW diffraction theory. For the case of an anisotropic material whose phase velocity dependence on angle about a pure mode

axis is parabolic, simple relationships and curves can be used to calculate the amount of diffraction and beam steering. The almost parabolic phase velocity relationship with angle near $\langle 110 \rangle$ for the SAW on GaAs is shown in Figure 2.4 for a zero degree wafer (i.e. the wafer normal direction is aligned with the $[001]$ crystal direction) [36, 39]. From the experimental curve, the anisotropy parameter β can be calculated using the relationship [47]:

$$v_s(\theta)/v_s(0 \text{ rad.}) \approx 1 + (\beta/2)\theta^2 \quad (2.3)$$

where θ is the angular propagation deviation from $\langle 110 \rangle$ in radians and v_s is the SAW phase velocity. The value obtained using Figure 2.4 is $\beta \approx -1.6$. The anisotropy parameter, transducer width, and propagation distance can be used to calculate the propagation loss between two identical transducers due to diffraction of the SAW beam.

For the ACT device built in this thesis, the transducer width and propagation distance were 75λ and 250λ respectively. Using the curves given by Szabo and Slobodnik [46] a diffraction related loss of ≈ 0.4 dB was obtained.

A second example could consist of a much narrower transducer (12.5λ) and a longer propagation distance (1250λ), corresponding to $100 \mu\text{m}$ and 1 cm respectively at 360 MHz . The diffraction induced loss in this case is ≈ 7 dB. Where long propagation distances are desired, the requirement for wide transducers is evident from these calculations. In general, the diffraction loss depends on the ratio of the

Figure 2.4: SAW phase velocity for propagation near $\langle 110 \rangle$

propagation distance to the square of the transducer width where these quantities are expressed in wavelengths at the desired frequency.

Values of β used to calculate the amount of beam steering using the parabolic velocity approximation have been tabulated for many SAW materials and orientations [46]. The present author was unable to find a value tabulated for $\langle 110 \rangle$ propagating SAW on (100) cut GaAs even though values were quoted for (110) cut $\langle 100 \rangle$ and (211) cut $\langle 111 \rangle$ propagation. Based upon the shape of the velocity curves for small angles about $\langle 110 \rangle$ for (100) cut GaAs, it does not seem unreasonable that the parabolic approximation can be applied to estimate diffraction related losses. This assumption ignores any effect of coupling to the bulk shear mode for directions off $\langle 110 \rangle$.

The anisotropy parameter is related to the beam steering angle (ϕ) through the relationship [46]:

$$\tan \phi = (1/v_s(0 \text{ rad.}))dv_s/d\theta \quad (2.4a)$$

where ϕ is the angle between the $\langle 110 \rangle$ pure mode axis and the power flow vector.

For small ϕ , $\tan \phi \approx \phi$ and (2.4a) can be simplified:

$$\phi \approx (1/v_s(0 \text{ rad.}))dv_s/d\theta \quad (2.4b)$$

For a material with phase velocity about a pure mode axis satisfying (2.3), equation (2.4b) can be used to determine the power flow angle ϕ :

$$\phi \approx \beta\theta \quad (2.4c)$$

Assuming an alignment accuracy of $\pm 1^\circ$ along $\langle 110 \rangle$, from

equation (2.4c) the beam steering angle is $\pm 1.6^\circ$. Over a propagation distance of 2 mm, this yields a lateral shift of the SAW beam equalling $\pm 56 \mu\text{m}$. Since the experimental ACT built for this work has a transducer width of $600 \mu\text{m}$ and a channel width of $400 \mu\text{m}$, the lateral shift in SAW caused by beam steering effects should not be a problem as long as the alignment accuracy of $\pm 1^\circ$ is maintained. For longer propagation distances, beam steering becomes increasingly important requiring very accurate alignment of SAW propagation direction with $\langle 110 \rangle$.

By plotting the reciprocal of the SAW phase velocity as a function of propagation direction, the so-called slowness surface is created [47]. SAW energy flow is perpendicular to the slowness surface. Figure 2.5 shows the curve generated for SAW propagation near $\langle 110 \rangle$ on (100) cut GaAs. In comparison to the case shown for an isotropic material, the $\langle 110 \rangle$ propagating SAW on GaAs is slightly focussing. A given diffraction pattern will occur at a greater distance from the acoustic transducer for $\langle 110 \rangle$ propagation on GaAs than for an isotropic material.

The curvature of the experimental and theoretical results shown in Figure 2.5 are quite different. This is due to the curvature difference in the velocity results shown in Figure 2.4. The cause of the difference was not explained by Hunt, Miller, and Hunsinger [39] who published the experimental results.

Figure 2.5: SAW slowness curve for propagation near $\langle 110 \rangle$

Other propagation effects that must be considered for SAW device design include viscous losses, air loading and surface scattering by defects [48]. The latter effect is considered in Section 3.2.

2.2 Depletion Potential in the ACT Channel

Charge transport in the ACT device under consideration is accomplished in the depleted n-epitaxial region of an n-GaAs:semi-insulating GaAs substrate configuration. The assumed band diagram under complete depletion of the n-epitaxial layer is shown in Figure 2.6. An "ohmic" contact alloyed to the back of the device biases the junction between the epitaxial layer and the semi-insulating substrate. The contact and substrate act as a large non-linear resistance. Hence, the effective bias appearing at the junction $|V'_s|$ is much smaller than the bias applied to the substrate contact $|V_s|$.

Potential in the channel is calculated using Poisson's equation and the depletion approximation subject to the following boundary conditions:

$$E_{x3}(x_3 = W) = 0 \quad (2.5a)$$

$$E_{x3} \text{ is continuous at } x_3=W_n \quad (2.5b)$$

$$\phi_d(x_3 = 0) = V_m + \phi_p - \phi_{m1} \quad (2.5c)$$

$$\phi_d \text{ is continuous at } x_3=W_n \quad (2.5d)$$

$$\phi_d(x_3 = W) = V'_s \quad (2.5e)$$

In the calculation it is assumed that the semi-insulating substrate is p-type with an acceptor doping of $\approx 10^{16} \text{ cm}^{-3}$ in proximity to the epitaxial layer junction. This doping density corresponds to the bulk trap density of the substrate. Traps are thought to be responsible for the capture of electrons and release of holes, making the substrate look p-type as far as the static depletion potential calculation is concerned [20, 45, 49]. The dynamics of the capture and release process are slow. Thus, while electrostatically the substrate can be treated as p-type, for time varying electric fields such as the propagating SAW, the substrate appears semi-insulating. Charge screening effects on the SAW can be ignored as a result.

The depletion potential calculation has been reported by Deyhimy, Eden, and Harris [20] for CCD devices and by Hoskins [45] for ACT devices. Although the geometries of the structures under consideration were identical for each calculation, the results differed slightly since Deyhimy et al. did not account for the GaAs and Schottky metal work functions. Hoskins' results for the configuration shown in Figure 2.6 are given below:

For $0 \leq x_3 \leq W_n$:

$$\phi_d(x_3) = \phi_{dm} - qN_D/2\epsilon(x_3 - x_{3m})^2 \quad (2.6)$$

$$\phi_{dm} = qN_D x_{3m}^2 / 2\epsilon + V_m + \phi_p - \phi_{m1} \quad (2.7)$$

$$x_{3m} = W_n - N_A W_p / N_D \quad (2.8)$$

For $W_n \leq x_3 \leq W$:

$$\phi_d(x_3) = qN_A/2\epsilon [x_3^2 - 2Wx_3 + W_n^2(N_D/N_A + 1)] - \phi_{m1} + V_m + \phi_p \quad (2.9)$$

$$W = [2\epsilon/qN_A]^{1/2} [qW_n^2(N_D+N_A)/2\epsilon - \phi_{m1} + V_m - V'_s + \phi_p]^{1/2} \quad (2.10)$$

$$W_p = W - W_n \quad (2.11)$$

where:

q = electronic charge

N_D = epitaxial layer donor density

N_A = equivalent acceptor density at interface between the semi-insulating substrate and the n-epitaxial layer (unoccupied deep trap density)

ϵ = permittivity of GaAs

V_m = bias on Schottky channel plate

V'_s = effective substrate bias

ϕ_p = semi-insulating GaAs work function

ϕ_{m1} = Schottky metal work function

W_n = thickness of n-epitaxial layer

W_p = thickness of depleted 'p-type' layer at interface between the semi-insulating substrate and n-epitaxial layer

W = total depletion layer thickness

Figures 2.7a to 2.7d show plots of the depletion potential for various values of substrate and channel plate bias combinations. The potential distribution is presented for typical bias values used with devices built for this research (Figure 2.7c). Since charge transport occurs at depth x_{3m} where the electric field is zero, these graphs

Figure 2.7a: ACT channel potential - experimental bias condition

Figure 2.7b: ACT channel potential - variation of frontgating bias

Figure 2.7c: ACT channel potential - variation of backgating bias

Figure 2.7d: ACT channel potential - variation of frontgating and backgating bias

show that the depth of charge transport is increased using larger channel plate or smaller substrate bias magnitudes.

Increasing the channel plate and substrate bias simultaneously by equal amounts had no effect on potential variation in the channel. By increasing $|V_m|$ to 8 Volts with $|V'_s| = 2.2$ Volts, the potential well disappears and injected electrons escape to the substrate. Similarly, for large enough backgating bias (V_s) with frontgating bias (V_m) held constant, injected charge flows into the channel plate.

The assumed depletion condition exists provided the epitaxial layer outside the depleted region is biased at a potential equal to or greater than the value indicated at x_{3m} , the electron potential minimum in the channel. The input and output ohmic contacts of the ACT must be biased at a potential equal to or more positive than the value indicated at x_{3m} to achieve depletion of the channel region.

2.3 SAW Transduction on GaAs

Generation of the large amplitude SAW on GaAs is accomplished with an interdigital transducer. The ACT devices built for this thesis used a $\lambda/4$ transducer design with finger and gap widths of $2\mu\text{m}$ as shown in Figure 2.8. The performance of such transducers has been well documented; for example by Milsom, Redwood, and Reilly [50].

One complication that arises when interdigital transducers are used to generate SAW on semiconducting substrates is screening of the SAW fields by free carriers.

Figure 2.8: Quarter wavelength finger width IDT design

By fabricating the transducer using a metal that forms a Schottky barrier with the semiconductor, a reverse bias applied to the fingers relative to a nearby ohmic contact will cause depletion of the region underneath the fingers [51]. The applicability of this approach depends on the transducer design, substrate doping, and substrate thickness since complete depletion to a depth at which SAW RF fields are small must be achieved before reverse voltage breakdown of the Schottky diodes formed by the fingers occurs.

A significant problem with the long $\lambda/4$ transducers is loss induced by internal acoustic reflections. For this IDT design, maximum internal reflections are present at the same frequency greatest radiation conductance is achieved. The internal reflections have two principal origins. One of these is due to the SAW electric field being short-circuited underneath the transducer fingers. This effect is significant for strong piezoelectrics such as LiNbO_3 , but small for weak piezoelectrics such as GaAs. The other mechanism is related to the periodic mass loading of the surface by the IDT. The latter effect is significant for GaAs at high frequencies as transducer finger height becomes a larger proportion of the electrode width.

The frequency response of a GaAs SAW delay line consisting of two 100 finger pair transducers with phase centres separated by 4.38 mm is shown in Figure 2.9. The wafer used was semi-insulating and the fingers consisted of 50 nm of chromium followed by 300 nm of aluminum. The

Figure 2.9: Frequency response of reflective $\lambda/4$ transducer delay line on GaAs

acoustic reflections completely destroyed the sinc function frequency response expected for an unapodized array.

Bogus, Hoskins and Hunsinger [52] described a method to reduce internal reflections in $\lambda/4$ transducers using a layered metal structure. The technique considers the sign and magnitude of acoustic reflections in an IDT structure given by analysis of the metal and substrate material properties. The magnitude of the acoustic reflection depends on the electrode height. By depositing two different metals during interdigital transducer fabrication, a reduction of the internal reflections occurs through a cancellation of the effects caused by each metal.

The basic theory is based on a transmission-line model for the acoustic impedance mismatch created by the periodic IDT array. The acoustic characteristic impedance of the GaAs without any deposited metal is Z_{A0} corresponding to the gap regions of the IDT. The acoustic characteristic impedance of the metallized regions is $(Z_{A0} + \delta Z_A)$ corresponding to the IDT fingers. Since δZ_A is found to vary in sign and magnitude with the deposited metal, the method seeks to make $\delta Z_A = 0$ to eliminate reflections. The assumption underlying this method is that the effects of two metals placed on top of each other linearly superimpose. The theory was experimentally verified using an IDT array with $28\ \mu\text{m}$ features and metal thickness of 250 nm.

Hoskins [45] suggested that a certain amount of transducer reflectivity may increase the SAW transduction

efficiency for materials with low electromechanical coupling coefficients. His argument was based on the fact that radiation conductance increases for a transducer in a SAW resonator configuration [53]. In the extreme case of perfect reflectivity, the SAW will be "trapped" within the IDT in the form of a standing wave. The internal losses of the IDT increase as resonance is approached along with the magnitude of the SAW. As such, Hoskins stated that some value of electrode reflectivity for a transducer of a given length should result in maximized SAW generation efficiency.

Another approach to reduce the problem of internal reflections is the $\lambda/8$, double-finger, or split-finger transducer [54] shown in Figure 2.10. Geometry of the IDT eliminates acoustic reflections in this design which is efficient at generating both the fundamental and third harmonic. Although launching the fundamental at 360 MHz would involve $1\text{ }\mu\text{m}$ geometry, the third harmonic design would require $3\text{ }\mu\text{m}$ features. There is no reported use of this type of transducer in the literature for ACT devices, although they are currently being used in recent ACT device research [55]. Whether these IDT's are being operated at their fundamental or third harmonic generation frequency is unknown.

An approximate equivalent circuit for the transducer is shown in Figure 2.11.

Figure 2.10: Split-finger IDT design

$$B(f) = B_a(f) + 2\pi f C_T$$

Figure 2.11: SAW transducer equivalent circuit

The radiation conductance and susceptance are given for a $\lambda/4$ design [56]:

$$G(f) \approx G_0^N \sin^2 X_N / X_N^2 = \text{radiation conductance} \quad (2.12)$$

$$B_a(f) \approx G_0^N (\sin 2X_N - 2X_N) / 2X_N^2 \quad (2.13)$$

$$B(f) = B_a(f) + 2\pi f C_T \quad (2.14)$$

= transducer susceptance

$$G_0^N = 8f_0 K_S^2 C_S N^2 b \quad (2.15)$$

$$X_N = N\pi(f - f_0)/f_0 \quad (2.16)$$

K_S^2 = SAW electromechanical coupling coefficient

$$C_S = C_T / Nb \quad (2.17)$$

C_T = DC capacitance of transducer

N = number of finger pairs

b = acoustic beamwidth

f = frequency

f_0 = centre frequency of transducer

For the transducer used in ACT devices built for this thesis:

$$N = 100$$

$$f_0 = 360.7 \text{ MHz}$$

$$K_S^2 = 6.4 \times 10^{-4}$$

$$b = 600 \text{ } \mu\text{m}$$

$$C_T = 7.67 \text{ pF}$$

Calculation of the transducer impedance at the resonant frequency of 360.7 MHz yielded:

$$G(f_0) = 1.42 \times 10^{-3} \text{ Siemens}$$

$$B(f_0) = 1.74 \times 10^{-2} \text{ Siemens}$$

This corresponds to an input impedance of:

$$Z_{IN}(f_0) = (4.66 - j57.1) \Omega$$

The lack of agreement between this value and the measured RF impedance of the transducer given in section 4.2 is believed to be due to the combined effect of package parasitics and large internal acoustic reflections which make the expressions for transducer impedance presented above invalid.

2.4 The Acoustoelectric Effect

The transfer of energy and momentum between acoustic waves and carriers in a semiconductor is called the acoustoelectric effect. Weinreich [57] derived a relationship between the magnitude of the acoustoelectric current density (J_{AE}) and the acoustic power density (S):

$$J_{AE} = -2\mu\alpha S/v_S \quad (2.18)$$

where: S = acoustic power density (W/m)

μ = carrier mobility

α = attenuation coefficient

v_S = acoustic wave velocity

J_{AE} = acoustoelectric current density (A/m)

Solving for the acoustoelectric attenuation coefficient:

$$\alpha = -J_{AE}v_S/2\mu S \quad (2.19)$$

The SAW power loss as a function of propagation along x_1 is then given by:

$$S = S_0 \exp(2\alpha x_1) \quad (2.20)$$

where: S_0 = initial acoustic intensity at $x_1 = 0$.

The Weinreich relation is valid for arbitrary acoustic

waveforms and amplitudes with no applied magnetic fields, no carrier trapping effects and only one type of carrier present.

For low values of SAW power, the acoustoelectric current is proportional to the acoustic power intensity. This is the linear or small signal regime in which charge transport occurs at subsonic velocity.

Large values of SAW power lead to bunching and transport of the free carriers at acoustic velocity in the wells defined by the periodic SAW potential. This is termed the large signal or non-linear regime and results when the SAW induced piezoelectric space charge density approaches, or exceeds, the space charge density of the free carriers. In this case it has been shown that the acoustoelectric current saturates [58]. The transported carriers are under the influence of a constant "synchronous" electric field [59] given by:

$$E_S = -v_S/\mu \quad (2.21)$$

This is simply the field required to move the carriers at a constant velocity equal to the sound velocity in a material with mobility μ . For SAW, most of the acoustic energy is confined to within a wavelength of the substrate surface. This fact, and the high electron mobility in GaAs, permit moderate SAW drive powers to be used to achieve the synchronous field.

As an example of the magnitude of SAW loss caused by the acoustoelectric effect an experimental device with charge

injection level adjusted to produce a time averaged output current of $6 \mu\text{A}$ is considered. The acoustoelectric current density is 15 mA/m taking into account a $400 \mu\text{m}$ channel width. The acoustoelectric attenuation coefficient calculated from (2.19) is $\alpha = -0.43 \text{ m}^{-1}$ assuming a wave power intensity of $1 \text{ mW}/\lambda$ (corresponding to a $\approx 0.5 \text{ V}$ optimized wave potential) and an electron mobility value of $4000 \text{ cm}^2/\text{Vs}$. The calculated attenuation coefficient corresponds to a SAW power loss of only 3.7 dB/m . The small magnitude of this attenuation means that very little power is required to "drive" the transport process. The practical device length based on this loss value could be extremely large without severe degradation of the SAW potential. Unfortunately, in an actual device, the SAW attenuation will be dominated by loss mechanisms associated with substrate imperfections and diffraction effects. These considerations limit current device lengths to several millimeters.

It is interesting to consider loss caused by the gap between the interdigital transducer and the ACT channel where the epitaxial layer is not depleted. In this case, the equivalent sheet charge density of the $4 \mu\text{m}$ thick epitaxial region was calculated as $8 \times 10^{11} \text{ cm}^{-2}$ assuming a donor density of $2 \times 10^{15} \text{ cm}^{-3}$. For the purpose of this calculation, it is assumed that the carriers could be transported at the sound velocity by a SAW whose power density is $1 \text{ mW}/\lambda$ acoustic beamwidth. For a $100 \mu\text{m}$ gap, the attenuation under these conditions is only 0.1 dB .

2.5 Optical Charge Injection

Light incident through semi-transparent windows on the ACT device creates electron-hole-pairs (EHP's) in the device channel whose initial spatial distribution can be described by:

$$n(x_3, t_0) = n_0 \exp(-\alpha_0 x_3) \quad (2.22)$$

$$p(x_3, t_0) = n_0 \exp(-\alpha_0 x_3) \quad (2.23)$$

assuming uniform optical intensity over the window in the x_1 and x_2 directions and monochromatic light. The parameter α_0 is the optical absorption coefficient for the given wavelength of illumination. The ACT channel potential is given by Hoskins [45] page 61:

$$\phi_C(x_1, x_3, t) = -(qN_D/2\epsilon)(x_3 - x_{3m})^2 + |\tilde{\phi}_0(x_3)| \cos(kx_1 - \omega t) \quad (2.24)$$

where the first term describes the parabolic depletion potential excluding components with no spatial dependence. The second term describes the travelling wave potential due to the SAW.

In practice, the x_3 -component of the depletion electric field is much larger than the x_3 -component of the SAW electric field. Hence, a simple analysis describing charge separation in the x_3 -direction can ignore this component to obtain an approximation of the charge separation behaviour.

The channel potential (ignoring the SAW) is therefore given by:

$$\phi_C(x_3, t) \approx -(qN_D/2\epsilon)(x_3 - x_{3m})^2 \quad (2.25)$$

The electric field due to the depletion potential alone is given by:

$$E_{CP}(x_3, t) = -d\phi_C(x_3, t)/dx_3 \approx (qN_D/\epsilon)(x_3 - x_{3m}) \quad (2.26)$$

Using some typical values:

$$N_D = 2 \times 10^{15} \text{ cm}^{-3}$$

$$\epsilon = 11.0 \epsilon_0$$

$$x_{3m} = 3.5 \text{ } \mu\text{m}$$

yields:

$$E_{CP}(x_3, t) \approx 3.29 \times 10^6 (x_3 - 3.5) \text{ V/m where } x_3 \text{ is in } \mu\text{m}. \quad (2.27)$$

The saturation drift velocity (v_{SAT}) for carriers in GaAs is $\approx 7 \times 10^6 \text{ cm/s}$ [60]. In the analysis that follows, it has been assumed that velocity saturation is achieved for electrons and holes with an electric field of 3.3 kV/cm and 25 kV/cm respectively.

For electrons, the depletion field exceeds the threshold over $0 \leq x_3 \leq 3.4 \text{ } \mu\text{m}$. Hence, to an approximation, the electrons move to the centre of the ACT channel at or above their saturated drift velocity.

For holes, the field for velocity saturation is achieved over $0 \leq x_3 \leq 2.7 \text{ } \mu\text{m}$. Hence, for carrier injection where the optical penetration depth is less than $2.7 \text{ } \mu\text{m}$, the holes and electrons are separated at their saturated drift velocities. This is a highly desirable condition for reduction of recombination and transit time.

At $\lambda_p = 730 \text{ nm}$, the optical absorption coefficient is $\alpha_0 = 1.92 \times 10^4 \text{ cm}^{-1}$ [61]. At a depth into the substrate of $(4/\alpha_0) = 2.1 \text{ } \mu\text{m}$, over 98% of incident photons have been absorbed by the GaAs. This is within the $2.7 \text{ } \mu\text{m}$ limit just calculated for velocity saturation for holes.

A qualitative analysis of the charge separation process will be discussed. Initial electron and hole distributions in the channel are given by equations (2.22) and (2.23). It is assumed that the electric field in the channel exceeds the electron saturation fields from the device surface at $x_3=0$ to the electron potential minimum at $x_3=x_{3m}$. The optical penetration depth is assumed to be small enough that hole velocity saturation applies. For $x_3>x_{3m}$ the electric field changes sign. Ignoring diffusion and recombination, the initial electron and hole charge distributions move in opposite directions at v_{SAT} until the electrons reach $x_3=x_{3m}$ and the holes reach $x_3=0$ as shown in Figure 2.12. In this simple analysis, the final electron and hole charge distributions can be regarded as forming a parallel plate capacitor with the electrons in the channel forming one plate and the holes, or positive charge, on the Schottky metal at the surface forming the other plate. A current flows in the channel plate when the electrons are removed from the channel, effectively discharging the capacitor. This last result has been verified with an actual ACT device by monitoring the channel plate current before, during, and after application of the optical pulse. The result is shown in Figure 2.13. A current pulse appeared when electrons were removed from the channel of the device at the output node. The polarity of this current pulse indicated electron flow into the channel plate, in agreement with this analysis.

Figure 2.12: Simplified analysis of optical charge separation

Top: Voltage to LED

5 V/div

Middle: Channel

Plate Current

$\approx 4 \mu\text{A}/\text{div}$

Bottom: Output Ohmic

Current

$\approx 4 \mu\text{A}/\text{div}$

Figure 2.13: Channel plate waveform after application of optical pulse

In summary, this simplified theory predicts that at $t \approx 3.4 \mu\text{m} / v_{\text{SAT}} \approx 50 \text{ ps}$, the charge has been completely separated and the photo-excited electrons have drifted to the centre of the channel where they are bunched and transported by the SAW analogous to electrically injected carriers. Velocity overshoot has been ignored, but in practice could reduce even further the charge separation times. Comparing the 50 ps charge separation with the minority carrier lifetime in GaAs, several ns, the amount of recombination should be small, justifying an assumption made earlier.

The calculation did not account for changes in static depletion potential due to the presence of injected carriers. The effect of this additional charge on the channel potential is expected to be small for low injection levels. For example, injection of a single carrier species into the ACT and comparison of the density of injected carriers with the background charge density due to the ionized donors was considered.

For an injection level of 1.74×10^5 electrons/well corresponding to a $10 \mu\text{A}$ output current, the maximum charge density due to the injected electrons is $7 \times 10^{13} \text{ cm}^{-3}$ for a slit dimension of $16 \mu\text{m} \times 300 \mu\text{m}$ and an optical absorption coefficient of $\alpha_0 = 1.92 \times 10^4 \text{ cm}^{-1}$ assuming all electron-hole-pairs are created at $t=t_0$. This charge density value applies only near the substrate surface since charge injection follows an exponential relationship with depth. The maximum injected charge density is relatively small

compared to the background doping density of $\approx 2 \times 10^{15} \text{ cm}^{-3}$ and is reduced further by a compensating effect of the holes in an actual optical injection situation and by the finite time during which the optical pulse can inject charge into a given well, approximately 2.8 ns corresponding to the amount of time that a SAW charge well is under the injection window.

Diffusion effects have not been included in this analysis. This appears justified in light of the high drift fields responsible for charge separation and the small separation time. Trapping effects caused by surface states have been ignored since these traps are expected to occupy a small fraction of the total electron-hole-pair generation volume.

2.6 Charge Capacity and Transfer Efficiency

The performance of a charge transfer device is quantified in terms of the amount of charge that can be carried in each potential well at a given transfer efficiency. In the case of a CCD, the charge transfer efficiency is the fraction of charge that is successfully carried through one complete clocking cycle. An analogous definition for an ACT would be the fraction of charge transported one acoustic wavelength.

In general, the amount of charge lost during each clocking interval is a non-linear function of the total charge load. In practice, a proportional loss mechanism is often assumed, where the total amount of charge lost is

proportional to the total amount being transported.

Results of the theory presented by Hoskins to describe charge transfer behaviour of the ACT device will be briefly described. A detailed development of the theory is given by Hoskins [45] and Hoskins and Hunsinger [62]. The assumptions in the analysis are listed below:

- . the channel centre to image conductor separation is at least one quarter of an acoustic wavelength
- . only electrons are assumed to exist in the channel
- . hot electron effects are ignored
- . no carrier generation-recombination is assumed
- . the acoustoelectric attenuation of the SAW is not considered
- . channel edge effects are ignored
- . the field required to move the electrons at the SAW velocity is assumed small compared to the SAW electric field magnitude
- . a plane wave approximation to the SAW wave potential is assumed

Most of the assumptions appear justified in light of the device geometry and normal operating conditions. One point of importance in a practical device which has been ignored in the analysis, is the SAW attenuation. Although acoustoelectric attenuation of SAW for high mobility materials is small, as discussed in section 2.4, the attenuation due to scattering from surface imperfections and diffraction effects is not insignificant even for reasonably

small propagation distances. Incorporation of a propagation distance dependent parameter into the model would perhaps be justified.

The charge packet shape is determined by a two-dimensional solution of Laplace's equation in the device channel taking into account the electrostatic depletion field, that due to the SAW, and the field due to transported charge. Shape of the charge boundary is shown in Figure 2.14 for a charge load of $1/2$. Packet shape depends on the channel constant ρ_D/ρ_W defined by [45, 62]:

$$\rho_D/\rho_W = qN_D/\phi_0\epsilon k^2 \quad (2.28)$$

where q is the electronic charge, N_D is the ionized donor density in the channel, ϕ_0 is the SAW potential, ϵ is the permittivity of GaAs, and k is the acoustic wavenumber. The channel constant is the ratio of the space charge density due to ionized donors to the piezoelectric polarization space charge density. Intuitively, one would expect large channel constants to increase the electron confining field in the device channel resulting in a flattening of the charge packets in the direction of transport. This is indeed the case (Figure 2.14). Under conditions of maximum charge load, the packet dimension in the propagation direction is approximately one half of an acoustic wavelength.

The mechanism of charge transfer inefficiency is assumed to be the diffusion of packet charge over the potential barrier imposed by the SAW. This is a symmetric effect with

From: Hoskins [45]

Figure 2.14: Charge packet shape for charge load of $1/2$

equal charge being dispersed into leading and trailing wells. As the charge load is increased, the potential barrier provided by the SAW is reduced, leading to increased diffusion induced spreading of the charge. To calculate the magnitude of this effect, a worst case assumption of thermionic emission of the packet charge over the barrier provided by the SAW was used. Results of this calculation are given in Figure 2.15, showing charge transfer inefficiency as a function of charge capacity, N_e = number of electrons per centimeter of channel width, with the SAW potential as a parameter.

A useful expression provided by Hoskins, Brophy, Dallesasse, Miller, and Peterson [27] relates the maximum charge capacity (N_e) to the wave potential (ϕ_0) and is given by:

$$N_e(\text{electrons/cm}) \approx 53 \times 10^6 (\phi_0 - 0.26 \text{ Volts}) \quad (2.29)$$

This expression indicates that the first $10kT/q \approx 260$ mV of wave potential is required to provide charge confinement, with additional wave potential contributing linearly to the charge capacity.

From: Hoskins [45]

Figure 2.15: Charge transfer inefficiency for different SAW wave potentials versus charge load

3.0 EXPERIMENTAL DEVICE DESIGN AND FABRICATION

3.1 Mask Layout

Figure 3.1 shows mask levels for the experimental ACT device. Patterns were generated on a Metheus $\lambda 700$ computer aided design station and fabricated at Precision Photomasks Inc. in St. Laurent, Québec. The device shown is similar in layout to the second generation ACT device described by Hoskins [45, 63] with the addition of windows in the channel plate so that optical charge injection could be achieved.

To simplify fabrication and reduce costs, the number of mask levels needed to implement the device was minimized. The guard ring isolation technique [19] was chosen since it could be implemented at the same time as other Schottky features. Isolation methods such as mesa etching [64] and proton bombardment [65] require additional mask levels and fabrication steps. Disadvantages of the guard ring approach include the fact that, by surrounding the active device, it becomes difficult to route additional metallization patterns into the channel area without resorting to multilevel techniques. The guard ring also requires an additional power supply for biasing.

An additional mask to build up device bonding areas was eliminated in the experimental device. While this would be a poor decision for a production ACT that must withstand environmental and vibrational stress, it did not seem to create any problems in the laboratory environment.

Figure 3.1: Experimental ACT mask layout

The mask levels shown in Figure 3.1 include:

- 1) Ohmic features
- 2) Schottky features
- 3) Window features

A fourth level labelled "dielectric" was also present for experiments designed to integrate optically induced charge outside the SAW beamwidth. This particular mask level was designed such that its omission would not affect the operation of the basic device, as it was not used for work presented in this thesis.

3.2 Epitaxial GaAs Substrate Material

The substrate material for the experimental ACT devices was a vapour phase epitaxy (VPE) n-GaAs layer grown on a Cr-doped horizontal Bridgman substrate. The geometric thickness and doping concentration of the epitaxial layer were $4\text{ }\mu\text{m}$ and $2 \times 10^{15}\text{ cm}^{-3}$ respectively. The light doping was near the lower limit achievable using standard GaAs VPE growth techniques when no intentional impurities are introduced in the growth process. Residual doping is believed to be a result of silicon "contamination" from the quartzware of the epitaxial growth chamber [66].

The first reported ACT device was built using molecular-beam-epitaxial (MBE) material [67]. Problems of excessive SAW attenuation ($\approx 6\text{ dB}/\mu\text{s}$) were attributed to the poor surface morphology of MBE layers that were several microns thick. The defects were found to be of approximately the

same dimension as the SAW wavelength ($\approx 8 \mu\text{m}$). In this growth technique, such surface imperfections are known as oval defects and typically number 10^3 - 10^4 cm^{-2} [68]. More recent ACT devices have been built on VPE grown n-epitaxial layers using an undoped liquid-encapsulated Czochralski (LEC) substrate.

The surface of the VPE wafers received from a commercial supplier for this work had small surface ripples that were visible with an optical microscope. The wavelength of these undulations was generally longer than the SAW wavelength. According to Hoskins [45] pages 41-42, VPE grown epitaxial layers gave considerably less SAW scattering ($\approx 3 \text{ dB}/\mu\text{s}$) than MBE grown epitaxial layers. Epitaxial material was found to have higher SAW attenuation than well-polished bulk semi-insulating substrates.

In a study of [111] propagating SAW on (211) cut and [100] propagating SAW on (110) cut GaAs at 1 GHz, Slobodnik [69] concluded that surface imperfections much smaller than the acoustic wavelength do not affect SAW attenuation. From these results, it may be possible to conclude that surface imperfections much smaller or much larger than the SAW wavelength do not adversely affect propagation.

Shortly after the VPE grown material had been received for this project, correspondence with M.J. Hoskins revealed that lightly doped epitaxial layers grown on Cr-doped substrates present several problems when used to build ACT devices [55]. One problem relates to the out-diffusion of

chromium from the substrate into the epitaxial layer during the growth process, causing a partial compensation of the n-doping [70]. Indeed, although an n-epitaxial layer geometric thickness of $4\text{ }\mu\text{m}$ was specified in the material order, the delivered wafers were only effectively doped at $2 \times 10^{15}\text{ cm}^{-3}$ to a depth of $\approx 2.2\text{ }\mu\text{m}$. The reduction of the n-epitaxial layer thickness impacted the depth of charge transport and hence the wave potential achieved in the device. This in turn led to a reduction in the charge load that could be transported since charge transport occurred at a depth less than the optimum ≈ 0.45 wavelengths below the surface of the substrate at which wave potential peaks.

An additional problem with the use of Cr-doped substrates in the context of ACT device performance is their tendency to exhibit less backgating behaviour [49]. Since backgating is used to vary the depth of charge transport, excessively large voltages could be required with some Cr-doped substrates [55]. Testing performed on devices made for this research (section 4.1) revealed that backgating voltages approximately 40% larger than those measured by Hoskins [45] page 118 were required. However, this increase was not so great as to place the voltages out of the range of a standard 50 volt laboratory supply. The wafers used were relatively thin (16 mils). This may have helped reduce backgating voltages.

While not considered in this thesis, the use of metal-organic chemical vapour deposition (MOCVD) grown epi-layers

might be preferable for future ACT device work. Good surface morphology has been obtained using lower growth temperatures than VPE ($\approx 650^{\circ}\text{C}$ versus $\approx 750\text{-}850^{\circ}\text{C}$ [71]). The lower growth temperatures might result in fewer substrate impurity out-diffusion problems. Since MOCVD growth uses a cold-wall vessel, Si contamination should be reduced [72], allowing easier growth of the thick epitaxial layers with the low doping density used for ACT devices.

3.3 Basic Fabrication Steps

Fabrication of the ACT devices involved the following steps:

- 1) Pre-clean of the wafer in a 1% filtered Alconox (TM) solution.

- 2) Deposition of a Au:Ge/Ni/Au ohmic metallization on the semi-insulating substrate side of the wafer.

Thickness: Au:Ge = 96 nm
 Ni = 20 nm
 Au = 130 nm

- 3) Patterning and deposition of ohmic features on the device frontside using the Au:Ge/Ni/Au metallization system.

(Same thicknesses as above)

Alignment of the propagation direction was made with a {110} GaAs cleavage plane.

4) Alloying of the ohmic features in an open tube furnace at 450 degrees Celsius for 1.5 minutes in a Nitrogen atmosphere.

5) Patterning and deposition of the Schottky features using a Cr/Al:Cu(1%) metallization system.

Thickness: Cr = 50 nm
Al:Cu(1%) = 300 nm

6) Patterning and deposition of the thin semi-transparent Cr metal window features.

Thickness: Cr = 25 nm

7) Dicing and mounting of the completed devices in a 16 pin dual-in-line package. RTV silicon rubber was applied to each end of the chip and wirebonding performed with 0.8 mil gold wire using a thermo-compression wirebonder. All bonds used one wirebond/pad except for the interdigital transducer pads where two wirebonds were made.

The ohmic contacts used for this thesis differed slightly from those reported by Hoskins [45, 63] in which a Au:Ge/Ni system was described. The addition of a Au over-

layer tends to produce ohmic contacts with smoother surface texture and may yield better specific contact resistivity values [73]. Since the ohmic bias ring and input/output contacts are in the propagation path of the SAW, it was felt that a smooth surface for low scattering was desirable.

Copper was added to the aluminum Schottky metallization in order to reduce the effect of "acoustomigration" of metal which could lead to destruction of the interdigital transducer [55]. The Cr/Al:Cu metallization system has been used in silicon work to reduce the effects of electromigration [74, 75]. Copper is an undesirable metal to have in direct contact with GaAs since it can diffuse into the substrate, getter onto crystalline defects, and has a high chemical reactivity as pointed out by Williams [76]. Hence, a chromium layer was evaporated first to act as a diffusion barrier. The excellent adhesion of chromium to GaAs was an additional benefit of adding this layer.

The lift-off technique [77] was used for all metallization patterning. Fabrication experience was first developed using undoped GaAs wafers and a scanning electron microscope (SEM) to check photoresist and metal edge profiles. Several problems that were encountered and solved during this phase of the work are discussed below.

To achieve undercut photoresist profiles essential for good lift-off, the chlorobenzene method was used [78]. After soft-baking the photoresist, the wafer is dipped in an aromatic solvent such as chlorobenzene either before or after

exposure to the ultraviolet light in the mask aligner. It is believed that low-molecular-weight resin and solvent is leached out of the photoresist by a diffusion process. This results in a toughened surface layer that dissolves at a greatly reduced rate in the developer solution relative to exposed photoresist. The hardened surface layer gives the desired undercut edge profile after development.

For devices built during this research, soak times were 7 minutes just prior to placing the wafer in the developer solution. Initially, a photoresist soft-bake temperature of 95° Celsius was used but this led to "wings" on the lifted-off metal patterns shown in Figure 3.2. Lowering the soft-bake temperature to 70° Celsius eliminated this problem and yielded acceptable metal edge profiles, (c.f. Figure 3.3). This observation is in agreement with results reported by Hatzakis, Canavello and Shaw [78]. They explained that the photoresist resin molecules are more tightly packed after a higher temperature soft-bake reducing the diffusion rate of solvent and low-molecular-weight resin.

No ultrasonic agitation was used during the lift-off process for fear of shattering the wafer. Instead, a spray-bottle of acetone was found to be an effective lift-off aid and possibly safer.

Good mask-to-wafer contact was required for proper photoresist edge profile because of optical diffraction effects. To achieve acceptable contact, a precision mask aligner, which operates by placing a small vacuum between the

Figure 3.2: SEM of metal "wings" on transducer fingers
caused by poor photoresist edge profile

Figure 3.3: SEM of 2 μm interdigital transducer fingers

wafer chuck and the mask, was used to pull the wafer and mask into intimate contact.

To provide a clean GaAs surface prior to all deposition steps, a 20 second $1\text{NH}_4\text{OH}:10\text{H}_2\text{O}$ dip was used in accordance with results presented by Miers [79]. Immediately after this final cleaning step, the wafer was loaded into the evaporation chamber and pump-down commenced.

Photoresist baking was observed during thermal evaporation of chromium from tungsten or molybdenum boats as shown in Figure 3.4. This in turn distorted the patterning and made lift-off very difficult. Switching to chromium coated tungsten rods eliminated this effect, probably because better thermal contact between the chromium and tungsten allowed evaporation to proceed without excessively heating the substrate.

A similar problem noticed while thermally evaporating gold from tungsten or molybdenum boats was eliminated using an integrated tungsten-alumina crucible. As such, the heat required to liquify the gold was more closely confined to the area where the melt was located.

During some of the lift-off processes, Hunt microstrip (TM) was used to remove residual photoresist. If any water was present in the microstrip solution, as might occur if a wet beaker were used, etching of aluminum metallization occurred (Figure 3.5). Care in ensuring that beakers were completely dry before adding microstrip solved this problem.

Figure 3.4: SEM of "baked" photoresist caused by excessive heat during evaporation

Figure 3.5: SEM of etched 1 μm aluminum fingers caused by moisture in Hunt microstrip (TM)

A diced ACT device, prior to application of acoustic absorber material, mounting and wirebonding is shown in Figure 3.6. Closer views of the input and output structures are displayed in Figures 3.7a and 3.7b respectively.

3.4 Test Structures and Process Control

In order to assess the success of each stage in the fabrication process from an electrical viewpoint, test structures were included in the ohmic and Schottky mask layers.

Ohmic contact resistance was measured using the test pattern found in the inset of Figure 3.8 [73]. By graphing the resistance versus gap width for this structure, the curve shown in Figure 3.8 was obtained. Approximate values for the GaAs sheet resistance and ohmic contact resistance were derived by measuring the slope and resistance axis intercept of this graph respectively.

By making a rough approximation that the sheet resistance of the GaAs under the alloyed contact was the same as the sheet resistance of the n-epitaxial wafer alone, values of specific contact resistivity and transfer length were obtained. With data derived from the graph, specific contact resistivity (r_c) and transfer length (l_t) were $\approx 2.7 \times 10^{-5} \text{ ohm-cm}^2$ and $0.75 \text{ } \mu\text{m}$ respectively. A more accurate estimate of specific contact resistivity and transfer length would require the measurement of end resistance. However, for contacts whose dimensions are of the order of $150 \text{ } \mu\text{m}$,

Figure 3.6: Photograph of completed ACT device

Figure 3.7a: Photograph of ACT input features

Figure 3.7b: Photograph of ACT output features

Figure 3.8: Ohmic contact test structure resistance variation with gap width

this is a difficult measurement since the end resistance rapidly approaches zero for contact lengths which are several transfer lengths long [80].

The doping profile of the n-epitaxial layer was measured by the capacitance-voltage or C-V method [81]. ACT device channel plate and surrounding ohmic bias ring were used for this measurement. Results shown in Figure 3.9 were in good agreement with data provided by the wafer manufacturer.

The Schottky ideality factor was measured using the diode shown in Figure 3.10. The current-voltage characteristic yielded an ideality of 1.05, indicating acceptable surface preparation and evaporation technique.

DOPING PROFILE

SAMPLE= ACT1.RUN1

.3V TO -8V SWEEP

Figure 3.9: Epitaxial wafer doping profile measured by the capacitance-voltage method

Figure 3.10: Schottky diode current-voltage plot used to determine the ideality factor

4.0 MEASURED PERFORMANCE

4.1 Frontgating and Backgating Behaviour

Omitting the interdigital transducer, the ACT device can be viewed as a MESFET with a very long gate. The input ohmic forms the source, the channel plate the gate, and the output ohmic the drain. Using the biasing arrangement shown in Figure 4.1, the MESFET drain-source current was modulated by applying bias voltages to the channel plate (frontgating) or substrate (backgating).

The measured characteristics for a typical ACT are shown in Figures 4.2a and 4.2b. Frontgating pinch-off voltage was between -5 to -6 V. This was in good agreement with the calculated value of -5.6 V assuming a donor density of $1.9 \times 10^{15} \text{ cm}^{-3}$ and epitaxial layer thickness of $2.2 \text{ } \mu\text{m}$. Backgating voltages were much larger to achieve the same amount of source-drain current change. This can be attributed to a large voltage drop between the "ohmic" contact alloyed to the semi-insulating substrate and the active device channel.

Curves shown in Figures 4.2a and 4.2b provide data used to calculate the depletion potential in the channel of the device (section 2.2) and hence the depth of charge transport. For instance, from the graphs, it can be seen that -2 to -3 V of frontgating bias has the same effect as -18 V of backgating bias on the drain-source current.

V_S = Semi-insulating substrate "ohmic" tied to Ground for Frontgating and made negative for Backgating Characteristic

Figure 4.1: Experimental set-up for measurement of frontgating and backgating behaviour

$V_{SR} = 0 \text{ Volts}$

Figure 4.2a: Frontgating characteristic

$V_G = 0$ Volts

Figure 4.2b: Backgating characteristic

4.2 Interdigital Transducer RF Impedance

The interdigital transducer return loss magnitude and phase were measured as a function of reverse bias applied to the transducer fingers using a network analyzer. The results are shown in Figures 4.3a to 4.3d.

With no bias applied to the IDT the peak in the return loss indicative of SAW generation was almost non-existent and the out-of-band return loss was greater than 0.5 dB indicating a wide-band power loss. As the reverse bias on the transducer fingers was incremented, return loss increased monotonically in magnitude to a maximum value of -5 dB at -14 V. Further increases in the finger bias produced no change in the transducer impedance, indicating depletion of the epitaxial layer at ≈ -14 V (Figure 4.3c).

The depletion condition could be obtained by a combination of frontgating and backgating. For example, Figure 4.3d shows the return loss and phase response with only -5 V of frontgate bias combined with -20 V of substrate bias.

The reverse bias required for complete depletion beneath the IDT was much larger than the pinchoff voltage of the MESFET portion of the ACT measured earlier (c.f. section 4.1). This is not surprising when one considers that depletion of the IDT gap regions is achieved solely by the fringing fields of the finger regions. To calculate the magnitude of depletion potential would require a two-dimensional field analysis of the periodic IDT structure.

(a)

(b)

Note: Top trace is return loss magnitude
Bottom trace is return loss phase

Figure 4.3: Variation of IDT return loss with biasing

<--- 0 dB

$V_S = 0$ V

$V_{IDT} = -14$ V

(c)

<--- 0 dB

$V_S = -20$ V

$V_{IDT} = -5$ V

(d)

Note: Top trace is return loss magnitude

Bottom trace is return loss phase

Figure 4.3: Variation of IDT return loss with biasing

A similar discrepancy between MESFET pinchoff voltage and voltage required for IDT depletion appeared in Hoskins' results [45] pages 32 and 118 although no explanation was provided.

Behaviour of the out-of-band loss requires clarification. Increasing the IDT finger bias from 0 V initially caused an increase in the out-of-band return loss to a maximum of ≈ -1 dB at -8 V. Further increases in the bias caused the out-of-band return loss to decrease to its minimum value at -14 V.

One possible explanation for this behaviour might be the large concentration of free carriers screening the transducer at 0 V reverse bias which made it look like an effective RF short circuit out-of-band. As reverse bias was increased, the density of free carriers was reduced and the transducer looked resistive, hence the finite return loss. Increasing still further the IDT reverse bias, a condition of complete depletion was approached indicating an open-circuit condition and a return loss near 0 dB.

Different behaviour was noted with some transducers. These had out-of-band loss that decreased monotonically with increasing IDT reverse bias. This is in agreement with the results presented by Hoskins [45] page 31.

For reasons not clear to the author, the return loss peak in the results presented by Hoskins [45] page 31 was much deeper, with -12.5 dB return loss at full depletion. The value indicated by the graph [45] page 31 was not

consistent with the centre frequency impedance quoted [45] page 33, from which a return loss of -8 dB was calculated. Even with a modest return loss of -5 dB, $\approx 68\%$ of the incident power was absorbed by the transducer. In future devices, the reflection loss could be reduced using matching techniques to transform the IDT impedance shown in Figure 4.4 to $50\ \Omega$.

The low frequency IDT capacitance measured using an LCR meter was 7.9 pF compared to a calculated value of 7.7 pF and one obtained by Hoskins [45] page 33 of ≈ 7 pF.

4.3 Calculation of Wave Potential

The relationship between RF power applied to the interdigital transducer and the resulting SAW wave potential will initially be discussed for an IDT with no internal acoustic reflections. In this case, the frequency response of the transducer is a sinc function with nulls every 3.6 MHz about 360 MHz for a 100 finger pair unit. By measuring the return loss at centre frequency and at one of the first frequency nulls, the SAW power can be calculated assuming loss mechanisms in the transducer are constant between these two frequencies. Figure 4.5 shows the power components under consideration.

Power incident on the transducer (P_{IN}) is partially reflected (P_{REF}). P_{REF} can be calculated from the measured return loss value. Of the net power incident on the transducer, a certain proportion is lost in the conversion

Figure 4.4: Polar plot of IDT reflection near resonance

IDT Frequency Response

$$X_N = \frac{N\pi(f - f_0)}{f_0}$$

Experimental Device

$$N = 100$$

$$f_0 = 360 \text{ MHz}$$

Figure 4.5: SAW generation power components

process between electrical and acoustic energy. This loss is represented by P_{CONV} .

Since the transducer is symmetric in the direction of propagation, it will launch SAW equally either way (i.e. $P_{\text{SAW}+} = P_{\text{SAW}-}$). Since no reflector array is being used to redirect the $P_{\text{SAW}-}$ component, this amount of power (i.e. 3 dB) is lost.

At the centre frequency the power balance equation can be written:

$$P_{\text{SAW}+} + P_{\text{SAW}-} = P_{\text{IN}} - P_{\text{REF}} - P_{\text{CONV}} \quad (4.1)$$

At the first frequency null, since $P_{\text{SAW}+} = P_{\text{SAW}-} = 0$, we can write:

$$0 = P_{\text{IN}} - P_{\text{REF}'} - P_{\text{CONV}} \quad (4.2)$$

where $P_{\text{REF}'}$ is the new value of reflected power at this frequency.

Equations (4.1) and (4.2) can be solved to yield the value of $P_{\text{SAW}+}$, the power in the forward launched wave in terms of measured quantities.

With the calculated SAW power value, the quasistatic potential curve (Figure 2.3b) is used, after appropriate denormalization, to give the distribution of SAW potential into the substrate. The depth of charge transport is calculated from the known bias values and substrate parameters using the analysis presented in section 2.2. Knowing the charge transport depth, the wave potential can be read directly from the denormalized quasistatic potential curve.

Unfortunately, internal reflections in long transducers of the type used for the device in this research introduce additional frequency dependent loss not accounted for in the analysis just presented. The internal reflections alter the sinc function frequency dependence of the transducer, as previously discussed in section 2.3. Hoskins [45] accounted for this additional loss by comparing one-port IDT impedance with the frequency response of the transducer, as measured with a laser probe [82].

An alternate approach to obtain the SAW potential in the presence of internal reflections will be derived. The frequency response of a two transducer delay line is considered. At some magnitude value $-Y_{dB}$ below the peak in the response at the centre frequency f_0 , the frequency width can be measured δf . Since the loss of a single transducer is desired, a correction for the difference in frequency response between the latter and a delay line configuration consisting of a transducer pair must be made. In the frequency domain, the transfer functions of each transducer multiply to give the delay line response (ignoring a phase term accounting for the distance between the transducers). Hence, the single transducer magnitude in decibels Y'_{dB} is simply half the delay line magnitude in decibels Y_{dB} . At $f_0 \pm \delta f$:

$$Y'_{dB} = Y_{dB}/2 \quad (4.3)$$

The return losses RL and RL' are measured at f_0 and $f_0 \pm \delta f/2$ respectively. The reflected powers are then:

$$P_{REF}(f_0) = P_{IN} 10^{-RL/10} \quad (4.4)$$

$$P'_{REF}(f_0 \pm \delta f/2) = P_{IN} 10^{-RL'/10} \quad (4.5)$$

At f_0 , conservation of energy requires:

$$P_{SAW+} + P_{SAW-} = P_{IN} - P_{REF} - P_{CONV} \quad (4.6)$$

For a bidirectional IDT, $P_{SAW+} = P_{SAW-}$ and (4.6) can be reduced to:

$$2P_{SAW+}/P_{IN} = 1 - P_{REF}/P_{IN} - P_{CONV}/P_{IN} \quad (4.7)$$

At $f_0 \pm \delta f/2$, conservation of energy requires:

$$P'_{SAW+} + P'_{SAW-} = P_{IN} - P'_{REF} - P_{CONV} \quad (4.8)$$

where $\delta f/2$ is assumed small enough so that the losses (P_{CONV}) are the same.

But, the forward wave SAW power at $f_0 \pm \delta f/2$ (P'_{SAW+}) is related to the forward wave SAW power at f_0 (P_{SAW+}) through:

$$P'_{SAW+} = 10^{-Y' \text{ dB}/10} (P_{SAW+}) \quad (4.9)$$

Hence, at $f_0 \pm \delta f/2$ for a bidirectional array, using (4.8) and (4.9):

$$2 \times 10^{-Y' \text{ dB}/10} P_{SAW+}/P_{IN} = 1 - P'_{REF}/P_{IN} - P_{CONV}/P_{IN} \quad (4.10)$$

Equations (4.7) and (4.10) can be used to solve for P_{SAW+}/P_{IN} , the fraction of incident electrical power converted to the forward travelling SAW in terms of measured quantities:

$$\frac{P_{SAW+}}{P_{IN}} = \frac{(P'_{REF}/P_{IN} - P_{REF}/P_{IN})}{2(1 - 10^{-Y' \text{ dB}/10})} \quad (4.11)$$

Individual loss components at f_0 can be calculated from the following expressions:

$$\text{Electrical} = -10\log[(P_{IN}-P_{REF})/P_{IN}] \quad (4.12)$$

Mismatch Loss

$$\text{Conversion} = -10\log[2P_{SAW+}/(P_{IN}-P_{REF})] \quad (4.13)$$

Loss

$$\text{Bidirectional} = 3 \text{ dB}$$

Loss

The loss due to acoustic reflections could be calculated by measuring the transducer return loss far enough away from f_0 such that the narrowband reflection losses were small relative to those due to other factors such as finger conductivity, shunt conductance between the fingers, bulk mode generation, and so on. At a frequency null far enough away from f_0 so that reflection loss can be ignored but close enough to f_0 so that the other loss components have not changed appreciably, the power balance can be written as:

$$0 = P_{IN} - P_{REF} - P'_{CONV} \quad (4.14)$$

$$\text{where: } P_{REF} = P_{IN}10^{-RL''/10} \quad (4.15)$$

and RL'' is the return loss in decibels measured at the frequency null.

The difference in power lost between the latter case (P'_{CONV}) and the former case (P_{CONV}) is that due to acoustic reflections.

The analysis above could not be used to estimate the transducer conversion loss for highly reflective IDT's employed in ACT devices built for this work. This is because

the validity of the analysis depends on: 1) identical transducers for the delay line and 2) no intertransducer reflections. Measurement of the return loss for each transducer in the delay line configuration yielded a centre frequency difference of 0.1-0.3 MHz between the IDT's. Thus, condition 1) above was violated. Also, the highly reflective nature of the IDT's built for this research violated condition 2).

The method would be valid for transducers which are less reflective, such as split-finger types. Also, the analysis could be used to derive loss components for a reflective array if a laser probe were available. One would have to modify equation (4.3) such that $Y'_{dB} = Y_{dB}$. Unfortunately, the laser probe is not a standard piece of equipment in most laboratories.

Assuming a conversion loss value similar to that quoted by Hoskins [45] page 33, and calculating mismatch loss using the return loss shown in Figure 4.3c and equation (4.12) the loss components are as given in Table 4.1:

Table 4.1: IDT loss components	
Electrical Mismatch Loss	= 1.7 dB
Conversion Loss	= 3.1 dB
Bidirectional Loss	= 3 dB

Total Loss	= 7.8 dB

Using the analysis presented in section 2.2 and the device bias values, the estimated depth of charge transport

was $\approx 1.7 \mu\text{m}$. After appropriate denormalization of the quasistatic potential curve (Figure 2.3b), the SAW potential in the ACT channel was calculated from the input electrical power. The relationship derived was:

$$\phi_0 \approx 13\sqrt{P_{\text{IN}}(\text{mW})} \text{ mV} \quad (4.16)$$

4.4 Device Biasing

Figure 4.6 shows the ACT biasing scheme including the output amplifier stage. The entire circuit was implemented with standard RF "breadboard" techniques using coaxial cable to reduce stray noise pickup.

Typical bias values are listed in Table 4.2:

Table 4.2: Typical ACT device bias values	
V(guard ring)	= -11.3 V
V(substrate)	= -18.0 V
V(IDT bias)	= -18.9 V
V(channel plate)	= -5.5 V
V(i/p gate)	= -5.5 V

The output amplifier was configured to have 49 dB gain when matched to 50 ohms at its input and output and was powered by a 9 V battery to reduce power supply noise. The measured 3 dB bandwidth of the amplifier was 70 MHz. Pickup of the 360 MHz SAW drive signal by the output circuit was removed with a simple coaxial stub notch filter.

Figure 4.6: ACT biasing and output amplification circuitry

The MESFET portion of the ACT device is biased into pinch-off under normal operating conditions. Hence, the ACT input and output impedances are large. Viewing the ACT output as a high-impedance current source, the output voltage rise time is controlled by the load resistance and any stray capacitance that might exist. Measurement of the device package capacitance alone with an LCR meter yielded a value of ≈ 0.5 pF. Hence, assuming a capacitance for the output amplifier of 4 pF and using a load resistor of $51\ \Omega$, an RC time constant of ≈ 0.2 ns was calculated. Although the output pad of the ACT was in direct contact with the n-epitaxial layer, the presence of the guard ring should isolate this area. The effective capacitance is the series combination of the depletion layer capacitance and the capacitance of the undepleted n-epitaxial layer to ground through the semi-insulating substrate. Calculation of this latter capacitance yielded a value of ≈ 18 fF, much smaller than the measured package value.

The package plus ACT device output capacitance was quantified using an LCR meter with the guard ring and channel plate voltages set to typical bias values and the bias ring grounded. The measured value was ≈ 0.5 pF, indicating that the output node capacitance was due almost entirely to the package under normal biasing conditions. This capacitance differs with the ≈ 4 pF value quoted by Hoskins [45] page 137, which he attributed mainly to the device bonding pad.

The interdigital transducer was driven with pulses of

360 MHz at drive levels up to 30.5 dBm and typical duty ratios of 5-10%. The pulse response suffered when attempts were made to drive the transducer in a continuous wave manner. There was a time constant of several seconds during which the output pulse rise and fall times were observed to increase dramatically from their minimum values of 10 ns.

It was subsequently found that heating the ACT substrate caused the return loss peak of the IDT to shift down in frequency. The amount of this shift could be calculated taking into account the expansion coefficients of the substrate, IDT metal, and the variation of the SAW velocity over temperature. Webster and Carr [38] have given a temperature coefficient of frequency drift of $-52 \text{ ppm}/^{\circ}\text{C}$. Whether or not this value is sensitive to the the IDT metal chosen is unclear to the present author.

By lowering the generation frequency to account for the IDT temperature rise, it was possible to operate the device at higher duty ratios. Since no attempt was made to heatsink the experimental device, continuous wave (CW) operation was not used. However, CW operation of ACT devices has been demonstrated using RF drive signals up to 33 dBm [55].

4.5 Pulse Response

4.5.1 Electrical Input

Application of a negative pulse to the input ohmic contact injects electrons into the SAW potential wells. Figure 4.7 shows the response of the device when a 1 V input

100 ns/div →

Top trace: ACT input ohmic (1V/div)

Bottom trace: ACT output ohmic ($\approx 4.1 \mu\text{A}/\text{div}$)

Figure 4.7: ACT pulse response for contact injection

pulse was applied. The 700 ns time difference between the input and output pulse agreed well with the value calculated using the known channel length of the device and the SAW velocity. The measured rise and fall time were both 10 ns.

The amount of charge injected, and hence the input impedance of the device, was strongly affected by the substrate, channel plate and input gate voltage for electrical charge injection. This effect was not observed when optical charge injection experiments were attempted.

4.5.2 Optical Input

Using the focussing scheme shown in Figure 4.8a optical pulses were applied to the ACT device. The spot was focussed over each of the four thin chromium windows shown in Figure 4.8b to get uniform charge injection across the SAW beamwidth.

Pulse response of the ACT device under these conditions is shown in Figures 4.9a to 4.9d. The delays of 0.61 ns, 0.52 ns, 0.43 ns and 0.34 ns were in good agreement with values calculated from the window positions and SAW velocity. The retention of pulse shape along the device was an indication of good charge transport efficiency.

Rise and fall times for the optically induced pulses were ≈ 20 ns, double the values for electrical injection. An independent measurement of the optical rise and fall time for the AlGaAs LED using a high speed Si PIN photodiode indicated

Figure 4.8a: Optical pulse generation and focussing

Figure 4.8b: Optical injection window positions

100 ns/div →

(a) : Over Window #1

100 ns/div →

(b) : Over Window #2

Top trace: Voltage to LED circuit

Bottom trace: ACT output current ($\approx 1.5 \mu\text{A}/\text{div}$)

Figure 4.9: Pulse response for optical charge injection

100 ns/div →

(c) : Over Window #3

100 ns/div →

(d) : Over Window #4

Top trace: Voltage to LED circuit

Bottom trace: ACT output current ($\approx 1.5 \mu\text{A}/\text{div}$)

Figure 4.9: Pulse response for optical charge injection

a value of ≈ 10 ns. Other contributions to the rise time include: output ohmic stray capacitance; finite bandwidth of the output amplifier; non-zero charge transfer inefficiency and light leakage into the gap region between the guard ring and channel plate. Any stray light in the channel plate/guard ring gap region will inject carriers beyond the limits defined by semi-transparent windows and degrade the pulse response.

4.5.3 Pulse Response vs. SAW Magnitude and Charge Injection Level

Figures 4.10a to 4.10c illustrate the output pulse waveform as the input RF power to the IDT, and hence the wave potential, was increased for optical charge injection. With low values of SAW potential, a great deal of injected charge was spilled into trailing wells.

Diffusion induced charge transfer inefficiency theory presented by Hoskins [45, 62] predicts an equal amount of charge spillage into both leading and trailing SAW wells, referred to as "ballooning". This symmetric effect was not observed either in the present or previous work [45]. Hoskins [45] page 123 attributed asymmetric behaviour which he observed to small potential wells created in $2\text{ }\mu\text{m}$ surface gaps present in the channel plate of his devices.

Initially it was thought that the asymmetric behaviour with the present device could be explained in terms of a voltage drop across thin chromium windows on the surface of

$$P_{elec} = +15 \text{ dBm}$$

$$(\phi_o \approx 70 \text{ mV})$$

(a) 100 ns/div --->

$$P_{elec} = +18.5 \text{ dBm}$$

$$(\phi_o \approx 110 \text{ mV})$$

(b) 100 ns/div --->

$$P_{elec} = +24 \text{ dBm}$$

$$(\phi_o \approx 200 \text{ mV})$$

(c) 100 ns/div --->

Vertical: ACT output current ($\approx 1.5 \mu\text{A}/\text{div}$)

(Optical charge injection)

Figure 4.10: Pulse response variation with SAW drive power

the device creating a potential hill over which the SAW would need to push transported charge.

Measurement of sheet resistivity for evaporated 25 nm chromium films yielded a value of $\approx 740 \text{ } \Omega/\text{square}$. Assuming a form of the SAW potential given by:

$$\phi_S(x_1, t) = \phi_0 \sin(kx_1 - \omega t) \quad (4.17)$$

where: ϕ_0 - magnitude of SAW potential

(assume no x_3 dependence)

k - acoustic wavenumber

ω - radian frequency

t - time variable

Application of Poisson's equation to ϕ_S given by (4.17) gives the magnitude of the piezoelectric charge density below:

$$\rho_W = \phi_0 \epsilon k^2 \quad (4.18)$$

Using some typical values in (4.18):

$$\phi_0 = 0.3 \text{ Volts}$$

$$\epsilon = 11.0 \epsilon_0$$

$$k = 2\pi/8 \text{ } \mu\text{m}$$

yields:

$$\rho_W = 18 \text{ C/m}^3$$

Since this charge moves at $v_S \approx 2860 \text{ m/s}$ in the x_1 direction, the equivalent current is $5.1 \times 10^4 \text{ A/m}^2$. Assuming an equal image current in the channel plate and a $400 \text{ } \mu\text{m}$ channel width, the channel plate current I_{cp} is calculated as:

$$I_{cp} = 82 \text{ } \mu\text{A}$$

The voltage drop across the chromium window is then 3 mV , much smaller than the $\approx 300 \text{ mV}$ SAW potential. As a result,

the asymmetric charge transfer behaviour must have another cause, not currently understood by the author.

Figures 4.11a to 4.11c describe output pulse response as a function of signal level for electrical charge injection. As the input was increased, charge spilled predominantly into trailing wells.

4.6 Charge Transfer Efficiency

The charge transfer efficiency was estimated by two independent techniques assuming a proportional loss mechanism.

In the first method, the frequency response of the ACT device was measured as RF bursts of increasing frequency were applied to the device input [83]. Figure 4.12a and 4.12b illustrate device response to RF bursts of increasing frequency applied to the device. Amplitude response was measured and the data curve fitted to the theoretical response given by:

$$|G(f)| = \exp(-N_t \epsilon_t [1 - \cos(2\pi f/f_0)]) \quad (4.19)$$

where: ϵ_t = charge transfer inefficiency
 f = RF burst frequency
 N_t = number of equivalent transfers
 f_0 = clock frequency

Using $f_0 = 360.7$ MHz, $N_t = 250$, and $\epsilon_t = 0.008$, a fit to the experimental data shown in Figure 4.13 was achieved. The charge transfer efficiency determined by this method was 0.992.

Input Ohmic Voltage
(-1 Volt)

(a) 100 ns/div →

Input Ohmic Voltage
(-1.15 Volt)

(b) 100 ns/div →

Input Ohmic Voltage
(-1.35 Volt)

(c) 100 ns/div →

Vertical: ACT output current ($\approx 4.1 \mu\text{A}/\text{div}$)

(Contact injection)

Figure 4.11: Pulse shape variation with injection level

200 ns/div →

(a) $f_{\text{BURST}} = 10 \text{ MHz}$

200 ns/div →

(b) $f_{\text{BURST}} = 42 \text{ MHz}$

Top trace: Input ohmic voltage (0.5 V/div)

Bottom Trace: Output ohmic current ($\approx 1.6 \mu\text{A}/\text{div}$)

Charge Load: 1.6×10^6 electrons/cm (for each $8 \mu\text{m}$ well)

Figure 4.12: ACT RF burst response

Figure 4.13: Charge transfer efficiency using frequency response method

An alternate time domain method [84] involved application of a long duration pulse to the input of the device and measurement the rise time and fraction of charge lost from the leading edge of the output pulse. For small charge transfer inefficiency, the expression given below was used [45]:

$$\epsilon_t = (t_r f_0 / N_t) (Q_L / Q_t) \quad (4.20)$$

where: t_r = rise time of the leading edge of the
output pulse

(Q_L / Q_t) = fraction of charge lost from the
leading edge of the output pulse

f_0 and N_t are as previously defined for the frequency response method. Calculation of charge transfer efficiency using this technique yielded a value of 0.993.

The estimates of charge transfer efficiency obtained using these methods were lower limits since the frequency response of the output amplifier will have degraded the output pulse rise time. Both results were close to the value of 0.996 obtained by Hoskins [45] on a similar device for a charge load of $3 \times 10^6 \text{ cm}^{-1}$.

Experiments were not undertaken to determine how charge transport efficiency varied with charge load though these measurements have been reported by Hoskins [45] page 146.

4.7 Device Linearity Characteristics and Input/Output

Impedance

Linearity of the device output current as a function of input ohmic voltage and focussed optical power are given in Figures 4.14 and 4.15 for electrical and optical input signals respectively. The maximum output current measured was constrained by spreading of the output pulse at high input levels into trailing charge wells for contact injection. The maximum light intensity available from the AlGaAs LED limited the amount of charge which could be optically induced.

The linearity of the device for optical input signals was apparent from Figure 4.15 over the range of optical power where measurements were made. The slope indicated an optical responsivity of approximately 5 mA/W. The linearity characteristic for electrical charge injection was similar to that measured by Hoskins [45].

An approximate value for the real part of the device input impedance is given by taking the reciprocal of the slope of the linearity characteristic for electrical injection. This yielded a value of $\approx 26 \text{ k}\Omega$, slightly higher than the 25 $\text{k}\Omega$ value reported by Hoskins. Although not perfectly true, the generally symmetric geometry of the ACT input and output would imply an output impedance of $\approx 26 \text{ k}\Omega$.

Figure 4.14: ACT output linearity for contact injection

Figure 4.15: ACT output linearity for optical injection

4.8 Quantum Efficiency for Optical Signal Injection

The ACT device quantum efficiency for near-infrared incident light was defined by relating the number of electrons detected at the device output to the number of photons incident on the input window.

Using the optical linearity graph (Figure 4.15) output current for an incident focussed optical power of 70 μW was 3.6 μA . The output current of 3.6 μA implied a net generation rate of 2.27×10^{13} electrons/second. Using this information and the LED peak wavelength of emission of 730 nm, the quantum efficiency was calculated using:

$$\eta = ghc / \lambda_p P_{\text{OPT}} \quad (4.21)$$

where: g - net generation rate

h - Planck's constant

c - speed of light in vacuum

λ_p - wavelength of LED light

P_{OPT} - optical power incident on Cr window

The value obtained was $\eta = 0.09$. To determine the device "intrinsic" quantum efficiency, the optical loss and reflection due to the chromium windows was calculated using the analysis presented by Schneider [85].

To perform the analysis, complex refractive indices of GaAs, Cr, and Air were required. Values used were: $N = 1$ for air; $N = 1.5 - j2.0$ for chromium [86] and $N = 3.74 - j0.111$ for GaAs [61]. The chromium layer thickness was 25 nm. Optical experiments used a Fujitsu AlGaAs LED (Part No. FED073K1WA)

with peak emission wavelength of 730 nm and spectral halfwidth of 25 nm. For the calculation, the light was assumed monochromatic. The calculated transmittance and reflectance were 0.37 and 0.46 respectively. Details of the calculation are given in Appendix B. Hence, with only $\approx 37\%$ of incident optical power reaching the GaAs, the recalculated quantum efficiency became 0.24.

Unfortunately, the accuracy of this result is debatable since the optical constants quoted in the literature for chromium vary widely. The refractive index for chromium is sensitive to: rate of deposition of the film; substrate temperature during deposition; vacuum pressure; thickness of the film and the frequency of light [87, 88]. The value used in the analysis was for a chromium film evaporated under similar experimental conditions as the present work.

The optical constants for chromium were measured at a wavelength of 546 nm. There is evidence that the optical constants for chromium films increase with frequency in the 500 nm to 1000 nm wavelength range [88]. The magnitude of this increase is of the order of 10% between 546 nm and 730 nm. Use of larger values for the refractive index of chromium improves the calculated quantum efficiency.

5.0 CONCLUSION

An ACT device was designed and fabricated. Interface circuitry and test fixtures were constructed allowing electrical and optical charge injection experiments to be undertaken. Results obtained have established an optical charge injection technique of possible use in future frontside illuminated ACT imaging devices.

Electron injection was achieved using an ohmic contact as described by Hoskins et al. [9, 45, 63]. Experimental results were in general agreement with the previous work. In particular, the ACT charge transfer efficiency for small SAW duty ratios ($\approx 5\%$) was in excess of 0.992 for a charge load of 1.6×10^6 electrons/cm. This compares to a value of 0.996 quoted by Hoskins et al. for a charge load of 4×10^6 electrons/cm and SAW potential of ≈ 0.32 V. The pulse shape was found to be a strong function of: device biasing; SAW potential; injection level and the RF duty ratio of the transducer. Minimum pulse rise and fall times were 10 ns, similar to results obtained by Hoskins [45] though smaller charge loads were achieved with the present device.

Optical charge injection through thin semi-transparent windows on the surface of the ACT was demonstrated for near-infrared light ($\lambda_p = 730$ nm). The injection process involved the creation and separation of electron-hole-pairs, with subsequent transport of the electrons in potential wells defined by the SAW. The presence of holes in the channel and the dynamics of their removal represented a physical

situation fundamentally different from the case of contact injection reported in the past.

The net quantum efficiency of the device built for this work was $\approx 9\%$. Calculations of the reflection and loss due to thin chromium windows on a GaAs substrate yielded an "intrinsic" quantum efficiency in excess of 24% . A precise determination was not possible due to the wide range of optical constants quoted in the literature for chromium.

Optical pulse responses had rise and fall times of ≈ 20 ns. A portion of the rise and fall times was directly attributed to the response of the AlGaAs LED used to provide the near-infrared light and leakage of the focussed optical signal into the gap between the channel plate and guard ring. Since the optical penetration depth at $\lambda_p = 730$ nm was approximately the same as the $2.2 \mu\text{m}$ thick epitaxial layer, very few electron-hole-pairs were created in the substrate. Hence, "tails" were not visible on the optically induced output pulse as might have been expected from electron trapping effects in the semi-insulating substrate. The pulse response shape and magnitude over the four semi-transparent windows was well preserved. ACT output current followed a linear relationship with respect to focussed optical power on the semi-transparent windows over the optical power range measured.

Several key issues need to be addressed in the design of an ACT imaging device based on the injection technique described. The problem of low external quantum efficiency

could presumably be overcome using a more sophisticated semi-transparent channel plate electrode, keeping the acoustic properties of the chosen material in mind. SAW scattering, diffraction, and beam steering would play a prominent role in determining the maximum practical imaging area. More work is needed to establish the best epitaxial wafer growth technique which will provide the lowest SAW attenuation.

Once launched, the SAW potential wells cannot be slowed down or stopped. A method is therefore required to integrate enough charge before the acoustic pixel has moved one wavelength. To prevent distortion or smearing of the image, the charge wells would need to be shielded from additional radiation until clocked out of the device.

Large amplitude RF pulses (> 24 dBm) were required to launch a SAW of sufficient wave potential to drive the charge transport process. To avoid breakdown of the transducer, continuous wave operation was not used. More work is required to achieve CW operation. One possible idea for improvement would be to use a more efficient IDT. A split-finger design appears promising based on its low-reflectivity properties. Use of a reflective grating to redirect the SAW energy propagating away from the charge transport channel would improve the IDT generation efficiency by approximately 3 dB.

Assuming, therefore, that the total IDT conversion loss could be reduced from ≈ 7 dB to ≈ 4 dB, a 1 Volt SAW wave potential would require an input RF drive signal amplitude of

≈ 29 dBm for a channel width of $600\ \mu\text{m}$. The level of this signal presents a challenging thermal dissipation problem for the device operated in CW mode. ACT devices have been operated with continuous wave drive signals up to $+33$ dBm [55]. Details of the thermal design used to achieve this drive level have not been published.

Compared to an ACT device, a four-phase CCD with 1 V clock voltages has smaller RF power requirements. To generate a multi-phase clock signal for the CCD, one might use a 180 degree hybrid to split a sinusoidal source into two equal power components followed by two 90 degree hybrids to generate four phases. Assuming a loss of 1 dB for each hybrid, approximately 18 dBm of drive power would be required for this CCD in a 50 ohm system.

Clearly, the much higher RF drive magnitude requirements for the ACT and associated thermal dissipation problems must be balanced against the intrinsic simplicity of its drive circuitry. A narrowband RF amplifier can be used to provide a single-phase clock signal for the ACT. The CCD will require additional circuitry for the multi-phase drive signal. Better charge transfer efficiency is achieved for CCD devices when they are driven with "square" clock signals since this increases the fringing fields under the gates. Obtaining such clock waveforms at VHF or UHF frequencies further increases the complexity of the CCD drive circuitry over that required by the ACT.

The wide channel width ($400\ \mu\text{m}$) of the device built for

this research made it possible for the relatively low SAW potential of ≈ 0.3 Volts to carry enough charge per well for simple detection of transported charge at the output. As previously discussed, wave potentials in excess of 0.26 V provide approximately linear increases in the charge capacity. Hence, an ACT device with a $10\text{ }\mu\text{m}$ channel width and 1.6 V wave potential could achieve the same maximum charge capacity as a device with a $400\text{ }\mu\text{m}$ channel width and 0.3 V wave potential. This would be an important consideration in the design of a two-dimensional imaging array in which narrow channel widths are desired for resolution in the transverse direction.

Future experiments will be required to establish the feasibility of ACT as a device capable of resolving ionizing radiation. Detection of the small number of injected electrons created by a single nuclear event will involve redesign of the output circuitry to minimize noise. Fortunately, typical charge loads will be small. Therefore, the requirement for a sophisticated transducer to generate a large amplitude SAW is reduced. Device degradation due to long-term radiation exposure and transient radiation effects discussed briefly in chapter 1 will be of considerable importance in this application.

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APPENDIX A: Calculation of the Quasistatic Potential for

<110> Propagating SAW on (100) cut GaAs

Quasistatic potential and displacement components have been calculated for <110> propagating SAW on (100) cut GaAs using the method of Tseng and White [42] and Campbell and Jones [43]. The method will be outlined below.

Relevant tensor equations are given, assuming Einstein summation over all subscripts, and orientation shown in Figure 2.1:

Elastic-wave equation:

$$\frac{dT_{ij}}{dx_j} = \frac{\rho_m d^2 u_i}{dt^2} \quad (A.1)$$

where: ρ_m - mass density

T_{ij} - stress tensor

u_i - particle displacement in the i'th direction

x_j - space coordinate

t - time variable

Strain-Displacement Relationship:

$$S_{kl} = \frac{1}{2} \left[\frac{du_k}{dx_l} + \frac{du_l}{dx_k} \right] \quad (A.2)$$

where: S_{kl} - strain tensor

Quasistatic Potential:

$$E_k = \frac{-d\phi_w}{dx_k} \quad (A.3)$$

where: E_k - electric field

ϕ_w - quasistatic potential

Constitutive Relations:

$$T_{ij} = c_{ijkl}^E S_{kl} - e_{kij} E_k \quad (A.4)$$

$$D_i = e_{ikl} S_{kl} + \epsilon_{ik}^S E_k \quad (A.5)$$

where: c_{ijkl}^E = stiffness coefficients

ϵ_{ik}^S = permittivity coefficients

e_{ijk} = piezoelectric coefficients

D_i = electric displacement

Gauss' Law (no uncompensated charge):

$$\frac{dD_i}{dx_i} = 0 \quad (A.6)$$

Substitution of (A.2) and (A.3) into (A.4) and (A.5) yields:

$$T_{ij} = c_{ijkl}^E \frac{1}{2} \left[\frac{du_k}{dx_l} + \frac{du_l}{dx_k} \right] + e_{ijk} \frac{d\phi_w}{dx_k} \quad (A.7)$$

$$D_i = e_{ikl} \frac{1}{2} \left[\frac{du_k}{dx_l} + \frac{du_l}{dx_k} \right] - \epsilon_{ik}^S \frac{d\phi_w}{dx_k} \quad (A.8)$$

Substitution of (A.7) and (A.8) into (A.1) and (A.6) gives:

$$c_{ijkl}^E \frac{1}{2} \left[\frac{d^2 u_k}{dx_l dx_j} + \frac{d^2 u_l}{dx_k dx_j} \right] + e_{ijk} \frac{d^2 \phi_w}{dx_k dx_j} = \rho_m \frac{d^2 u_i}{dt^2} \quad (A.9)$$

$$e_{ikl} \frac{1}{2} \left[\frac{d^2 u_k}{dx_l dx_i} + \frac{d^2 u_l}{dx_k dx_i} \right] - \epsilon_{ik}^S \frac{d^2 \phi_w}{dx_k dx_i} = 0 \quad (A.10)$$

$$\text{Since: } c_{ijkl}^E = c_{ijlk}^E \quad (A.11)$$

$$\text{and: } e_{ijk} = e_{ikj} \quad (A.12)$$

equations (A.9) and (A.10) can be simplified:

$$c_{ijkl}^E \frac{d^2 u_l}{dx_k dx_j} + e_{ijk} \frac{d^2 \phi_w}{dx_k dx_j} = \rho_m \frac{d^2 u_i}{dt^2} \quad (A.13)$$

$$e_{ikl} \frac{d^2 u_l}{dx_k dx_i} - \epsilon_{ik}^S \frac{d^2 \phi_w}{dx_k dx_i} = 0 \quad (A.14)$$

Applying abbreviated subscript rules [89] equations (A.13) and (A.14) can be written for the cubic GaAs crystal class after modification of the stiffness and piezoelectric constants to apply along the $\langle 110 \rangle$ propagation direction. The modification involves a rotation of the x_1 and x_2 axes by 45 degrees. The modified constants are given below (primed) in terms of the unrotated values (unprimed) [90]:

$$[c'] = \begin{bmatrix} c'_{11} & c'_{12} & c'_{12} & 0 & 0 & 0 \\ c'_{12} & c'_{11} & c'_{12} & 0 & 0 & 0 \\ c'_{12} & c'_{12} & c'_{11} & 0 & 0 & 0 \\ 0 & 0 & 0 & c_{44} & 0 & 0 \\ 0 & 0 & 0 & 0 & c_{44} & 0 \\ 0 & 0 & 0 & 0 & 0 & c'_{66} \end{bmatrix} \quad (A.15a)$$

$$c'_{11} = \frac{(c_{11} + c_{12} + 2c_{44})}{2} \quad (A.15b)$$

$$c'_{12} = \frac{(c_{11} + c_{12} - 2c_{44})}{2} \quad (A.15c)$$

$$c'_{66} = \frac{(c_{11} - c_{12})}{2} \quad (A.15d)$$

$$[e'] = \begin{bmatrix} 0 & 0 & 0 & 0 & e_{14} & 0 \\ 0 & 0 & 0 & -e_{14} & 0 & 0 \\ e_{14} & -e_{14} & 0 & 0 & 0 & 0 \end{bmatrix} \quad (A.16)$$

Using equations (A.15) and (A.16), equations (A.13) and (A.14) are rewritten in abbreviated subscript notation where for the Rayleigh wave solution, all terms involving spatial derivatives with respect to x_2 have been dropped:

$$c_{11} \frac{d^2 u_1}{dx_1^2} + c_{55} \frac{d^2 u_1}{dx_3^2} + (c_{13} + c_{55}) \frac{d^2 u_3}{dx_1 dx_3} + 2e_{15} \frac{d^2 \phi_w}{dx_1 dx_3} - \rho_m \frac{d^2 u_1}{dt^2} \quad (A.17)$$

$$c_{66} \frac{d^2 u_2}{dx_1^2} + c_{44} \frac{d^2 u_2}{dx_3^2} - \rho_m \frac{d^2 u_2}{dt^2} \quad (A.18)$$

$$(c_{13} + c_{55}) \frac{d^2 u_1}{dx_1 dx_3} + c_{55} \frac{d^2 u_3}{dx_1^2} + c_{33} \frac{d^2 u_3}{dx_3^2} + e_{15} \frac{d^2 \phi_w}{dx_1^2} - \rho_m \frac{d^2 u_3}{dt^2} \quad (A.19)$$

$$2e_{15} \frac{d^2 u_1}{dx_1 dx_3} + e_{15} \frac{d^2 u_3}{dx_1^2} - \epsilon \left[\frac{d^2 \phi_w}{dx_1^2} + \frac{d^2 \phi_w}{dx_3^2} \right] = 0 \quad (A.20)$$

Assuming a partial wave solution of the form:

$$\begin{bmatrix} u_1 \\ u_2 \\ u_3 \\ \phi_w \end{bmatrix} = \begin{bmatrix} U_1 \\ U_2 \\ U_3 \\ V \end{bmatrix} \exp(-\sigma k x_3) \exp(j k x_1 - j \omega t) \quad x_3 > 0 \quad (A.21)$$

$$\phi_w = V_0 \exp(k x_3) \exp(j k x_1 - j \omega t) \quad x_3 < 0 \quad (A.22)$$

where: ω = radian frequency of SAW

$$k = \omega/v_S \text{ is the acoustic wavenumber} \quad (\text{A.23})$$

and: v_S - SAW velocity

Substitution of (A.21) into (A.17)-(A.20) gives a matrix equation involving U_1 , U_2 , U_3 , and V .

$$\begin{bmatrix} (c_{55}\sigma^2/v_S^2 - c_{11}/v_S^2 + \rho_m) & 0 & j(c_{13}+c_{55})\sigma/v_S^2 & j2e_{15}\sigma/v_S^2 \\ 0 & (c_{44}\sigma^2/v_S^2 - c_{66}/v_S^2 + \rho_m) & 0 & 0 \\ j(c_{13}+c_{55})\sigma & 0 & (c_{33}\sigma^2 - c_{55} + \rho_m v_S^2) & -e_{15} \\ j2e_{15}\sigma & 0 & -e_{15} & \epsilon(1-\sigma^2) \end{bmatrix} \begin{bmatrix} U_1 \\ U_2 \\ U_3 \\ V \end{bmatrix} = 0 \quad (\text{A.24})$$

The u_2 solution is decoupled from the sagittal plane solution involving the u_1 and u_3 displacement components. The Rayleigh wave is of interest, involving the constants U_1 , U_3 , and V . Denoting the matrix coefficients in (A.24) that involve U_1 , U_3 , and V as A , we can write:

$$A \begin{bmatrix} U_1 \\ U_3 \\ V \end{bmatrix} = 0 \quad (\text{A.25a})$$

where:

$$A = \begin{bmatrix} (c_{55}\sigma^2/v_S^2 - c_{11}/v_S^2 + \rho_m) & j(c_{13}+c_{55})\sigma/v_S^2 & j2e_{15}\sigma/v_S^2 \\ j(c_{13}+c_{55})\sigma & (c_{33}\sigma^2 - c_{55} + \rho_m v_S^2) & -e_{15} \\ j2e_{15}\sigma & -e_{15} & \epsilon(1-\sigma^2) \end{bmatrix} \quad (\text{A.25b})$$

For a non-trivial solution, $\det[A] = 0$. When the determinantal equation is expanded, a cubic in σ^2 results. For solutions that decay into the crystal, the roots with positive real parts are chosen. The three roots are denoted σ_y , $y=1,2,3$. For each σ_y , $y=1,2,3$ there will be an eigenvector normalized to U_1 that satisfies:

$$[A] \begin{bmatrix} 1 \\ a_y \\ b_y \end{bmatrix} = 0 \quad y = 1,2,3 \quad (\text{A.26a})$$

$$\text{where: } a_1 = U_3/U_1 \quad (\text{A.26b})$$

$$b_1 = V/U_1 \quad (\text{A.26c})$$

The solution of interest can be expressed in terms of the eigenvector components (a_y, b_y) and weighting factors (W_y) that are determined by the boundary conditions. The form of the solutions are as below:

For $x_3 > 0$:

$$u_1 = \sum_{y=1}^3 \{W_y \exp(-\sigma_y k x_3) \exp(j k x_1 - j \omega t)\} \quad (\text{A.27})$$

$$u_3 = \sum_{y=1}^3 \{a_y W_y \exp(-\sigma_y k x_3) \exp(j k x_1 - j \omega t)\} \quad (\text{A.28})$$

$$\phi_w = \sum_{y=1}^3 \{b_y W_y \exp(-\sigma_y k x_3) \exp(j k x_1 - j \omega t)\} \quad (\text{A.29})$$

For $x_3 < 0$:

$$\phi_w = V_0 \exp(k x_3) \exp(j k x_1 - j \omega t) \quad (\text{A.30})$$

The solution for a shorted surface is of primary interest for ACT operation since the channel plate on the surface of the device provides this condition. The boundary conditions at an electrically shorted surface are given by:

$$\phi_w(x_3 = 0) = 0 \quad (\text{A.31})$$

$$T_{31}(x_3 = 0) = 0 \quad (\text{A.32})$$

$$T_{33}(x_3 = 0) = 0 \quad (\text{A.33})$$

From (A.31) and (A.29):

$$\sum_{y=1}^3 b_y W_y = 0 \quad (\text{A.34})$$

From (A.2), (A.3), and (A.4):

$$T_{31} = c_{55} \left[\frac{du_3}{dx_1} + \frac{du_1}{dx_3} \right] + e_{31} \frac{d\phi_w}{dx_1} \quad (\text{A.35})$$

and:

$$T_{33} = c_{31} \frac{du_1}{dx_1} + c_{33} \frac{du_3}{dx_3} \quad (\text{A.36})$$

Substituting (A.27)-(A.29) into (A.35) and (A.36) and setting these expressions equal to zero in accordance with equations (A.32)-(A.33) yields:

$$\sum_{y=1}^3 \{W_y [c_{55}(ja_y + \sigma_y) + je_{31}b_y]\} = 0 \quad (\text{A.37})$$

$$\sum_{y=1}^3 \{W_y [jc_{31} + c_{33}a_y\sigma_y]\} = 0 \quad (\text{A.38})$$

Equations (A.34), (A.37), and (A.38) together form a matrix equation for W_y :

$$\begin{bmatrix} b_1 & b_2 & b_3 \\ c_{55}(ja_1+\sigma_1) + je_{31}b_1 & c_{55}(ja_2+\sigma_2) + je_{31}b_2 & c_{55}(ja_3+\sigma_3) + je_{31}b_3 \\ c_{33}a_1\sigma_1 + jc_{13} & c_{33}a_2\sigma_2 + jc_{13} & c_{33}a_3\sigma_3 + jc_{13} \end{bmatrix} \begin{bmatrix} W_1 \\ W_2 \\ W_3 \end{bmatrix} = 0 \quad (\text{A.39})$$

Equation (A.39) is of the form:

$$[D] \begin{bmatrix} W_1 \\ W_2 \\ W_3 \end{bmatrix} = 0 \quad (\text{A.40})$$

For a non-trivial solution:

$$\det[D] = 0 \quad (\text{A.41})$$

The method of solution follows the sequence below:

- 1) Assume a value of v_s
- 2) Calculate σ_y such that $\det[A] = 0$ using (A.24)
- 3) Calculate a_y and b_y from (A.26)
- 4) Check if $\det[D] = 0$. If not, change v_s and repeat steps 1) to 3).

Once the SAW velocity has been obtained by this method, the displacement components and quasistatic potential can be calculated from (A.27)-(A.29) after solving for W_2 and W_3 in terms of an arbitrary constant W_1 using (A.39). For the numerical calculation, $W_1 = 1$ was used. Denormalization for arbitrary acoustic power takes place at a later stage in the calculation.

The acoustic power flow per unit area is given by Poynting's theorem for acoustic waves using the quasistatic approximation:

$$\bar{P}_{avg} = 1/2 \operatorname{Re}[-\bar{v}^* \bar{T} + \phi_w(j\omega \bar{D})^*] \quad (A.42)$$

$$\text{where: } v_i = \frac{du_i}{dt} \quad (A.43) \text{ is the particle velocity}$$

In the x_1 direction, the power flow is given by:

$$P_{x1} = -1/2 \omega \operatorname{Re}[j(u_1^* T_{11} + u_3^* T_{31} + \phi_w D_1^*)] \quad (A.44)$$

Noting that:

$$T_{11} = c_{11} \frac{du_1}{dx_1} + c_{13} \frac{du_3}{dx_3} + e_{31} \frac{d\phi_w}{dx_3} \quad (A.45)$$

and:

$$D_1 = e_{15} \left[\frac{du_1}{dx_3} + \frac{du_3}{dx_1} \right] - \epsilon \frac{d\phi_w}{dx_1} \quad (A.46)$$

After substitution of (A.27)-(A.29), (A.35), (A.45), and (A.46) into (A.44) and integration from $x_3 = 0$ to $x_3 = \infty$ yields the expression for the acoustic power flow in units of Power per unit Acoustic Beamwidth:

$$P_{AI} = \operatorname{Re} \left[\sum_{y=1}^3 \frac{-j\omega}{4\operatorname{Re}(\sigma_y)} |W_y|^2 (jc_{11} + c_{13}\sigma_y a_y + e_{31}\sigma_y b_y + a_y^* (jc_{55} a_y + c_{55}\sigma_y) + b_y (e_{15}\sigma_y^* + j\epsilon b_y^*)) \right] \quad (A.47)$$

APPENDIX B: Transmission of Light through Chromium onto GaAs

To evaluate the reflection and loss due to the chromium "windows" on the ACT, the method presented by Schneider [85] has been used. Figure B.1 shows the substrate configuration under analysis. Incident light on the chromium window is partially reflected (component r_2) and partially transmitted to the GaAs (component t_2).

Generalized immittances for each (non-magnetic) layer are given by:

$$g_k = N_k \cos \gamma_k \quad (\text{B.1}) \quad k = 0, 1, 2$$

for a "p" (TE) wave

$$g_k = \frac{\cos \gamma_k}{N_k} \quad (\text{B.2}) \quad k = 0, 1, 2$$

for an "s" (TM) wave

To simplify the analysis, the case of normal incidence will be considered. In actual fact, the lens focus system used to provide the light for the optical experiments gave $0^\circ \leq |\gamma_0| < 8^\circ$. With normal incidence ($\gamma_0 = \gamma_1 = \gamma_2 = 0$), and the expressions for transmittance and reflectance are identical for TE and TM incident waves. Hence, only the TE solution will be considered. Pertinant equations used in the analysis are given below:

Phase constant for chromium film:

$$\rho_1 = \frac{j2\pi N_1 \cos \gamma_1}{\lambda_p} \quad (\text{B.3})$$

where: λ_p = free space optical wavelength

Figure B.1: Optical model of thin chromium window on GaAs used in calculation of quantum efficiency

Reflection coefficient for air-chromium interface:

$$r_2 = \frac{(g_2 - g_1)(g_1 + g_0)\exp(\rho_1 d) + (g_2 + g_1)(g_1 - g_0)\exp(-\rho_1 d)}{(g_2 + g_1)(g_1 + g_0)\exp(\rho_1 d) + (g_2 - g_1)(g_1 - g_0)\exp(-\rho_1 d)} \quad (\text{B.4})$$

Reflectance of air-chromium interface:

$$R = r_2 r_2^* \quad (\text{B.5})$$

Transmission coefficient for chromium-GaAs interface:

$$t_2 = \frac{4g_1 g_2}{(g_1 + g_2)(g_1 + g_0)\exp(\rho_1 d) + (g_2 - g_1)(g_1 - g_0)\exp(-\rho_1 d)} \quad (\text{B.6})$$

Transmittance of chromium-GaAs interface:

$$T = \frac{|g_0|}{|g_2|} t_2 t_2^* \quad (\text{B.7})$$

Fraction of incident optical power lost in the chromium film:

$$L_{Cr} = 1 - R - T \quad (\text{B.8})$$