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Studies in Solid State Image Sensors

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IMAGE SENSORS

BY

KONSTANTINOS NIKOLAOU ANAGNOSTOPOULOS

A THESIS SUBMITTED IN PARTIAL FULFILLMENT
OF THE REQUIREMENTS FOR THE DEGREE OF
DOCTOR OF PHILOSOPHY
IN
ELECTRICAL ENGINEERING

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UNIVERSITY OF RHODE ISLAND

1975

ABSTRACT

DOCTOR OF PHILOSOPHY DISSERTATION

OF

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ABSTRACT

16 x 16 element silicon image sensor arrays, utilizing the photodiode-bipolar transistor (PBT) element, were fabricated, tested and scanned in a television raster. Area arrays utilizing the PBT element are simple to fabricate requiring only two diffusions and one metallization. The spike noise cancellation scheme inherent in the PBT element was effectively utilized in the area arrays to reduce the peak value of the spike noise by more than an order of magnitude and to increase the peak value of the video signal.

The circuit analysis was performed using ECAP-1620. The photodiode was modeled by a non-linear capacitor and the transistor by a modified π -equivalent circuit. The variation of the base-emitter junction resistance and voltage drop with forward bias and the variation of the transistor gain with base current were included in the model. The effect of all the elements of the array on the video signals was taken into account. Very good agreement was obtained between the calculated and experimental results.

The collection process of the photogenerated charge in charge injection device (CID) image sensors was examined and a method is presented by which the collection efficiency can be calculated. Numerical techniques were utilized in the analysis. The method was applied to a specific epitaxial type structure. The contribution of the depleted and undepleted regions to the total collection efficiency of the

sensor was calculated.

The relation between output signal charge and incident light intensity was also investigated. It is shown that the collection efficiency decreases with increasing signal charge and this results in non-linear transfer characteristics at high signal levels.

A design is presented for a CID type image sensor that can be operated in the back illumination mode. Except for the thinning step, the proposed structure is as simple to fabricate as present designs.

Calculations are also presented of the transmittance of $\text{Air/SiO}_2/\text{Polysilicon/SiO}_2/\text{Si}$ structures in the spectral region between 0.4 and $1.0\ \mu\text{m}$. It is shown that a judicious choice of the thicknesses of the oxide films, for a given polysilicon film thickness, can substantially increase the transmittance over a narrow wavelength band or over the entire wavelength region of interest here.

The dependence of the photocurrent on the energy of the incident photons was investigated in silicon p-n junctions at room and higher temperatures for photon energies between 0.75 eV and 1.08 eV. The sensitivity of the method enabled high resolution measurements in the absorption tail. At room temperature, thresholds were found at ~ 0.91 eV, 0.99 eV, and 1.026 eV. The derivative of the response showed extensive fine structure in this tail. The TO and LO phonon assisted transitions to the ground and excited state of the exciton, reported by Shaklee and Nahory in the phonon emission region, were seen here with phonon absorption, occurring around 1.054 eV and 1.065 eV.

PREFACE

This thesis is prepared in accordance with the new option -- "Manuscript plan for thesis preparation", approved on November 22, 1968 by the Graduate Council of the University of Rhode Island. The main body of this thesis is comprised of Manuscripts I - IV which have been written in the contemporary style required for Manuscripts to be published in the IEEE J. of Solid State Circuits, Solid State Electronics, Physical Review or other standard journals. An overall introduction and review of the problem is given in Chapter V. The references pertaining to each Manuscript may be found at the end of the respective Manuscripts. A complete list of all references cited, in order of appearance in the thesis, is given in Chapter VI. A number of Appendices are written as supplements to the Manuscripts and are placed at the end of their corresponding Manuscript.

The following Manuscripts are already published or have been accepted for publication:

- IV. Phys. Rev. B-7, 733 (1973);
- III. IEEE J. Solid State Circuits SC-10, (1975);
- II. Solid State Electronics, (1975).

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Finally, the author wishes to express his sincere appreciation to his wife, Eleni, for her understanding and encouragement and for her careful typing of this thesis.

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Manuscript IV: Fig.

1. Abstract

16 x 16 element silicon image sensor arrays, utilizing the photoconductive bipolar transistor (PBT) element, have been fabricated, tested and scanned in a television raster. These arrays utilizing the PBT element are simple to fabricate requiring two diffusions and one metallization. Also, the spike noise cancellation scheme inherent in the PBT element is effectively utilized in the array design to reduce the peak value of the spike noise by more than an order of magnitude while at the same time increasing the peak value of the video signal.

The circuit analysis is performed using SPICE-630. The photoconductor is modeled by a non-linear capacitor and the transistor by a modified Ebers-Moll circuit. The variation of the base-emitter junction resistance and voltage drop with time and the variation of the transistor gain with base current are included in the model. The effect of all the elements of the array on the video signal is taken into account. Very good agreement is obtained between the calculated and experimental results.

PHOTODIODE - BIPOLAR TRANSISTOR IMAGE SENSOR

1. Abstract

16 x 16 element silicon image sensor arrays, utilizing the photodiode-bipolar transistor (PBT) element, have been fabricated, tested and scanned in a television raster. Area arrays utilizing the PBT element are simple to fabricate requiring two diffusions and one metallization. Also, the spike noise cancellation scheme inherent in the PBT element is effectively utilized in the area arrays to reduce the peak value of the spike noise by more than an order of magnitude while at the same time increasing the peak value of the video signal.

The circuit analysis is performed using ECAP-1620. The photodiode is modeled by a non-linear capacitor and the transistor by a modified π -equivalent circuit. The variation of the base-emitter junction resistance and voltage drop with bias and the variation of the transistor gain with base current are included in the model. The effect of all the elements of the array on the video signals is taken into account. Very good agreement is obtained between the calculated and experimental results.

2. Introduction

Integrated solid state matrix type image sensors are presently of considerable interest and importance. Large silicon area arrays have been fabricated^{1,2} utilizing the phototransistor structure at each matrix point. We have fabricated 16 x 16 element arrays in which the phototransistor is replaced by the photodiode-bipolar transistor (PBT) structure^{3,4}. Although the photodiode-bipolar transistor element is very similar in many aspects to the phototransistor, utilization of the PBT element results in simplified fabrication process requiring only two diffusions and one metallization for an area array. Furthermore, the spike-noise cancellation scheme inherent in the PBT element, has been effectively utilized in the area arrays to reduce the peak value of the spike noise by more than an order of magnitude, while at the same time increasing the peak value of the video signal.

The operation of a single phototransistor element in the charge storage mode and the limitations of the p-n junction as a switch have been extensively studied^{5,6,7,8}. Calculations of the spike noise and video signal in area arrays have been, however only very approximate¹. In phototransistor, as well as PBT area arrays, the scanning pulses produce additional spikes across the load resistor, because of the capacitive paths formed by the elements of the array, other than the one being addressed. In addition, the load resistors are shunted by equivalent capacitances which are

again due to the other elements of the array. The shunt capacitances reduce the peak value of the video signal and limit the speed of operation of the sensor.

In this paper we show that the PBT array can be modeled so that circuit analysis can be performed using a simple electronic circuit analysis program, ECAP-1620⁹. We find good agreement between the characteristics of the calculated signals and those obtained experimentally from the array. Phototransistor arrays can be similarly modeled.

3. Description of the Image Sensor

A. Operation of a Single Element in the Charge Storage Mode

The equivalent circuit of one Photodiode-Bipolar Transistor sensor element is shown in Fig. 1. C_d , C_{bc} and C_{be} are the junction capacitances of the photodiode to base, the base to collector and the base to emitter p-n diodes, respectively. R_b denotes the base resistance. R_E and R_C are the load resistances. V_S is a pulse applied periodically to read the charge collected during the previous integration period and to reset the element for the next integration period. V_C is the collector bias supply. The principle of operation of the PBT element is exactly the same as that of the phototransistor^{6,7}, thus only a brief account will be given here.

The operation of the device can be understood with the aid of Fig. 2. In Fig. 2a, V_S is plotted as function of time. The broken line on the time axis indicates that the integration time T_i is much longer than the sampling

Fig. 1. Equivalent circuit for one PBT element. As connected, the element operates in the charge storage mode.

Fig. 2. a) Addressing pulse. b) Base potential as function of time. c) Emitter video signal. d) Collector video signal. The spikes in the outputs due to capacitive coupling between input and output are also shown.

(addressing) time T_S . T_r designates the risetime and fall-time of the addressing pulse. The base potential v_B as function of time is plotted in Fig. 2b. The video signal appears across the emitter and collector load resistances. Fig. 2c shows the emitter video and Fig. 2d shows the collector video signal.

At the end of the integration period, time t_1 , in Fig. 2a, the base potential is negative or at most a few tenths of a volt positive depending on the incident light intensity during the integration period. As V_S starts to increase, at t_1 , v_B also increases. The transistor will not turn-on however until v_B becomes sufficiently positive. Thus between t_1 and until the transistor turns-on a voltage will appear across R_E and R_C due to the capacitive coupling between input and output. This part of the video signal is termed the spike and is so noted in Figs. 2c and 2d.

With the transistor on, C_d begins charging towards a reverse bias of V_D volts. The charging current for C_d is the base current of the transistor. The time constant for charging C_d , neglecting the emitter diode resistance, is thus given by $(\beta + 1)C_d R_E$, where β is the transistor current gain. The relationship between the peak value of the emitter signal and the charge collected during the integration period has been studied in detail for the phototransistor by Joy and Linvill⁶ and Brugler et al⁷. Their analyses are exactly applicable to the PBT element and will not be reproduced here. The analysis showed that the relationship between collected charge and peak value of v_E

was not exactly linear. However, the alternative, of taking as signal the total charge that flows through R_E during the sampling period is a worse choice since this charge is exactly proportional to β . It is well known that uniformity in β is almost impossible to achieve in large area arrays and, furthermore, β is strongly dependent on the base current¹⁰.

If the sampling period is sufficiently long, v_B , v_E and v_C will be close to zero volts at t_3 . As V_S begins to decrease, v_B is driven negative and the transistor turns-off. Between t_3 and t_4 , a negative going spike will be produced across R_E and R_C again because of the capacitive coupling. In Fig. 2c and 2d it is observed that while the spikes of the emitter video and the collector video are in phase, relative to V_S , the signal peaks are out of phase. Thus if v_C is inverted and multiplied by a suitable constant and added to v_E , the spikes will cancel while the signals will add. This is one of the major advantages of the PBT over the phototransistor.

During the integration period those electron-hole pairs created by the light, either in the base region or outside it, that reach the depletion region edge of any of the three p-n junctions are collected and neutralize an equal amount of depletion charge in the base. During the sampling period the neutralized charge is replaced. The charge that flows in the outside circuit during T_S is, as mentioned above, $(\beta + 1)$ times larger because of the transistor gain.

B. The PBT Area Array

A schematic representation of a photodiode-bipolar transistor image sensor area array and simplified block diagrams of the scanning circuits and video amplifier are shown in Fig. 3. Reading and resetting of an element is achieved by putting the row buss line, to which its emitter is connected, to ground via the load resistor R_E and then connecting the column buss line, to which its photodiode is connected, to $+V_D$. To prevent crosstalk all the other emitter lines must be held at $+V_E$. The photodiode column lines are normally grounded. Random addressing of any one element or any portion of the array is possible. In imaging applications, addressing is normally done sequentially. For example, the first emitter line is grounded through R_E and then each of the photodiode columns are successively put to V_D for a time interval equal to the sampling period desired and then grounded. After all the elements in the first row have been addressed, the first emitter line is returned to $+V_E$ and the second emitter line is grounded through R_E . The photodiode columns are then again addressed sequentially, and so on until each of the elements of the array have been addressed, and then the process is repeated.

A top view and cross sectional views of a small portion of the PBT array fabricated on silicon is shown in Fig. 4. The base is diffused first and driven-in. The photodiode and emitter diffusions follow and are done simultaneously. The area of the emitter diode is much smaller than that of the photodiode to keep the emitter spike small. The base

Fig. 3. PBT area array image sensor and simplified block diagram of the scanning circuits and of the video amplifier.

Fig. 4. Top view and cross sectional views of a portion of a PBT area array integrated structure.

collector junction capacitance is much smaller than the photodiode junction capacitance since the impurity concentration in the base-collector junction region is much less than that in the base-photodiode junction region and since the collector is strongly reverse biased. Thus the collector spike is kept small. Isolation between the column and row buss lines is achieved simply by utilizing the n^+ photodiode diffusion as cross-under as shown in Fig. 4. The simplicity in fabricating integrated area arrays utilizing the PBT element is the second major advantage compared to phototransistor arrays.

There are a number of other advantages when the PBT is used instead of the phototransistor. First, to optimize the spectral response, only the depth of the base-collector junction can be adjusted in the phototransistor. The emitter area is usually made very small in order to minimize the base-emitter junction capacitance and prevent reverse-transistor action¹, thus its contribution in collecting photo-generated carriers is negligible. In the PBT all the photo-generated carriers collected by the photodiode and collector junctions contribute to the signal. Thus, while the photodiode can be optimized for maximum response in the blue-green region of the spectrum⁸ the collector junction depth can be chosen for maximum response in the red region of the visible spectrum, giving a much broader and more efficient total spectral response. Second, in the phototransistor it is desirable to have high β values to separate the spike noise from the signal^{1,5,11,12}, particularly at low signal levels. The PBT does not have such a requirement because of

the spike noise cancellation scheme available." The value of β can therefore be designed independently of the spectral response, for example to achieve more uniformity in β over the whole array. Third, in phototransistor arrays it is required that $C_{be} \ll C_{bc}$ to eliminate crosstalk during addressing as mentioned above. No such requirement exists for PBT arrays. And fourth, the storage capacitance per unit area in the PBT is normally much higher than that of the phototransistor because of the higher impurity concentration in the photodiode-base junction as compared to the base-collector junction of the phototransistor. As a result, the PBT has a larger dynamic range.

4. Circuit Analysis

A. Modeling of the Array

The equivalent circuit of a PBT array at the moment element "B" is addressed, is shown in Fig. 5. Transients arising from the switching of the emitter rows need not be considered, since, in typical imaging applications, they occur during blanking. R_{sw} denotes the switch and line resistances. The equivalent capacitances of the switches can be included, but are neglected here. The ac equivalent circuit for an $n \times n$ element array is shown in Fig. 6. Addition of the parallel and series combinations of capacitances and resistances reduces the circuit further to that of Fig. 7. Also, in the figure, the transistor has been replaced by its η -equivalent. Since the transistor operates between cut-off and the active region, it is necessary

Fig. 5. Equivalent circuit of the PBT area array at the moment element "B" is addressed.

Fig. 6. AC equivalent circuit for an $n \times n$ element PBT array during the time one element of the array is being addressed.

Fig. 7. Addition of the parallel and series combinations of resistors and capacitors reduces the circuit of Fig. 6 to that shown here. In addition the transistor has been replaced by its π -equivalent circuit and the photodiode is represented by a voltage dependent capacitor. The current sources in parallel with C_d , C_μ and C_n account for the light and thermal generated currents of the corresponding junctions.

to model r_n by a voltage dependent resistor. V_{BE} , which corresponds to the voltage drop of the emitter diode biased in the forward direction, is also voltage dependent. The transistor current gain is strongly dependent on the base current and its variation will be included in the analysis. The current sources in parallel with C_μ and C_n account for the light generated and thermal current of the base-collector and base-emitter junctions during the integration period. The photodiode is, likewise, modeled by a voltage dependent capacitor in parallel with a current source. During the sampling period these sources contribute very small currents⁶ and their effect will be neglected during that period. Except for the junction capacitance of the element being addressed, all other capacitances in the circuit are assumed constant. Only a minor error results when C_μ and C_n are assumed constant, primarily because of their small values compared to C_d .

Analytic solution for the video signal is practically impossible both because of the presence of the non-linear elements and the complexity of the network. One method of solution is by the utilization of that section of the IBM Electronic Circuit Analysis Program, ECAP-1620⁹, that performs transient analysis. The problem then is to model the various elements of the circuit in Fig. 7 to be compatible with ECAP.

ECAP solves for the current through each branch and the voltage at each node of the network as function of time. Each branch may contain a switch, a fixed voltage or current

source and a resistor, a capacitor or an inductor. The switch in a given branch may be on or off depending on the direction of the current in that branch. The effect of changing the state of the switch is to change the value of the element in the same branch and in other designated branches to a second predetermined value. A subsequent change in the state of the switch returns the element values it controls to their first value. A branch may contain a dependent current or voltage source, but not other elements or a switch, and the gain of the source may change between two predetermined values depending on the state of a switch at another branch. Non-linear elements may be modeled by piece-wise linear approximation of their characteristics.

A typical small signal Capacitance-Voltage (C-V) curve for the photodiodes fabricated is shown in Fig. 8. This C-V curve can be fitted almost exactly by

$$C = 40.0 [1 - (V_A / .6)]^{-1/3} \text{ pF}$$

where V_A is the applied reverse bias. In the same figure, the depletion charge is plotted as function of applied reverse bias⁶. The line segments between the letters in the Q-V curve are linear. The number near each letter indicates more accurately the value of the depletion charge at that point. The piece-wise linear approximation of the Q-V characteristics serves to model the non-linear capacitance⁹. The equivalent circuit of C_d for ECAP transient analysis is shown in Fig. 9. The notation V_i indicates the voltage in the i^{th} mode. B_i indicates the i^{th} branch and

Fig. 8. Typical small signal Capacitance-Voltage (C-V) curve for the photodiode made. This curve can be fitted with the expression shown. The calculated depletion charge versus applied voltage is also shown.

Fig. 9. Equivalent circuit for the non-linear photodiode junction capacitance for ECAP transient analysis.

S_i denotes the presence of a switch in the i^{th} branch. T_i indicates a dependent current source and in this case, with a resistor r_o in parallel, simulates the current gain of a bipolar transistor. r_o is often assumed very large and omitted from the circuit. The broken line between a dependent current source and a branch indicates that this source produces a current β times that which flows in the designated branch. E is a fixed voltage source and G is the conductance.

A piece-wise linear approximation of the I-V characteristics of a typical forward biased emitter diode, is shown in Fig. 10. The parallel combinations of conductances and fixed voltage sources that simulate the forward biased emitter diode for ECAP transient analysis is shown in the left side of Fig. 11. The dependent current sources on the right simulate the β variation of the transistor. The broken lines indicate the branch current amplified by each current source. C_n is the base-emitter junction capacitance and C_μ is the base-collector junction capacitance. r_x is the base resistance. The complete network of Fig. 11 is the model for the bipolar transistor.

B. Computer Calculations

For a 16 X 16 element array, the equivalent circuit is shown in Fig. 12. The average values assumed for the various junction capacitances are; $C_{be}=4$ pF, $C_{bc}=5$ pF, and $C_d=30$ pF. The average base resistance is assumed to be $R_b=150 \Omega$ and the switch and line resistance $R_{sw}=150 \Omega$. Calculation of the video signal begins at t_1 , see Fig. 2, that is at the end of the integration period and the onset of V_S . The

Fig. 10. Piece-wise linear approximation of the I-V characteristics of a typical forward biased emitter diode.

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Fig. 12. Equivalent circuit for a 16 x 16 element PBT array.

charge collected during the integration period brings the base potential, point B in Fig. 12, from the initial negative value to a final one $v_B(t_1)$. The change in base potential Δv_B is not linearly related to the total charge collected because; a) as discussed earlier, the Q-V characteristics of the photodiode are not linear, and b) for sufficiently high incident illumination the base voltage becomes positive at some time during the integration period thus forward biasing the photodiode and allowing some of the collected photogenerated charge to leak out. Also, the collection efficiency of each junction changes during the integration period due to the decrease in the size of its depletion region^{8,13} so that even under steady illumination conditions, the photocurrent of each junction does not remain constant over the integration period. For these reasons it is convenient and more general to present the results with respect to the base voltage at the end of the integration period rather than the charge collected over the same period or the total photocurrent during the integration period.

To demonstrate the validity of the method of analysis used, we modeled the phototransistor with a n -equivalent circuit, similar to the one shown in Fig. 11, but replaced the base-collector capacitance C_μ by a circuit similar to the one shown in Fig. 9. The phototransistor was not assumed to be in an array. In Fig. 13, the peak emitter video signal is plotted as function of the base voltage $v_B(t_1)$ and is compared to the results obtained analytically by Joy and Linvill⁶ for the same phototransistor. It is

Fig. 13. The values of the peak emitter signal, for a photo-transistor modeled similarly, were calculated using ECAP and are compared here to those of Joy and Linvill (6).

seen that very good agreement is obtained.

The calculated emitter video signal as function of time and with the base voltage $v_B(t_1)$ as a parameter, is shown in Fig. 14. The equivalent circuit of Fig. 9 was used for the photodiode and the circuit of Fig. 11 was used for the transistor in Fig. 12. The emitter load resistance $R_E = 1 \text{ k}\Omega$. The dashed line in Fig. 14 corresponds to the signal obtained when $v_B(t_1)$ is sufficiently negative so that the transistor does not turn-on at all during the sampling period. A 5-volt photodiode addressing pulse is used with a 50 ns risetime and is shown on top of Fig. 14 for reference. For base voltages less negative than ~ -3.3 volts the transistor does turn-on but very weakly. Local variations of the decay RC time constant of the spike, however, will make detection of such small signals practically impossible.

The positive going portion of the calculated collector video signal is shown in Fig. 15a and its negative going part in Fig. 15b. The collector resistance is $R_C = 46 \Omega$. The dashed line in Fig. 15a again indicates the collector signal obtained for $v_B(t_1)$ sufficiently negative such that the transistor does not turn-on.

Spike noise cancellation is achieved by subtraction of the collector video from the emitter video signal. Since the spike noise depends on the signal level, this scheme of noise reduction is very desirable. The capacitive path between the addressing pulse, V_S , and the collector load resistor enhances the applicability of this method. The essentially simultaneous emitter and collector signals and

Fig. 14. Calculated emitter video signal, v_E , with $v_B(t_1)$ as a parameter. The broken line corresponds to signal obtained when the transistor does not turn-on during the sampling period.

Fig. 15a. The positive going portion of the calculated collector video signal v_C .

Fig. 15b. The negative going portion of the calculated collector video signal v_c .

the nearly tuned emitter and collector loads facilitate very efficient spike noise cancellation. The calculated composite video signal, $v_E(t) - av_C(t)$, is shown in Fig. 16, where $a = 1.02$ is used to equate the maximum peak values of the spike noise across R_E and R_C . The dashed curve is now the noise. Its peak value has been reduced by about a factor of 25. Thus the range of signals from ~ 10 to ~ 150 mv, corresponding to very low signal levels, now become accessible as is seen in Fig. 16.

The peak value of the composite video may be taken as the signal. The peak values are plotted in Fig. 17 as function of base voltage and correspond to the curve marked S40. In practice measurement of the peak value is done by a sample and hold circuit¹⁴. As seen from Fig. 16, however, the peak value occurs at different times depending on the signal level so that some error will occur in sampling the signal. The non-linearity of the curve S40 in Fig. 17 at high signal levels is due to the rapidly increasing value of the photo-diode junction capacitance for small reverse bias conditions, while at the low signal levels the non-linearity arises because of the variation of the transistor current gain with base current. To demonstrate the validity of these statements, the peak values of the composite video signal were calculated as function of $v_B(t_1)$ assuming that the photo-diode junction capacitance is constant. The value for C_d chosen was 22.5 pF. The results are plotted in Fig. 17 as curve S30. It is seen that the curve is perfectly linear except for low signal levels. Curve S40 is more non-linear

Fig. 16. Calculated composite video signal, $v_E - av_C$, where $a=1.02$. The dashed curve is the signal when the transistor does not turn-on.

Fig. 17. Calculated peak value of the composite video signal as function of $v_B(t_1)$, curve S40. For curve S30, the photodiode has been replaced by a fixed value capacitor.

than S_{30} at small signal levels because the effective value of C_d is less than 22.5 pF in this region. This reduces the forward bias across the base-emitter junction during sampling producing smaller base currents which in turn cause the transistor to operate at lower current gains.

In actual arrays, the current gain of the transistors will vary considerably over the total area of the sensor. In Fig. 18, a typical curve is shown of the peak value of the composite video as function of the maximum current gain of each transistor. $v_B(t_1)$ was assumed to be -2.0 volts. In the region of $\beta_{\max} \approx 10$, $v_{\text{peak}} \approx \ln \beta_{\max}$. The variation of β from element to element is not therefore a serious problem. Since the transistor is connected in the emitter follower configuration, for sufficiently large β , such as $\beta_{\max} > 100$ in the circuit of Fig. 12, v_{peak} becomes practically independent of β . Excessively large β values are not desirable since, as discussed earlier, the charging time constant of the photodiode junction capacitance is approximately proportional to β . As $\beta_{\max}$ becomes less than 1, the contribution of the transistor gain to the signal diminishes and as such the dependence of the signal on β becomes negligible. It may therefore be advantageous to design PBT arrays in which β is very small. Since the lifetime in the base region must have the maximum obtainable value for efficient collection of the photogenerated minority carriers, the β of the transistors can be kept small by making the base width large.

The exponential dependence of the diode forward current

Fig. 18.

Change in the peak value of the composite video as function of transistor current gain, for $v_B(t_1) = -2$ volts.

on the bias, makes the p-n diode a poor choice for a switch¹¹. In the top curve of Fig. 19, the base voltage $v_B(t)$ is plotted as function of time. It is seen that the base voltage decays very slowly once the forward bias is reduced below ~ 0.3 volts. The bottom curve in the figure corresponds to the emitter video $v_E(t)$ obtained. The three different time constants occur because of the decrease in the transistor current gain. They are easily distinguished here, of course, because of the piece-wise linear approximation of the transistor current gain.

In area arrays containing a large number of elements, the sampling time available per element, T_s is limited. At standard TV frame rates, for example, each horizontal line is scanned in $\sim 50 \mu s$. If a line consists of 200 elements, then $T_s = 0.25 \mu s$. This time is clearly insufficient to completely charge C_d as is evident from Fig. 19. In Fig. 20 the circles show the peak value of the composite video signal obtained each time the element is addressed. The sampling pulse shape and duration is also shown in the figure. For these calculations, it was assumed that the base potential was at -3.571 volts, its maximum possible value, at the time the light was first turned on and that during the first and each successive integration period the collected photogenerated charge produces an 0.5 volts change in the base voltage. C_d was again assumed constant and equal to 22.5 pF. The peak value of the signal obtained after the light is turned off is also shown. The crosses in the figure indicate the base potential at the conclusion of each of the sampling

Fig. 19. Decay of the base potential and the emitter video signal as function of time during the sampling period, for $v_B(t_1) = -2.0$ volts.

Fig. 20. Demonstration of "lag" of PBT elements.

periods. Equilibrium is reached when the charge removed from the base during the sampling period T_s is equal to the charge collected during the preceding integration period, T_i . The mechanism giving rise to this "lag" is well understood¹¹. Briefly, the charging time constant for C_d is at first dominated by the high resistance of the base-emitter diode, when $v_B(t_1)$ is sufficiently negative and then decreases as $v_B(t_1)$ becomes less negative until the transistor current gain and the emitter load impedance dominate the time constant.

It is evident from the above discussion that if the sampling period is reduced further, the base potential at the end of each sampling period at equilibrium would be less negative than shown in Fig. 20, for the same light level. From Figs. 16 and 17, it is seen that the peak value of the composite video signal at equilibrium would then be higher. This suggests a mode of operation for PBT element arrays that is particularly applicable at low light levels. The lag will however increase both when the light is turned-on and when it is turned-off, as a consequence of decreasing T_s further. The dynamic range will be reduced as well.

5. Experimental Details

Several 16 x 16 element area arrays were fabricated and tested. Figure 21a shows a closeup of the photosensitive area of the array and Fig. 21b shows the array mounted on a 2" by 1" metallized ceramic substrate. Since the scanning circuits are not on-chip, outside contacts are provided for each of the vertical and horizontal lines and

(a)

(b)

Fig.21. a) Close-up of a fabricated PBT array. The column lines make contact to the photodiodes and the row lines make contact to the emitters. This integrated structure is slightly different (15) from that shown in Fig.4. b) The array mounted in the metallized substrate.

for the common collector.

A. Fabrication Process

Standard masks obtained from Qualitron consisting of simple squares and rectangles placed on 20 mils centers, were used for the fabrication of the array. The element size was, as a result, $\sim 18 \times 18 \text{ mils}^2$ and the total area occupied by the 256 elements was $\sim 320 \times 320 \text{ mils}^2$. Because of the large size of the array and the small number of samples fabricated, the arrays made contained a number of defects.

Primarily Shipley AZ-1350J positive photoresist was used. Kodak KTR negative photoresist was also used for the following purpose. At some time during the processing of the wafer the available masks were inadequate to form the desired pattern. It was found that KTR could be applied on the wafer and a pattern developed on it. A layer of AZ-1350J photoresist could then be applied and a new pattern developed on this photoresist, without affecting the underlying KTR layer. In areas where either the KTR or the AZ photoresist were removed the oxide could be etched but not from where either of the photoresists or both were present. The capability for the KTR photoresist to act as a mask for the AZ photoresist has not been previously reported. Standard processing was followed in working with these photoresists¹⁵.

N-type, phosphorous doped, $1 \Omega\text{-cm}$, $\langle 111 \rangle$ orientation silicon wafers were used as the starting material for the fabrication of the PBT arrays. The wafers were cleaned¹⁵ and an $1.1 \mu\text{m}$ oxide was grown in 33% O_2 to 66% N_2 atmosphere at 1200°C . Base windows were opened on the oxide and

Boron was deposited at 875°C for 10 min. Boron Nitride¹⁶ type M wafers obtained from the Carborundum Corporation were used as the dopant source and were activated at 950°C for one hour in pure O_2 atmosphere just prior to the deposition step. The resulting surface resistance, following the deposition step, was $\sim 180 \Omega/\square$. The wafers were transferred directly from the B-deposition furnace to the Drive-in furnace and driven-in for 2.5 hours at 1200°C in a 50% O_2 to 50% N_2 atmosphere. This was followed immediately with an 0.5 hour annealing step at 1200°C in a N_2 atmosphere.

The emitter and photodiode windows were opened next in the oxide and phosphorous was deposited for 30 min at 1000°C . The POCl_3 liquid system¹⁷ was used. The surface resistance was $\sim 13 \Omega/\square$. The wafers were given a 10 sec, 5:1 buffered HF etch and were washed in hot distilled water. The wafers were then inserted in the oxidizing furnace, and oxidized for 0.5 hours at 1100°C . The oxidizing atmosphere consisted of water vapor obtained by bubbling N_2 gas through 80°C water. The wafers were kept an additional 1.5 hours in the same furnace but with only N_2 flowing. The resulting emitter and photodiode surface resistance were $10 \Omega/\square$ and the base surface resistance was $150 \Omega/\square$. The photodiode base junction depth was $\sim 3.4 \mu\text{m}$ and the base-collector junction was located at $\sim 6.15 \mu\text{m}$. The impurity concentrations at the surface, using Irving's curves¹⁸ are $\sim 2 \times 10^{18} \text{ cm}^{-3}$ for the base diffusion and $\sim 1.75 \times 10^{20} \text{ cm}^{-3}$ for the photodiode and emitter diffusions. A typical curve for transistor current gain versus base current is shown in Fig. 22.

Fig. 22. Typical curve of transistor current gain versus base current for the transistors in each PBT element.

A detailed manual has been written¹⁵ describing the processing steps for the fabrication of the 16 x 16 element PBT area array, including the metallization, bonding and mounting techniques.

B. Testing of the Array

A complete closed circuit TV system was built to test the array as an image sensor. The system consists of a Tektronix 604 CRT display monitor, in which the electron beam position and beam intensity are externally controlled, the circuits needed to produce the signals which are fed into the X and Y amplifiers of the 604 monitor, the scanning circuits for the 16 x 16 array, the differential video amplifier and the start and synchronizing circuits. A detailed description of these circuits and the timing diagrams are presented in Appendix I-A.

To be able to address only one element of the array at any frequency desired and to allow greater flexibility in the addressing pulse shape and amplitude, a second system was setup. In this arrangement, only one photodiode line was addressed and only one emitter line was continuously grounded through a load resistor R_E . Thus the same element was addressed always. Physically moving the array into another position in the special DIP-type socket, allowed another element to be selected. All the other emitter lines were connected together and held at an adjustable potential $+V_E$ and the remaining photodiode lines were also connected together and kept at 0 volts. Periodically all the photodiode lines were put at $+V_D$ at the same time that all emitter

lines were grounded. This was done to reset all the other elements of the array and to prevent crosstalk.

6. Experimental Results

The 16 x 16 element PBT arrays have been scanned in a television raster with the circuits presented in Appendix I-A. Figure 23a shows the picture in the face of the monitor produced from a uniformly illuminated array and Fig. 23b shows the picture produced with the array in the dark. Figure 23c shows the transmitted picture of number 2-shaped light images. In all cases the array is scanned at the rate of 200 frames per second. The brightness of each dot in the pictures is related to the amplitude of the video signal produced by the corresponding PBT element in the array. Several defects are present, such as bright spots, which correspond to leaky or shorted diodes in the array, missing elements and non-uniformity in brightness due to variation in the resistance of the lines or the commutator switches, variation in the transistor gain and other reasons.

Figure 24 is a series of oscilloscope display photographs of the emitter video signal. The top trace in Fig. 24a shows the complete raster, that is the video output from all 256 elements of the array, under uniform illumination conditions. The light level is sufficiently high in this case so that most elements saturate within the integration period. The bottom trace shows the complete raster with no light incident on the array. The transients arising from the switching of the emitter lines are present in both traces. These

(b)

()

(a)

(~)

(c)

Fig. 23.

- Photographs of the face of the monitor.
- a) Picture produced when the PBT array is illuminated uniformly.
 - b) Picture produced when no light is incident on the array.
 - c) Transmitted picture of number 2-shaped light images.

Fig. 24.

Oscilloscope face Photographs of the emitter video signal.

- a) Complete raster at saturation light level and at dark.
- b) Video output from 12 of the 16 elements of one line of the array under high and uniform illumination and at dark.
- c) Complete raster at two light levels below saturation.

spikes occur during the blanking period and do not produce an image on the monitor. The emitter video signal from 12 of the 16 elements of one line of the array is shown in the top trace of Fig. 24b, also for uniform illumination conditions and saturation light level. The bottom trace is the video output with no light incident on the line. The first large spike in this trace corresponds to the switching of the emitter line. It is clear from Figs 24a and 24b that the output is fairly uniform, for each line, at sufficiently high levels of illumination, but that a large number of defects are present. The non-uniformity in peak value of the video signal becomes worse at lower light levels¹². This is demonstrated in Fig. 24c where the complete raster is presented for uniform illumination conditions but for two light levels below saturation.

A more accurate and more detailed view of the emitter and collector video signals is obtained if the array is tested in the special test-jig, discussed in section 4B. There are no shift registers or commutator switches in this case and the various voltages are supplied externally from pulse generators and power supplies. With the array in place, however, only one element can be interrogated. The data to be presented in Figs. 25, 26 and 27 were obtained from this test arrangement.

A drawing from an oscilloscope photograph of v_E , v_C and $v_E - av_C$ at high light levels is shown in Fig. 25a. The top trace is the emitter video and the middle trace is the collector video signal inverted. The composite video signal

Fig. 25

View of the emitter and collector video signals. Demonstration of the spike noise cancellation capability of the PBT arrays and the resulting improvement in the signal to noise ratio.

is shown in the bottom trace. These signals bear close resemblance to the calculated ones in section 3. Experimentally it is not possible, however, to know the base potential at the end of the integration period since there is no contact to the base. The addressing pulse height is 2.1 volts and the pulse risetime and falltime are 50 ns. T_s is $\sim 1.6 \mu s$ and T_i is $\sim 50 \mu s$. The spikes produced at the outputs when the pulse turns off are also shown in the figure. The improvement in the signal to noise ratio in the composite video, because of the cancellation of the spike noise and the addition of the signal, is quite evident.

It was mentioned in Section 3 that, in arrays containing a large number of elements, the sampling time per element could be of the order of few tenths of a μs . The spike noise cancellation that is possible when short duration addressing pulses are used is demonstrated in Fig. 25b. There is no light incident on the element in this case. The pulse width is ~ 200 ns and its amplitude is 5 volts. The top double trace in Fig. 25b is the emitter and collector signals superimposed. The vertical scales are 0.05 V/cm for v_E and ~ 0.1 V/cm for v_C . Inverting v_C and adding it to v_E results in the bottom trace of Fig. 25b, where the vertical scale for this trace is 0.05 V/cm. $v_E - av_C$ is shown in an expanded scale in Fig. 25c. It is seen that the spike noise has been reduced to ~ 10 mV in the composite video as compared to 100 mV in v_E and ~ 200 mV in v_C .

The experimentally obtained peak values of the composite video signal as function of integration time and constant

light level are shown by the crosses in Fig. 26. The computer calculated values, as function of charge collected during the integration period, are shown by the dots in the same figure. The calculated values are the same with those of curve S40 in Fig. 17, plotted here against collected charge rather than $v_B(t_1)$. The two curves in Fig. 26 are very similar except at the very high signal levels. The saturation exhibited by the experimental curve occurs because of the loss of collected charge once v_B becomes positive during the integration period which forward biases the diodes. This charge loss mechanism is not taken into account in the model and it is the reason for the discrepancy. The shift between the two curves occurs because the base potential is not zero at the end of the sampling period⁶, as it was assumed for the calculations, but it remains few tenths of a volt positive, as demonstrated in Fig. 19. The shift, therefore, represents the amount of charge that is not removed from the base. The addressing pulse amplitude was 5 volts and had a risetime of 50 ns. The sampling time was 10 μ s in the experiment and effectively $T_s = \infty$ in the calculations. Both the experimental and calculated curves in Fig. 26 are linear over a wide range and the experimental curve saturates quite abruptly. This behavior for the video signal is very desirable in image sensors.

It was shown theoretically in Section 3 that as the sampling period is reduced, while the integration time and light level are held constant, the peak value of the video signal increases. This is demonstrated experimentally in

Fig. 26

Comparison of the calculated S40 and experimental (S#37-II) results of the peak value of composite video signal for the fabricated PBT arrays.

Fig. 27. This figure has been drawn from photographs of the oscilloscope CRT display. The top trace in each of the three portions of Fig. 27, shows the addressing pulse. The second trace from the top in each portion shows the emitter video signal when the array is illuminated and equilibrium has been reached and the bottom traces show v_E when there is no light incident on the element and again at equilibrium. The integration time in all cases is 0.2 ms.

It is clear that when a sampling pulse 50 μ s wide is used, it is practically impossible to detect the signal while if a sampling pulse 0.2 μ s wide is used the same signal is easily detectable. To take full advantage of this effect in PBT area arrays it would be required to design the shift registers, that address the photodiode lines, in such a way as to permit adjustment of the sampling period without affecting the frame rate. Thus when viewing scenes at low light levels or where there is more interest in viewing the dark portions of a scene, T_S could be set at a low value while for brightly illuminated scenes T_S would be set at a high value. As it was previously discussed, the lag increases as T_S decreases.

7. Conclusions

16 x 16 element silicon image sensors utilizing the photodiode-bipolar transistor (PBT) structure at each site were fabricated, tested and scanned in a television raster successfully transmitting light images.

It was shown that the spike noise cancellation scheme, inherent in the PBT element, can be effectively utilized in

Fig. 27

Effect of addressing pulse width on the peak value of the emitter video signal.

area arrays to reduce the peak value of the spike noise by more than an order of magnitude while at the same time increasing the peak value of the signal.

The simplicity in fabricating PBT area arrays was demonstrated by making several working sensors using standard masks consisting of various size rectangles and squares on 20 mil centers and requiring only two diffusions and one metallization.

It was found that the Kodak thin film resist (KTFR) could be used as a mask for the AZ-1350J photoresist.

The PBT element and the complete array were modeled so that the video signal calculations could be performed using the computer program ECAP-1620. The following effects were taken into account in modeling the PBT element; a) the non-linearity of the photodiode junction capacitance with reverse bias, b) the variation of the base-emitter junction resistance and p-n junction voltage drop with forward bias and c) the variation of the current gain of the transistor with base current. Very good agreement was obtained between the calculated and experimental results.

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APPENDIX I - A

Description of the Closed Circuit TV System

A block diagram of the system is shown in Fig. 1 and a diagram of the start and synchronizing circuit is shown in Fig. 2. The timing diagram for the system is shown in Figs. 3 and 4. In table I are listed the various integrated circuits used.

Each of the 16 outputs of the column shift register (fast SR), see Fig. 1, is connected to one photodiode line of the PBT array. The shift register is connected in a ring counter configuration and it is arranged that only one of the outputs is up, at a time. Each of the outputs of the row shift register (slow SR) is connected to the gate of a COS/MOS switch. The 16 COS/MOS switches are the commutator. One of the two source outputs from each COS/MOS switch is connected to R_E and the other to $+V_E$. Each emitter line is connected to the common drain of the corresponding COS/MOS switch. When the gate voltage is positive, the COS/MOS switch connects the emitter line to R_E . When the gate voltage is zero, the emitter line is connected to $+V_E$. The slow SR is connected also in a ring counter configuration and arranged so that only one stage is up at each time.

Fig. 1. Block diagram of part of the closed circuit TV system.

Fig. 2. Block diagram of the scan and synchronizing circuits.

NOTE:

In (A) beam in monitor is positioned during last half period each of the fast SR outputs is on.
In (B) beam is positioned at the same time as each Q output of fast SR goes up.

Fig. 3. Timing diagram for the closed circuit TV system.

Fig. 4. Timing diagram for the closed circuit TV system.

TABLE I

THE INTEGRATED CIRCUITS USED IN
THE CLOSED CIRCUIT TV SYSTEM

NUMBER	DESCRIPTION	MANUFACTURER
DM8570	8-Stages, Serial-in Parallel-out Shift Register	National Semi- conductor Corp.
SN7476	Dual J-k Master-Slave Flip-Flops	Texas Instru- ments (TI)
SN7400	Quad NAND Gates	(TI)
SN7402	Quad NOR Gates	(TI)
SN7404	Hex INVERTERS	(TI)
SN72733	Differential Video Amplifier	(TI)
CD4007A	COS/MOS Dual Complementary Pair plus Inverter. (Connected in the triple Inverter configuration).	(RCA)
DAC-19BBI	8-Bit, Digital to Analog Converter	Datel Systems Inc.

Fig. 5. Synchronous horizontal counter and digital to analog converter. The logic in the bottom is for synchronization. The logic on top and extra flip-flop allows for 17 pulse cycle while the D/A converter produces only 16 voltage steps per cycle.

II

COLLECTION EFFICIENCY AND TRANSFER CHARACTERISTICS OF CID IMAGE SENSORS

1. Abstract

The collection process of the photogenerated charge in CID image sensors is examined and a method is presented by which the collection efficiency can be calculated. Numerical techniques are utilized in the analysis. The method is applied to a specific epitaxial type structure. The contribution of the depleted and undepleted regions to the total collection efficiency of the sensor is calculated.

The relation between output signal charge and incident light intensity is also investigated. It is shown that the collection efficiency decreases with increasing signal charge and this results in non-linear transfer characteristics at high signal levels.

2. Introduction

The Charge Injection Device (CID) is presently used in fabricating solid state area type imaging arrays. Experimental data have been presented^{1,2}, describing the performance of the sensors. In the CID a substantial portion of the surface and volume, in which the visible radiation

is absorbed, is undepleted. The efficiency with which the minority carriers created in this region contribute to the signal has not been discussed. Also, the non-linearity between incident intensity and output signal charge has not been adequately explained. It is the purpose of this paper to examine these two points.

Numerical techniques are utilized in the analysis so that the complicated boundaries of the CID structure can be taken into account, and simplifying assumptions be avoided as much as possible. The calculations are restricted to a two-dimensional structure but the method can be extended to the three-dimensional case. The results, however, provide adequate understanding of the performance of actual devices.

3. Description of the CID Image Sensor

A top view of an epitaxial type CID image sensor is shown in the upper portion of Fig. 1 and the cross section along the dotted line is shown in the lower part. The CID image sensor consists of a two-dimensional array of MOS capacitors. Each capacitor consists of two gates, one connected to an x-row buss and the other to a z-column buss. A depletion potential is continuously applied to both gates. Minority carriers created by the incident radiation or by thermal generation are collected and stored as surface inversion charge in the Si-SiO₂ interface under the gates. Signal readout is achieved by setting both gates to zero volts thus injecting the interface charge into the n-layer. The resulting majority current entering the n-layer to

Fig. 1 Upper portion of figure shows top view of a CID image sensor. Bottom part is a cross-sectional view along the broken line.

neutralize these minority carriers is integrated over the addressing period. This integrated charge constitutes the signal. The use of two gates permits x-z addressing of the array. When a line is set to zero volts, the charge under each gate connected to that line is transferred to its neighboring gate. Net charge is released to the substrate only from that element for which both gates are set to zero volts. The p-diffusion shown in Fig. 1 is used to couple the charge between the two gates, thus relaxing the requirement for close proximity of the two gates.

Originally the sensors were made on bulk wafers but recent ones are built on epitaxial wafers because they offer several advantages. The reverse biased epitaxial junction underlying the whole array; (a) collects efficiently the injected minority carriers thus increasing the speed of operation while maintaining the dynamic range, (b) prevents capture of the injected charge by the neighboring potential wells, thus minimizing cross-talk and image smearing, (c) limits "blooming" arising from radial spreading of the charge from an intensely illuminated point, by limiting the distance carriers can travel³ and (d) improves the MTF of the sensor since carriers created deep in the silicon, which would otherwise give rise to image spreading, are not collected. The main disadvantage is that the collection efficiency of the epitaxial sensors is a factor of two smaller than the bulk ones or worse². It is of interest to note that for present types of CID sensors this decrease cannot be compensated by thinning the wafer and having the light incident

from the back. The reason is that high injection efficiency requires a high surface recombination velocity at the back surface (provided now by the reverse biased epitaxial junction) while high collection efficiency requires low surface recombination velocity at the same surface.

For the 100 x 100 element array described by Hooker⁴ the dimensions of each element were $\sim 64 \times 84$ microns. The area occupied by each pair of the semitransparent polysilicon gates was several times smaller than the total element area, primarily to minimize the dark current.

4. Collection Efficiency

The light image is usually incident on the CID image sensor from the electrode or front side. From the regions covered only with SiO_2 , (see Fig. 1) very little of the incident visible radiation is reflected if the SiO_2 film thickness is properly chosen. The light incident on the metallic interconnecting strips, shown by the cross-hatched regions in Fig. 1, is almost completely reflected. The strong interference effects produced by the polysilicon films deposited on top of SiO_2 to form the gates lead to substantial wavelength dependent reflection losses⁵ of the light incident on the gates. In addition, the polysilicon films although normally very thin, absorb a considerable amount of the incident radiation, particularly at short wavelengths where the absorption coefficient of silicon is very large. The photon flux ϕ_A that reaches the silicon surface at every point, therefore, depends both on the

wavelength of the incident photons and the position on the surface.

The generation rate of electron-hole pairs inside the semiconductor is given by

$$g_{\lambda}(x,y,z) = \alpha_{\lambda} \varphi_{\lambda}(x,z) \exp(-\alpha_{\lambda}y) \quad (1)$$

where λ is the wavelength, y is the distance inside the semiconductor from the top surface, α_{λ} is the absorption coefficient in cm^{-1} , g_{λ} is measured in $\text{pairs}/\text{cm}^3\text{sec}$ and φ_{λ} is the photon flux, in $\text{photons}/\text{cm}^2\text{sec}$, that enter the semiconductor at the surface $y = 0$.

Referring to Fig. 1, it is seen that electron-hole pairs are created by the light in four distinct regions, namely, the substrate, the depletion region of the epitaxial junction, the depletion region beneath the gates, and the undepleted portion of the epitaxial layer. All the holes created inside the depletion region (well) of the gates will be attracted to the surface and add to the surface inversion charge. Parenthetically we note here that of the electrons created some flow to the gate to compensate for the excess holes at the Si-SiO_2 surface while the remaining neutralize some of the ionized donors at the edge of the depletion region to adjust for the change of the surface potential. The exact relationship between charge out and interface inversion charge will be discussed in Section 5. None of the holes created in the substrate contribute to the surface inversion charge because they cannot cross the epitaxial junction. Holes created inside the depletion region of the

epitaxial junction cannot contribute to the surface charge because the built-in field of the junction sweeps them towards the substrate rather than toward the epitaxial layer. Finally of the holes created in the undepleted region of the epitaxial layer some diffuse toward the well where they are captured and add to the surface charge. The others are lost either (a) by diffusing to the edge of the depletion region of the epitaxial junction and being swept across to the substrate, (b) by diffusing to the surface and recombining there at a rate determined by the surface recombination velocity or (c) by recombining in the epitaxial layer because of their finite lifetime. As is evident from Fig. 1, most of the surface and volume of the epitaxial layer is undepleted. The spectral response or collection efficiency of the sensor will therefore depend strongly on how many of the holes created in this region are collected by the well as function of wavelength, epitaxial layer thickness, lifetime and surface recombination velocity. In steady-state uniform illumination conditions, the concentration of holes $p_{\lambda}(x,y,z)$ everywhere inside the undepleted region of the epitaxial layer is found by solving the diffusion equation

$$D \nabla^2 p - \frac{p}{\tau} = -g_{\lambda}(x,y,z) \quad (2)$$

where p is in cm^{-3} , D is the hole diffusion constant in cm^2/sec and τ is the hole lifetime in seconds. Once $p_{\lambda}(x,y,z)$ is known, the current density at any point can be calculated using

$$\vec{J}_p = -D \vec{\nabla} p \quad (3)$$

and the number of holes per second flowing into the well, or lost to surface recombination or flowing across the epitaxial junction, can be found by integrating J_p over the appropriate surfaces.

Efficient, straightforward numerical methods using the finite difference approximation and employing the computer^{6,7} are presently available to solve the diffusion equation (2). The irregular boundaries which constitute the main difficulty in obtaining an analytic solution are easily taken into account. Analytic expressions for the boundaries are not needed. In the CID image sensor, for example, the boundary of the depletion region due to the potential on the gates can be determined numerically by solving Poisson's equation^{8,9} and then entered into the program solving the diffusion equation. Also, the computer programs can be written in a general fashion so that design parameters can be varied until optimum values are determined.

The cross section of a specific CID structure for which we will calculate the collection efficiency is shown in Fig. 2. The numbers in the figure represent dimensions in microns. To save computer time and reduce program complexity we will assume that the gates of each element are almost as long as the element. If, in addition, we assume uniform illumination, then to a good approximation p_λ can be taken to be constant along the z-direction. Equation (2) then becomes two-dimensional and takes the form,

$$D\left(\frac{\partial^2 p}{\partial x^2} + \frac{\partial^2 p}{\partial y^2}\right) - \frac{p}{\tau} = -g_\lambda(x,y) \quad (2a)$$

Fig. 2. The cross-section of the CID image sensor used for the calculations. The numbers indicate dimensions in microns.

The boundary conditions are (see Fig. 2):

- (i) $p = 0$ at the depletion region boundaries of both the well and the epitaxial junction,
- (ii) $D(\partial p / \partial y) = sp$ at the undepleted portion of the surface, where s is the surface recombination velocity,
- (iii) $(\partial p / \partial x) = 0$ at $x = 0$ and $x = 32$ microns because of the symmetry of the structure and the assumed uniform illumination condition.

In Fig. 2 we have neglected the p -diffusion between each pair of gates and instead have assumed that the gates are sufficiently close together. Also, the depletion region at the edge of the gate was assumed to form a quarter circle with radius equal to the well depth measured directly beneath the gate. The last two assumptions produce a small error but obviate the need of solving the two-dimensional Poisson's equation to find the potential inside the semiconductor. The well depth y_d directly beneath the gate is calculated then from the one-dimensional solution of Poisson's equation, and is given by $y_d = (2\epsilon_s |V_S| / qN_D)^{1/2}$, where ϵ_s is the dielectric constant of the semiconductor, q is the electronic charge, N_D is the donor impurity concentration and V_S is the surface potential. The relation between y_d and the surface inversion charge Q_{inv} is

$$y_d = (-b + \sqrt{b^2 - 4ac}) / 2a \quad (4)$$

where $b = qN_D$, $a = (\epsilon_{ox} q N_D / 2\epsilon_s x_o)$, $c = -(\epsilon_{ox} |V_G| / x_o) + Q_{inv}$, ϵ_{ox} is the oxide dielectric constant, x_o is the oxide thickness, and V_G is the gate voltage measured from flat

band conditions. V_G and V_S are negative quantities.

Finally, we will neglect any reflection or absorption losses due to the SiO_2 or the polysilicon gates, since neither the thickness of the films nor a reflection spectrum have been published. Such a spectrum can, however, be included in our present computer programs without any difficulty.

Equation (2a) was solved using the "successive over-relaxation" or "extrapolated Liebman" point iterative method^{6,7} and a square mesh of 0.5μ was chosen. Numerical integration and differentiation¹⁰ techniques were also applied to equation (3) to evaluate the current flowing from the undepleted region into the well, across the epitaxial junction into the substrate, and into the surface. The accuracy of the calculations was checked by comparing the total calculated current to the total number of holes created per second in the undepleted region, the latter being determined accurately by integrating equation (1). In all cases the two values agreed to within 5%. Their difference was less than 1% for zero surface recombination velocity and epitaxial layer thickness less than 20 microns.

In the calculation the values for the diffusion constant and lifetime of the minority carriers have been taken as $D = 10 \text{ cm}^2/\text{sec}$ and $\tau = 10 \mu\text{sec}$ which are reasonable values for the devices. The values of the absorption coefficient α_λ have been taken from the data of Dash and Newman¹¹. Finally, the following quantities are defined to aid in the presentation of the results:

H_{dep} = holes/sec created in the depleted region.

H_{und} = holes/sec created in the undepleted region.

H_{sub} = holes/sec created in the substrate.

$H_{tot} = H_{dep} + H_{und} + H_{sub}$ = total number of holes generated per second.

I_{wel} = holes/sec flowing into the well from outside it.

In Fig. 3 we plot the collection efficiency of the undepleted region, $\epsilon = I_{wel}/(H_{und} + H_{sub})$, as function of the epitaxial layer thickness for different wavelengths. The light absorbed by the well H_{dep} has been omitted from the denominator of the above expression. Clearly, the presence of the epitaxial junction substantially lowers the collection efficiency of the undepleted region both for the short and long wavelength radiation. It is of interest to point out that because $\tau = 10 \mu\text{sec}$ the loss of holes to recombination in the bulk of the epitaxial layer is negligible even for the 40 micron thick layer.

The set of curves of Fig. 4 depict the collection efficiency of the undepleted region as function of wavelength with the surface recombination velocity s as parameter. The collection efficiency is defined here as $\epsilon = I_{wel}/H_{tot}$ and it now includes H_{dep} in the denominator. These curves apply for the structure of Fig. 2 where the epitaxial layer thickness is equal to 10 microns. To put these curves into perspective one must use the curves of Fig. 5 where the amount of light absorbed inside the well and that absorbed in the undepleted region is plotted as a function of wavelength. For example, at $\lambda \sim 0.62 \mu$, $\sim 50\%$ of the incident radiation is absorbed by the well and $\sim 50\%$ by the undepleted

Fig. 3. The portion of the holes created in the undepleted region and substrate and captured by the potential well as function of epitaxial layer thickness with the wavelength as parameter.

Fig. 4. Collection efficiency $\Theta = I_{\text{wel}}/H_{\text{tot}}$ of the undepleted region as function of wavelength for various surface recombination velocities. Curve (a) is the collection efficiency $\rho = H_{\text{dep}}/H_{\text{tot}}$ of the well.

Fig. 5. The portion of light absorbed in the depleted and undepleted regions of the structure in Fig. 2 as function of wavelength.

region. From Fig. 4, $\Theta \sim 0.2$, for $s = 0$, which indicates that only $\sim 40\%$ of the light absorbed in the undepleted region contributes to signal. Two points of interest in Fig. 4 are: (a) as expected, high surface recombination velocities are more detrimental to short wavelength radiation and (b) that Θ changes very little for s between 0 and 10^3 cm/sec which is interesting since values of $s \sim 10^3$ cm/sec or less are presently obtainable¹² in Si-SiO₂ interfaces.

The collection efficiency of the well, $\rho = H_{\text{dep}}/H_{\text{tot}}$ is, of course, equal to the absorption efficiency. ρ is plotted as function of wavelength in Fig. 4, curve (a). Since most of the depleted region is covered by the polysilicon gates which produce large interference effects and also absorb some portion of the light at the short wavelengths, the curve presented here can only be thought of as an ideal efficiency. The total collection efficiency of the structure would then be $\eta = \rho + \Theta$, and for $s = 0$ this is plotted as the top curve of Fig. 6. The set of points in this figure correspond to the data of Michon and Burke². The smooth line between the points was drawn for comparison with the top curve. The greatest discrepancy between these two curves occurs in the short wavelength region where the losses due to the polysilicon gates and surface recombination are expected to be important.

The analysis presented so far applied to small signal levels. At higher signals the size of the depletion region decreases because of the increasing surface inversion

Fig. 6. Top curve represents the ideal maximum collection efficiency, $\eta = \theta + \rho$, that can be expected from the structure of Fig. 2. The smooth line through the experimental points is drawn for comparison with the top curve.

charge, and this causes a reduction in the collection efficiency. These effects are treated in the following section.

5. Transfer Characteristics

The charge released to the epitaxial layer at the moment of addressing one element is equal to $Q_{inv} + qN_D y_d$, where Q_{inv} is the surface inversion charge and is equal to the photon and thermally generated holes collected by the well during the integration period. The second term represents the depletion region charge at the moment of addressing. y_d is the well depth. Thus $(Q_{inv} + qN_D y_d)/q$ represents the number of majority carriers that must flow into the epitaxial layer. We may assume that by the end of the addressing period the holes comprising Q_{inv} have either recombined or have been collected by the epitaxial junction. Thus when V_G goes up at the end of the addressing pulse, the depletion region acquires its maximum width y_{dmax} (zero inversion charge)¹³. The majority charge that must leave the epitaxial layer is $qN_D y_{dmax}$. The net signal charge Q_{sig} is then

$$Q_{sig} = Q_{inv} + qN_D y_d - qN_D y_{dmax} \quad (5)$$

Implicit in the above discussion is the assumption that the charge that enters the epitaxial layer at the onset of addressing, to neutralize the charge of the depletion regions of the gates connected to the same line as the addressed gate, is equal to the charge leaving at the end of the addressing period, when the depletion regions are established again. This assumption is justified if the

addressing period is short relative to the integration period.

Equation (5) indicates that a non-linear relationship between the charge out to the photo-generated charge will occur if the difference between the last two terms of the equation are significant relative to Q_{inv} . In Fig. 7, we plot the signal versus the inversion charge using equation (5). The parameters used are: gate voltage $V_G = -12$ volts, oxide thickness $x_o = 0.1 \mu$, and impurity concentration $N_D = 5 \times 10^{14} \text{ cm}^{-3}$, which are typical values. In the same graph the well depth is plotted as a function of Q_{inv} using equation (4). Figure 7 clearly shows that if the collection efficiency η_λ remains constant as the well depth decreases, that is, Q_{inv} remains proportional to H_{tot} , then the transfer characteristic is linear and saturates abruptly. A small non-linearity would appear at high signal levels either for $N_D > 1 \times 10^{15} \text{ cm}^{-3}$ if $x_o = 0.1 \mu$ or for $x_o > 0.15 \mu$ if $N_D = 5 \times 10^{14} \text{ cm}^{-3}$. The impurity concentration or oxide thickness required to produce an appreciable non-linearity are therefore significantly above typical values.

A pronounced non-linear behavior between H_{tot} and Q_{sig} occurs because of the reduction in the collection efficiency η_λ with increased signal charge. The reduction in η_λ is a direct consequence of the shrinking of the well as Q_{inv} increases. This is demonstrated in Fig. 8 where η_λ is plotted as function of y_d for three different wavelengths. The instantaneous increase in Q_{inv} is given by

$$dQ_{inv} = \eta_\lambda(y_d) \phi_\lambda q \, dt \quad (6)$$

Fig. 7. Transfer characteristic of the CID assuming constant μ . The depletion region depth is also plotted as function of Q_{inv} .

Fig. 8. Variation of η with well depth for three wavelengths.

where the functional dependence of n_λ to y_d is obtained by curve-fitting of the curves in Fig. 8. Using equation (4), n_λ can be expressed in terms of Q_{inv} instead of y_d . The total charge, Q_{invf} , collected over an integration period T_i can then be found from the following equation

$$\int_0^{Q_{invf}} \frac{dQ_{inv}}{n_\lambda(Q_{inv})} = \int_0^{T_i} q \varphi_\lambda dt \quad (7)$$

The transfer characteristic calculated using both equations (7) and (5) is shown as curve (a) in Fig. 9. The calculation was made for $\lambda = 0.67 \mu$ and $s = 0$ cm/sec. The relationship between n_λ and y_d was taken to be linear, which is a good approximation as is evident from Fig. 8. The significant feature of this curve is, of course, that it shows non-linearity at higher signal levels. The non-linearity becomes worse and the dynamic range is reduced if the injection efficiency is less than unity, i.e., some of Q_{invf} is collected again in the well at the beginning of the next integration period. This is included in the analysis by substituting Q_{invo} instead of zero as the lower limit of the integral on the left hand side of equation (7). The calculated curve when $Q_{invo} = Q_{invf}/2$ is shown as (b) in Fig. 9. The points in the figure are the experimental results of Michon and Burke². The radiant power scale corresponds to their data. We should point out, however, that the reported injection efficiency of the epitaxial structure is close to 100%.

Fig. 9. Transfer characteristics for η varying as shown in Fig. 8 for $\lambda = 0.67 \mu$. Curve (a) assumes 100% injection efficiency, while curve (b) assumes 50% injection efficiency. The points correspond to experimental data of Michon and Burke (2).

6. Conclusions

We have examined the photogenerated charge collection process in the CID image sensor and have presented a method by which the collection efficiency can be calculated. The method has been applied to a particular CID structure, shown in Fig. 2, and its collection efficiency has been calculated as function of epitaxial layer thickness, wavelength and surface recombination velocity. In addition the transfer characteristics of the structure, relating the total number of carriers created in the silicon to signal charge have been studied.

It has been shown that:

- (a) the collection efficiency ϵ of the undepleted region is substantially reduced as a result of the epitaxial junction. The highest rate of increase of ϵ with epitaxial layer thickness occurs for thicknesses between 5 and 20 microns;
- (b) at most $\sim 46\%$ of the holes created in the undepleted region are collected by the well. This occurs for $s=0$ and short wavelength radiation. The efficiency decreases for longer wavelengths or higher surface recombination velocities;
- (c) one significant source of non-linearity in the transfer characteristics is the decreasing collection efficiency with signal level. The decrease in η_λ occurs because of the shrinking of the potential well with signal level and the rate of decrease depends on the wavelength of the incident radiation.

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APPENDIX II - A

Numerical Solution of the Diffusion Equation

To solve equation (1) numerically,

$$D \left(\frac{\partial^2 p}{\partial x^2} + \frac{\partial^2 p}{\partial y^2} \right) - \frac{p}{\tau} = -g_{\lambda}(y) \quad (1)$$

the undepleted portion of the epitaxial layer in Fig. 1, between the vertical broken lines marked $x = 0$ and $x = 32 \mu\text{m}$, was divided into a square mesh of $0.5 \mu\text{m}$ sides. From Taylor series expansion and neglecting higher order terms we have;

$$p(x+h, y) \approx p(x, y) + \frac{\partial p}{\partial x} h + \frac{\partial^2 p}{\partial x^2} \frac{h^2}{2}$$

$$p(x-h, y) \approx p(x, y) - \frac{\partial p}{\partial x} h + \frac{\partial^2 p}{\partial x^2} \frac{h^2}{2}$$

where h is the incremental distance, in cm, between neighboring mesh points. Adding and solving for $\frac{\partial^2 p}{\partial x^2}$ gives,

$$\frac{\partial^2 p}{\partial x^2} \approx \left[p(x+h, y) - 2p(x, y) + p(x-h, y) \right] / h^2$$

Similarly for the y -direction;

$$\frac{\partial^2 p}{\partial y^2} \approx \left[p(x, y+h) - 2p(x, y) + p(x, y-h) \right] / h^2$$

Fig. 1. The cross-section of the CID image sensor used for the calculations. The numbers indicate dimensions in microns.

Letting $p(x+h, y) = p(i+1, j)$, etc. equation (1) becomes;

$$\left(\frac{D}{h^2}\right) \left(p(i+1, j) + p(i-1, j) + p(i, j-1) + p(i, j+1) - 4p(i, j) \right) = -g(j) + p(i, j)/\tau$$

Thus,

$$p(i, j) = A \left(p(i+1, j) + p(i-1, j) + p(i, j-1) + p(i, j+1) + \left(\frac{h^2}{D}\right) g(j) \right)$$

where $A = \left(\frac{D}{h^2}\right) \left\{ 1 / \left[\left(\frac{4D}{h^2}\right) + \left(\frac{1}{\tau}\right) \right] \right\}$.

For the iteration proceeding from left to right and from top to bottom,

$$p^{(n+1)}(i, j) = A \left\{ p^{(n)}(i+1, j) + p^{(n+1)}(i-1, j) + p^{(n)}(i, j+1) + p^{(n+1)}(i, j-1) + \left(\frac{h^2}{D}\right) g(j) \right\}$$

where $p^{(n)}$ denotes the n th iteration value.

In the "extrapolated Liebmann" or "successive over-relaxation" iterative method,

$$p^{(n+1)}(i, j) = p^{(n)}(i, j) + \omega R(i, j)$$

where ω is the "relaxation factor" and $R(i, j)$ is the residual at any mesh point defined as the amount by which the finite difference equation differs from zero and is given therefore by;

$$R(i, j) = p^{(n+1)}(i, j) - p^{(n)}(i, j)$$

The final difference form for the hole concentration at any

mesh point except at the Si-SiO₂ interface and the edges of the depleted regions, is;

$$p^{(n+1)}(i,j) = \omega A \left\{ p^{(n)}(i+1,j) + p^{(n+1)}(i-1,j) + p^{(n)}(i,j+1) + p^{(n+1)}(i,j-1) + \left(\frac{h^2}{D}\right)g(j) \right\} + (1-\omega) p^{(n)}(i,j) \quad (2)$$

At the edges of the depleted regions, $p(i,j) = 0$ at all times. In the curved portion of the depletion region of the well many of the mesh points do not fall exactly on the boundary. The approximation made was to assume that the depletion region boundary passes through the closest mesh point. At the Si-SiO₂ interface, the following boundary condition must be satisfied,

$$D \frac{\partial p}{\partial y} = sp \quad (3)$$

From Taylor series expansion and neglecting high order terms;

$$p(x, y+h) \approx p(x,y) + \frac{\partial p}{\partial y} h + \frac{\partial^2 p}{\partial y^2} h^2$$

$$p(x, y-h) \approx p(x,y) - \frac{\partial p}{\partial y} h + \frac{\partial^2 p}{\partial y^2} h^2$$

subtracting and solving for $\frac{\partial p}{\partial y}$ we get;

$$\frac{\partial p}{\partial y} = [p(x, y+h) - p(x, y-h)] / 2h$$

The mesh points at the surface have index $j = 2$. Thus;

$$\left. \frac{\partial p}{\partial y} \right)_{j=2} = [p(i,3) - p(i,1)] / 2h$$

Equation (3) becomes;

$$[p(i,3) - p(i,1)] / 2h = \left(\frac{S}{D}\right) p(i,2)$$

and;

$$p(i,1) = - \left(\frac{2hs}{D}\right) p(i,2) + p(i,3) \quad (4)$$

Since at the Si-SiO₂ interface equations (1) and (3) must be solved simultaneously, we have from equation (2);

$$p^{(n+1)}(i,2) = \omega A \left\{ p^{(n)}(i+1,2) + p^{(n+1)}(i-1,2) + p^{(n)}(i,3) + p^{(n+1)}(i,1) + \left(\frac{h^2}{D}\right) g(2) \right\} + (1 - \omega) p^{(n)}(i,2)$$

Substituting for $p(i,1)$ from equation (4), gives;

$$p^{(n+1)}(i,2) = \omega A \left\{ p^{(n)}(i+1,2) + p^{(n+1)}(i-1,2) + p^{(n)}(i,3) - \left(\frac{2hs}{D}\right) p^{(n)}(i,2) + p^{(n)}(i,3) + \left(\frac{h^2}{D}\right) g(2) \right\} + (1 - \omega) p^{(n)}(i,2) \quad (5)$$

Equation (5) is then the difference form used to calculate the hole concentration at the Si-SiO₂ interface. Point (i,1) is a fictitious point and the hole concentration is never calculated there.

In the calculations, presented in manuscript II,

$\omega = 1.875$ provided the maximum rate of convergence. Between 75 to 150 iterations were needed for convergence depending on the width of the epitaxial layer and the depth of the well. The solution was assumed to have converged when

$$|p^{(n+1)}(i,j) - p^{(n)}(i,j)| \leq 10^3 \text{ holes/cm}^3.$$

APPENDIX II - B

Design for Back Illuminated CID Image Sensor Array

Present designs of CID image sensors cannot be operated in the back illumination mode. The reason is that high injection efficiency requires a high surface recombination velocity at the back surface while high collection efficiency, for a thinned sensor, requires low surface recombination velocity at the same surface.

A design which would allow back illumination operation is shown in Fig. 10. The major difference between this sensor and the already existing types is that it is made on a bulk (rather than epitaxial) wafer and that vertical p-channels are diffused between the elements. These channels are as long as the array and can be diffused at the same time as the coupling diffusion between each pair of gates.

In normal operation all the channels are reverse biased relative to the n-substrate. When an element is addressed and the surface charge released, the channel nearest to that element serves as a sink for the released charge. The signal may thus be taken from the addressed column, as in present sensors, or it may be taken from the channels.

Fig. 1. Proposed structure for a novel CID image sensor suitable for operation with back illumination.

All the channels may be connected together and biased to a low dc potential, or the channels may be separate and connected to a shift register. In the later case the reverse bias on the channel adjacent to a column is increased when the column is addressed. This makes it a more efficient sink. The other channels can be held at some low reverse bias to serve as blooming inhibitors.

The advantages of the proposed sensor are:

- (1) It can be thinned and used with light incident through the back of the wafer, thus getting better spectral response and collection efficiency.
- (2) It can be operated in the Silicon Intensifier Target (SIT) mode, i.e., with the signal being generated by high energy (few kV) electrons bombarding the back of the wafer.
- (3) The channels provide antiblooming control, beside serving as injected charge collectors.
- (4) If necessary narrow metallic stripes can be used for the rows and columns. In thinned structures this may be important, since this could reduce the stress on the wafer.

The proposed design does not have any disadvantages when compared with present ones as regards fabrication, dark current, and capacitance loading.

- (1) The proposed sensor requires the same fabrication steps as current ones, since the channels can be diffused in at the same time as the coupling p-region between gates.
- (2) The dark currents in this sensor should not be larger

than that of a conventional CID. It is true that the channels will occupy and keep depleted a large portion of the surface, however the dark current generated will not flow into the wells but into the channels.

- (3) The capacitance of an output column will not increase, since the channels are parallel to the columns. The capacitance of the rows will increase, but the rows are addressed at slow speeds during blanking and hence this increased capacitance will not cause any degradation in performance.

For the proposed structure to work well it is necessary that when one of the gates of an element is set to zero volts, the surface charge beneath that gate gets transferred to the neighboring gate and is not released to the nearest channel. This could be achieved by making the falltime of the addressing pulse sufficiently long. An alternative method of ensuring that such undesirable injection does not occur would be to use different oxide thicknesses under the two gates of each element. It could be arranged that the collecting channel was next to the gate with the thicker oxide. When injection at an element occurs the other elements in that column have the signal charge in the gate farthest from the channel. However, this stepped oxide structure would lead to more complicated fabrication procedures, and possibly a structure which would be harder to thin down.

TRANSMITTANCE OF
AIR/SiO₂/POLYSILICON/SiO₂/Si STRUCTURES

1. Abstract

The transmittance of Air/SiO₂/Polysilicon/SiO₂/Si structures is calculated in the spectral region between 0.4 and 1.0 μ . A proper choice of the thicknesses of the oxide films can substantially increase the transmittance over a narrow wavelength band or over the entire wavelength region of interest here.

2. Introduction

Polycrystalline silicon films are presently used as the semi-transparent gate electrodes in front illuminated Charge Coupled Device (CCD)¹ and Charge Injection Device (CID)² image sensors. The thin film structure on the gate region of these sensors is shown in Fig. 1. The gate oxide is first thermally grown on the silicon wafer. The polysilicon film is deposited next, usually by chemical vapor deposition, and doped to reduce its resistivity. A top silicon dioxide film is then either deposited or grown thermally on the polysilicon.

In this paper are presented some calculated curves of

Fig. 1. Thin film structure in the gate region of polysilicon-gate image sensors. $m = 1$ denotes the top oxide and $m = 3$ denotes the gate oxide films.

the transmittance of Air/SiO₂/Polysilicon/SiO₂/Si structures, in the spectral region 0.4 to 1.0 μ . In the literature, little attention seems to have been paid^{1,3} in selecting the gate or top oxide film thicknesses when optimizing the transmittance. In our calculations all three films on top of the silicon are taken into account. It is demonstrated that for fixed polysilicon film thickness, the Reflectance/Transmittance (R/T) spectra of such structures are significantly altered by the thicknesses of the oxide films. Dependence of the R/T spectra on the oxide film thicknesses has been noted by Brown and Chamberlain⁴. It is also shown that a substantial portion of the incident illumination is absorbed within the polysilicon film even for 0.1 μ thick films.

3. Analysis

The Matrix method⁵ was used for the calculation of the R/T spectra, for light incident normally on the surface of the sensor. The mathematical expressions were evaluated using a digital computer. The recurrence relations from which the reflectance and transmittance are calculated are given in⁵ and will not be reproduced here.

The values of n and k , the real and imaginary parts of the refractive index, in the wavelength region between 0.4 to 1.0 μ , for the polysilicon films used as the gate electrodes, are not known. In this paper the optical constants of single crystal silicon are used for both the polysilicon films and the silicon wafer. These were taken from⁶ and are shown in Table I, the values at other wavelengths were

TABLE I

OPTICAL CONSTANTS OF SINGLE CRYSTAL SILICON FROM⁶

WAVELENGTH λ , MICRONS	REFRACTIVE INDEX, n	EXTINCTION COEFFICIENT, k
0.386	6.00	0.42
0.410	5.11	0.17
0.440	4.67	0.13
0.475	4.33	0.11
0.515	4.16	0.10
0.560	4.00	0.10
0.62	3.92	0.05
1.00	3.75	0.0008

found by interpolation. The k values for the heavily doped polysilicon films should be significantly larger⁷ than those for pure single crystal silicon; the values of n on the other hand are not expected to differ by more than a few percent^{8,9}. The variation of n and k with wavelength has been included in the calculations. The refractive index of silicon dioxide was assumed real and equal to 1.5^{10,11}. Since the silicon substrate is assumed semi-infinite, the transmittance gives the amount of light that is absorbed within the silicon and contributes to the efficiency of the sensor.

4. Results

A. Effect of the Gate Oxide

In Fig. 2a and 2b the transmittance is plotted against wavelength with the thickness d_3 of the gate oxide as a parameter. The polysilicon film thickness $d_2 = 0.15 \mu$ and the top oxide film thickness $d_1 = 0.1 \mu$ for all the curves in the figure. It is seen that as the gate oxide thickness increases; a) all the peaks shift towards longer wavelengths, b) the amplitude of each peak increases as it is shifted and then decreases after a maximum is reached and c) the peaks in the short wavelength side of the spectrum shift more than peaks occurring at the longer wavelengths. The same behavior is observed for structures with different polysilicon film thicknesses. Values for d_2 between 0.09 to 0.4μ were used.

It is clear from Fig. 2 that the thickness of the gate oxide must be considered when optimizing the transmittance

Fig. 2a. Change in the transmittance spectrum, of the Air/SiO₂/Polysilicon/SiO₂/Si structure shown in Fig. 1, for different gate oxide film thicknesses, in the range 0.1 to 0.26 μ

Fig. 2b. (see caption Fig. 2a).

over a narrow wavelength band, or over a wide spectral range.

A judicious choice of the gate oxide thickness can enhance the transmittance considerably. The commonly used gate oxide thickness of $d_3 = 0.1 \mu$ is a poor choice for maximizing the transmittance. For all polysilicon film thicknesses, an 0.18μ oxide film produced an overall enhancement of the transmittance without shifting the peaks appreciably as compared to the 0.1μ gate oxide film.

B. Effect of the Top Oxide

In Fig. 3 the transmittance is plotted for top oxide film thicknesses $d_1 = 0.0, 0.1$ and 0.18μ . For all curves the polysilicon film thickness d_2 is 0.15μ and the gate oxide film thickness d_3 is 0.18μ . It is seen that the 0.1μ top oxide film produces an overall enhancement of the transmittance. This is not surprising since it is well known that an 0.1μ SiO_2 film on top of silicon is an optimum antireflection coating for the visible region. The 0.1μ top oxide film thus acts primarily to reduce the reflection from the Air/ SiO_2 /Polysilicon portion of the structure of Fig. 1.

Some shift in the peaks occurs for top oxide film thicknesses varying from zero to 0.5μ , but no shift larger than $\pm 0.05 \mu$ was observed for any peak.

C. Effect of the Extinction Coefficient

The extinction coefficient is large in the blue-green region of the spectrum, thus a considerable amount of the incident short wavelength radiation is absorbed within the polysilicon film. This is demonstrated in Fig. 4 where the top curve is for $k = 0$ and in the bottom curve k has the

Fig. 3. Change in the transmittance spectrum as the top oxide film thickness is varied from 0.0 to 0.18 μ .

Fig. 4. Effect of the extinction coefficient k on the transmittance spectrum.

values from Table I. The film thicknesses in this figure are $d_1 = 0.1 \mu$, $d_2 = 0.1 \mu$ and $d_3 = 0.18 \mu$.

The extinction coefficient k of polysilicon is much smaller than the difference of the refractive indices of polysilicon and air or polysilicon and silicon dioxide. The Fresnel coefficients are therefore dominated by their real part and no noticeable phase shift occurs due to k , at any of the interfaces. The positions of the maxima and minima in the R/T spectra are not therefore affected by the value of k and this is clearly demonstrated in Fig. 4.

5. Conclusions

1. The transmittance spectrum will be in error unless the thickness of the gate and top oxide films are included in the calculations.

2. A proper choice of the gate and top oxide film thicknesses can result in substantial increase of the transmittance over a particular wavelength band or over the entire spectral region between 0.4 to 1.0 μ . This latter point is illustrated by the bottom curve of Fig. 4.

3. Considerable absorption occurs within the polysilicon even for very thin films. Since the k values used here are smaller than the values of polysilicon films, in practice the absorption losses will be higher than indicated in the figures. However, the positions of the peaks and valleys are not affected by the values of k , that are likely.

6. References

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APPENDIX III - A

Additional Data for the Transmittance of Air/SiO₂/Polysilicon/
SiO₂/Si Structures

In this Appendix we elaborate on some areas that were not discussed extensively enough in Manuscript III.

A. Effect of Polysilicon

Polysilicon, compared to air and SiO₂, is characterized by a high index of refraction in the wavelength region between 0.4 to 1.0 μ , and as a result dominates the reflectance/transmittance spectra of the air/SiO₂/polysilicon/SiO₂/silicon structures. Small changes in the thickness of the polysilicon film shift the position of the peaks in the transmittance appreciably. The transmittance spectra for polysilicon film thicknesses between 0.11 to 0.25 μ are shown in Fig. 1a, 1b and 1c. The top oxide thickness $d_1=0.1$ μ and the gate oxide thickness $d_3=0.18$ μ for all the curves.

B. Effect of the Extinction Coefficient

In Figs. 1a, 1b and 1c, it can be seen that as the thickness of the polysilicon is increased, the average value of the transmittance in the blue-green region of the spectrum decreases. This occurs because of the absorption of light in the polysilicon layer. The absorption coefficient, α ,

Fig. 1a. Change in the transmittance spectra of Air/SiO₂/Polysilicon/SiO₂/Si structures, for various polysilicon film thicknesses.

Fig. 1a. Change in the transmittance spectra of Air/SiO₂/Polysilicon/SiO₂/Si structures, for various polysilicon film thicknesses.

Fig. 1b. (see caption Fig. 1a.)

Fig. 1c. (see caption Fig. 1a).

is related to the extinction coefficient by $\alpha = 4\pi k/\lambda$, and is appreciable in this region. For the values of k used in these calculations, which are most likely smaller than the values of k for polysilicon, significant absorption occurs within the polysilicon even when its thickness is 0.1μ as it was demonstrated in Fig. 4 in Manuscript III. As the thickness of the polysilicon film is increased, the oscillations in the blue-green region are severely subdued and the transmittance curve becomes smooth. This effect is shown in Fig. 2 where the polysilicon thickness $d_2 = 0.4\mu$.

As discussed in Manuscript III, the position of the maxima and minima in the R/T spectra are not, however, affected by the value of k and this is demonstrated again in Fig. 2.

In Fig. 3 we have plotted the transmittance, T , for different values of k , ranging from $k = 0$ to $k = 5k_{DN}$, where k_{DN} is the extinction coefficient of single crystal silicon measured by Dash and Newman¹². For these calculations $d_1 = 0.1\mu$, $d_2 = 0.2\mu$ and $d_3 = 0.18\mu$. Using the data from this figure, it can be verified that if the transmittance, T , is known when the extinction coefficient is k , then the transmittance, T' , at the same wavelength but for a different k is given to a good approximation by;

$$T' = T \exp \left[(4\pi/\lambda) (k' - k) d \right]$$

where λ is the wavelength in vacuum and d is the polysilicon film thickness. Thus, if the values of k for polysilicon films are different from the values used here, all the figures presented, both in Manuscript III and in this Appendix,

Fig. 2. Effect of the extinction coefficient k on the transmittance spectrum, for thick polysilicon film.

Fig. 3.

Transmittance spectra with k as a parameter.

can still be used by simply correcting for the amplitude of T using the above formula.

C. Effect of Top Oxide

It was stated in Manuscript III that an 0.1μ thick SiO_2 film on top of Si is an optimum antireflection coating in the spectral region between 0.4 to 1.0μ . This is demonstrated in Fig. 4 where we have plotted the calculated reflectance spectra of $\text{air/SiO}_2/\text{Si}$ structures with the SiO_2 film thickness as a parameter. For SiO_2 film thicknesses greater than those shown, the reflectance spectrum for each thickness shows several maxima and minima within the wavelength region of interest here. The calculations are for light incident normally on the structure. The thickness of the Si substrate was assumed infinite and the n and k values from table I were used for Si. For SiO_2 , $n = 1.5$ and for air $n = 1.0$.

Fig. 4.

Calculated reflectance spectra of an $\text{air/SiO}_2/\text{Si}$ structure with the SiO_2 film thickness as a parameter.

Fig. 4. Calculated reflectance spectra of an AIR/SiO₂/Si structure with the SiO₂ film thickness as a parameter.

APPENDIX III - B

The Reflectance and Transmittance Formulae

The Reflectance R_j and Transmittance T_j , for normal incidence, of a structure consisting of j -thin films bounded by two semi-infinite media are given by;

$$R_j = (t_{1(j+1)}^2 + u_{1(j+1)}^2) / (p_{1(j+1)}^2 + q_{1(j+1)}^2)$$

$$T_j = (n_{j+1}/n_0) \left\{ \prod_{i=1}^{j+1} [(1+g_i)^2 + h_i^2] \right\} / (p_{1(j+1)}^2 + q_{1(j+1)}^2)$$

where n_0 is the index of refraction of the incident medium (air in this case) and n_{j+1} is the index of refraction of the exit medium (silicon). $\prod$ indicates product in the above expression.

Recursive formulae for the other quantities are given below. From their definition, the subscript i must take values ≥ 1 only.

For $i \geq 2$,

$$p_{1(i+1)} = p_{1i}p_{i+1} - q_{1i}q_{i+1} + r_{1i}t_{i+1} - s_{1i}u_{i+1}$$

$$q_{1(i+1)} = q_{1i}p_{i+1} + p_{1i}q_{i+1} + s_{1i}t_{i+1} + r_{1i}u_{i+1}$$

$$r_{1(i+1)} = p_{1i}r_{i+1} - q_{1i}s_{i+1} + r_{1i}v_{i+1} - s_{1i}w_{i+1}$$

$$s_{1(i+1)} = q_{1i}r_{i+1} + p_{1i}s_{i+1} + s_{1i}v_{i+1} + r_{1i}w_{i+1}$$

$$t_{1(i+1)} = t_{1i}p_{i+1} - u_{1i}q_{i+1} + v_{1i}t_{i+1} - w_{1i}u_{i+1}$$

$$u_{1(i+1)} = u_{1i}p_{i+1} + t_{1i}q_{i+1} + w_{1i}t_{i+1} + v_{1i}u_{i+1}$$

$$v_{1(i+1)} = t_{1i}r_{i+1} - u_{1i}s_{i+1} + v_{1i}v_{i+1} - w_{1i}w_{i+1}$$

$$w_{1(i+1)} = u_{1i}r_{i+1} + t_{1i}s_{i+1} + w_{1i}v_{i+1} + v_{1i}w_{i+1}$$

For $i = 1$, the above constants become;

$$p_{12} = p_2 + g_1 t_2 - h_1 u_2, \quad q_{12} = q_2 + h_1 t_2 + g_1 u_2$$

$$t_{12} = t_2 + g_1 p_2 - h_1 q_2, \quad u_{12} = u_2 + h_1 p_2 + g_1 q_2$$

$$r_{12} = r_2 + g_1 v_2 - h_1 w_2, \quad s_{12} = s_2 + h_1 v_2 + g_1 w_2$$

$$v_{12} = v_2 + g_1 r_2 - h_1 s_2, \quad w_{12} = w_2 + h_1 r_2 + g_1 s_2$$

For $i \geq 1$,

$$p_i = \cos(\gamma_{i-1}) \cdot \exp(a_{i-1})$$

$$q_i = \sin(\gamma_{i-1}) \cdot \exp(a_{i-1})$$

$$r_i = [g_i \cdot \cos(\gamma_{i-1}) - h_i \cdot \sin(\gamma_{i-1})] \cdot \exp(a_{i-1})$$

$$s_i = [h_i \cdot \cos(\gamma_{i-1}) + g_i \cdot \sin(\gamma_{i-1})] \cdot \exp(a_{i-1})$$

$$t_i = [g_i \cdot \cos(\gamma_{i-1}) + h_i \cdot \sin(\gamma_{i-1})] \cdot \exp(-a_{i-1})$$

$$u_i = [h_i \cdot \cos(\gamma_{i-1}) - g_i \cdot \sin(\gamma_{i-1})] \cdot \exp(-a_{i-1})$$

$$v_i = \cos(\gamma_{i-1}) \cdot \exp(-a_{i-1})$$

$$w_i = -\sin(\gamma_{i-1}) \cdot \exp(-a_{i-1})$$

$$a_{i-1} = (2 k_{i-1} d_{i-1}) / \lambda, \quad \gamma_{i-1} = (2 n_{i-1} d_{i-1}) / \lambda$$

$$g_i = (n_{i-1}^2 + k_{i-1}^2 - n_i^2 - k_i^2) / [(n_{i-1} + n_i)^2 + (k_{i-1} + k_i)^2]$$

$$h_i = 2(n_{i-1}k_i - n_i k_{i-1}) / [(n_{i-1} + n_i)^2 + (k_{i-1} + k_i)^2]$$

where, λ is the wavelength of the incident radiation in vacuum, n_m and k_m are the real and imaginary parts of the complex index of refraction, defined as $N_m = n_m - ik_m$, of the m th medium and d_i is the thickness of the i th layer.

FINE STRUCTURE IN THE OPTICAL ABSORPTION EDGE OF SILICON

1. Abstract

Details of the structure in the indirect optical absorption edge of silicon were studied by measuring the dependence of the photocurrent in p - n junctions on the energy of the incident photons. The measurements were made at room and higher temperatures for photon energies $0.75 \text{ eV} < h\nu < 1.05 \text{ eV}$. The sensitivity of the method enabled high resolution measurements in the absorption tail. At room temperature, thresholds were found at $\sim 0.9 \text{ eV}$, 0.99 eV , and 1.025 eV . The derivative of the response showed extensive fine structure in this tail. The 70 and 10 phonon assisted transitions to the ground and excited states of the exciton, reported by Hinkley and Mahony in the phonon excitation region, were here seen with phonon absorption, occurring around 1.054 eV and 1.065 eV . There was additional structure in this region, whose origin is not known.

2. Introduction

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2. Introduction

The optical absorption spectrum near the indirect band gap of silicon was measured by Macfarlane, McLean, Quarrington and Roberts¹ (MMQR). The absorption arises from indi-

rect allowed transitions with momentum being conserved by the emission or absorption of a phonon. The electron-hole pairs created may either be unbound or exist as excitons with binding energy $\epsilon(n)$, where n denotes the n^{th} exciton level. The theory for the spectral dependence of the absorption coefficient α for such transitions was developed by Elliott². A summary of the theory and an analysis of the experimental results can be found in the extensive review by McLean³.

The agreement between theory and experiment was, in general, excellent, but there were a few discrepancies. The experimentally determined value of 5.5 meV for $\epsilon(1) - \epsilon(2)$ led to an estimate of ~ 7 meV for the exciton Rydberg, whereas the value calculated from theory by McLean and Loudon⁴ was 14 meV. Another discrepancy involved the absence of any contribution from the longitudinal optical (LO) and longitudinal acoustic (LA) phonons, although transitions aided by these phonons are allowed. Finally, the experimental results showed absorption at photon energies lower than the minimum threshold energy for one-phonon aided transitions. The value of α in this absorption tail was very small and could not be accurately determined from the transmission technique used by MMQR.

Dean et al⁵ measured absorption and luminescence at low temperatures and obtained results in substantial agreement with MMQR. The higher resolution they had enabled them to see additional structure near the thresholds which they attributed to splitting of the ground state of

the exciton by valley-orbit interactions. Recently, Shaklee and Nahory⁶ presented results from wavelength derivative type experiments which indicated that the energy separation between the ground and first excited state of the exciton was 11.0 ± 0.2 meV, and the binding energy of the exciton was 14.7 ± 0.4 meV. In addition they pointed out theoretical reasons that preclude valley-orbit splitting of the exciton ground state. They identified the additional structure near the threshold for absorption of a photon with the emission of a transverse optical (TO) phonon as due to LO phonon assisted transitions.

In this work we have investigated the dependence of the photocurrent generated in silicon p-n junctions on the energy of the incident photons. The ratio of the photocurrent to light intensity was measured and corrected to give the response R for constant photon flux. R was taken to be proportional to the absorption coefficient α , which is true as long as the incident light is weakly absorbed. This point is discussed further in Section 3. The sensitivity of the present method arises from the fact that α is directly related to the measured photocurrent, whereas in transmission or reflection experiments it is related to the small difference between two large measured quantities. The method proved particularly useful for measuring accurately the absorption tail. Measurements were also made in the TO phonon absorption region, i.e. where the absorption of a photon is accompanied by the absorption of a TO phonon.

The results can be summarized as follows:

(1) At room temperature there are three prominent thresholds in the absorption tail. The first, which is not too well defined, occurs at 0.91 eV, the second at 0.99 eV and the third at 1.026 eV. The latter two were observed by MMQR. The first lies beyond their range of measurement.

(2) The dependence of α on the photon energy following the latter two thresholds is not in agreement with that found by MMQR. This is due to the fact that there are a number of other lesser thresholds present.

(3) In the region from 1.015 eV to 1.045 eV the results are accurate enough to permit numerical differentiation of the data to get the derivative of the response. The derivative plots show fine structure which has not previously been seen.

(4) The threshold energies and temperature dependence of α suggest that two and three phonon processes give rise to the absorption tail. The fine structure indicates that the phonons participating are not just from the Γ point and Δ symmetry direction of the Brillouin zone, but from throughout the zone.

(5) In the TO phonon absorption region our experiments show that there is more structure in the absorption coefficient than accounted for by the theory. Furthermore it is evident from our data that the fine structure, in the region following TO phonon aided transitions to the ground state of the exciton, has been the source of disagreement in previous determinations of $\epsilon(1) - \epsilon(2)$. The value of $\epsilon(1) - \epsilon(2)$ we find is in excellent agreement with that

found by Shaklee and Nahory⁶.

(6) Finally, participation of LO phonons in the absorption process, first observed by them, is confirmed in these experiments.

3. Experimental Details

Commercially available large area silicon p-i-n photo-diodes and epitaxial n-on-p wafers were used. The n epitaxial layer was either phosphorous or arsenic doped and its thickness was approximately $15\ \mu$. The substrate was always boron doped and its thickness was of the order of $250\ \mu$. The resistivities of the epitaxial layer and the substrate were the same in each sample and ranged in value between 1 and $10\ \Omega\text{ cm}$ for different wafers. The measurements were made with the samples at room temperature and at higher temperatures.

A tungsten lamp, chopper and a double-grating monochromator with a spectral bandwidth of 0.5 nm were used to provide a chopped light beam. A beam splitter was used at the exit slit to irradiate the sample and a reference thermopile. The voltage developed across a load resistor connected across the sample was measured with one lock-in amplifier and the output from the thermopile on a second lock-in amplifier. The ratio of the outputs from the two lock-in amplifiers was measured with a ratiometer and recorded. This reading was multiplied by $1/\lambda$ to give R the response at different wavelengths for constant photon flux. The illumination level was kept low enough that the photocurrent

varied linearly with light intensity. R is then proportional to the total number of electron-hole pairs created within an effective collection region near the junction.

The reasons for taking R to be proportional to α are as follows. In traversing a distance l the light is attenuated from intensity I_0 to $I(l)$ where $I(l) = I_0 \exp(-\alpha l)$. In the region of interest α is less than 1 cm^{-1} , and with wafer thickness of 250μ , we have $\alpha l \ll 1$. Hence $I_0 - I(l) \approx I_0 \alpha l$ and the light absorbed in the effective collection region is proportional to α . We assume the quantum efficiency to be independent of photon energy. This assumption is substantiated by the photoconductive measurements made on germanium by Moss and Hawkins⁷. They found that in the absorption tail their calculated absorption coefficient from photoconductive measurements was identical to that obtained by Macfarlane et al⁸ from transmission measurements. These expectations were experimentally verified by shining the light through either surface of the wafer. Despite the large asymmetry of the junction depth relative to the sample surfaces the response obtained was identical. Again there were five room temperature measurements of the response in the region from 1.015 eV to 1.045 eV. These included the p-i-n diode and several different epitaxial wafers. All runs gave results which were the same to within a multiplying factor.

The monochromator which was used in the experiments had a linear wavelength scale. Where extreme accuracy was required, the data points were taken by manual setting of this scale. The derivative spectrum was obtained by taking

the difference between the responses at wavelengths separated by 1 nm. This gave $\Delta R/\Delta \lambda$ rather than $\Delta R/\Delta h\nu$, but for small ranges of λ one is proportional to the other. The value of 1 nm for $\Delta \lambda$ was chosen for highest resolution consistent with discrimination against noise.

A number of steps were taken to make sure of the genuineness of the structure seen in the derivative plots. The correctness and linearity of the monochromator scale were checked by measuring persistent lines and doublets in atomic spectra. The accuracy and reproducibility of the settings was found to be better than 0.02 nm. Effects due to the grating and beam splitter were eliminated by using a near-infrared transmitting filter and polarizer in front of the entrance slit of the monochromator. This was verified by repeating the experiments with a second thermopile and a Ge photodiode in place of the sample. In neither case was any structure observed. The use of several different samples has already been mentioned. Finally the samples were heated and measurements made at temperatures 10 to 50°C above room temperature. Large changes in the shape of the absorption curve are not expected for such small temperature changes and, since the energy gap decreases with temperature, all genuine structure should appear displaced to longer wavelengths. All structure discussed in Section 4 showed this expected behavior. Incidentally, there was no structure found that was caused by the system.

4. Experimental Results and Discussion

A. The Absorption Tail

The method was useful for photon energies below 1.077 eV. At higher energies the absorption was so large that R did not increase linearly with α , resulting in loss of sensitivity. The only information beyond 1.077 eV that was obtained was the energy at which onset of transitions to the ground state of the exciton occurred with the simultaneous emission of a TO phonon. This threshold gave an easily identifiable peak in the $\Delta R/\Delta \lambda$ versus λ curve, and occurred at 1.1692 eV. The peak due to the onset of transitions to the ground state of the exciton involving the simultaneous absorption of a TO phonon was easily established and occurred at 1.0538 eV. The average of these two energies gives the room temperature (296 K) exciton indirect gap, i.e. the energy gap minus the exciton binding energy, as 1.1115 ± 0.5 eV. Half of the spacing between the two peaks gives the TO phonon energy to be 57.7 ± 0.5 meV. These values are in excellent agreement with previous results.

A semi-logarithmic plot of the response against photon energy in the region of the absorption tail is shown in Fig. 1. There is a measurable response at 0.775 eV, and it increases by about an order of magnitude on going to 0.90 eV. Not much can be said about the shape of the curve in this region, as the signal was too small to be measured accurately. The signal was just above the noise level at the low energy end and was determined with about 6 percent

Fig. 1. The photoresponse of silicon p-n junction in the absorption tail. For reference the arrow points to the TO phonon absorption threshold.

accuracy at 0.90 eV.

Towards higher energies the response rises sharply, with clearly observable thresholds at 0.91 eV and 0.99 eV. Various simple expressions were tried for describing the response between these two thresholds. The formula

$$R = 4.9 + 1.09 x + 0.56 \exp x \quad (1)$$

with $x = \beta(h\nu - h\nu_0)$, $\beta = 67 \text{ eV}^{-1}$, and $h\nu_0 = 0.925 \text{ eV}$ gave a good fit to the data in the region between 0.925 eV and 0.983 eV. According to this formula the response is an exponential, superimposed on a background consisting of a constant and a linear term. The data in this region at $T = 331 \text{ K}$ was fitted by

$$R = 10.8 + 2.30 x + 1.20 \exp x \quad (2)$$

where β was again 67 eV^{-1} , and $h\nu_0$ was 0.916 eV. The shift of 9 meV in $h\nu_0$ is in agreement with the change in the energy gap due to the rise in temperature. The latter was measured from the shift of the TO absorption peak and was 9 meV. It is of interest to note that the raising of the temperature increased the response by a factor of about 2.15.

The response between the second threshold and the TO absorption region is shown in greater detail in Fig. 2. The top curve is data obtained at 331 K and goes with the upper scale; the bottom curve is the room temperature data and goes with the lower energy scale. The top scale is shifted towards lower energy by 9 meV relative to the bottom scale to offset the change in energy gap with

Fig. 2. Expanded plot of response versus photon energy in the region from 0.96 to 1.06 eV. The upper curve has been shifted by 9 meV towards lower energies to offset the decrease in energy gap with tempera-

temperature; the arrow points to the TO absorption threshold for both the room and high temperature data. In the region from 0.99 eV to 1.05 eV, MMQR fitted their room temperature data with an expression of the form

$$\alpha = a(h\nu - 0.989)^2 U(h\nu - 0.989) + b(h\nu - 1.021)^2 U(h\nu - 1.021) \quad (3)$$

where $h\nu$ is measured in eV, a and b are constants and U is the unit step function. This implies that the absorption rises as the square of the energy from thresholds at 0.989 eV and 1.021 eV. In trying to fit such an expression to our measurements one of the problems was to find the correct extrapolation of the response at lower energies and subtract it from the actual data. The mechanism giving rise to the absorption below 0.98 eV is not known, and there is some concern about extrapolating an increasing exponential. For want of a better procedure it was assumed that the same mechanism continued to operate beyond 0.98 eV, and the extrapolated value R_{ext} was found from Eq. (1). Fig. 3 shows the square root of the difference between the actual and extrapolated values against photon energy. The bottom two curves are room temperature data for two different samples, the top curve is for a heated sample. The energy scale for the heated sample has again been shifted by 9 meV relative to the scale for the room temperature data. Between (a) and (b) one can draw approximate straight lines, with nearly common origin for the three curves. The threshold thus established is 0.982 eV at room temperature,

Fig. 3. Square root of the difference between the actual photoresponse and the extrapolated low energy response from Eq. (1), plotted against photon energy. The bottom curves are for different samples at room temperature. The top curve is for a sample at higher temperature, and has been shifted by 9 meV to facilitate comparison with the lower curves. The arrows point to structures seen in all

as compared to the value of 0.989 eV found by MMQR. There is a break in the curve at point (b). This occurs at 1.0213 eV, and coincides with the second threshold of Eq. (3).

To a very rough approximation, one can fit the tail in this region according to the dependence suggested by MMQR. The agreement, however, is not very good. There are really no straight line portions in Fig. 3. There is more structure in the response than indicated by Eq. (3). The arrows point to places where a break is apparent in every one of the curves. The deviations between the experimental values and the curve calculated to give the best fit are well beyond the experimental error.

The structure in the response can be seen more clearly in the derivative plots. As explained in Section 3, it was more convenient for handling the data to choose λ as the independent variable instead of $h\nu$.

In Fig. 4 $\Delta R/\Delta \lambda$ is plotted against decreasing wavelength in the vicinity of the threshold at 0.99 eV. The top curve is the room temperature data, and the bottom curve is the data at higher temperature for the same sample. The scales are offset, as in the previous figures, to facilitate comparison. The similarity of these curves is evident. It is clear that not one but two peaks are present. The photon energies corresponding to these peaks are 0.989 eV and 0.994 eV for the room temperature data, and are shifted by 9 meV towards lower energies at higher temperature.

Fig. 5 shows $\Delta R/\Delta \lambda$ against decreasing wavelength in the region from 1.22 μ to 1.18 μ . The points plotted are the

Fig. 4. $\Delta R / \Delta \lambda$ against decreasing wavelength near the threshold at 0.990 eV (1.250μ). The bottom curve has been shifted by 9 meV towards lower energies.

Fig. 5. $\Delta R/\Delta \lambda$ showing fine structure in the absorption tail. The points are experimental values, the line is drawn through features seen consistently in all runs. The photon energies corresponding to the arrows are shown in the figure.

average of five runs with each run showing essentially the same features, while averaging eliminated some of the noise. There are several features with different and well defined shapes; the photon energies corresponding to these features are noted in the figure. It is worth noting that the step increase denoted by (a) occurs at 1.0256 eV, which corresponds to point (c) of Fig. 3. Beyond (h) the derivative is dominated by the TO phonon peak and its thermal broadening.

The mechanism giving rise to the absorption tail and the fine structure is not known⁹. We can rule out the effects of the electric fields in the space-charge region, as the diodes gave the same results with and without applied bias. Although we cannot completely disregard impurity effects, the fact that As and P doped wafers with resistivities ranging from 1 to 10 ohm-cm gave identical results suggests that impurity effects are negligible. The known levels introduced by the impurities do not correlate well with the observed energies of the structure. Our results are more in accordance with multiphonon effects.

The conduction band minima in silicon occur in the [100] direction of the Brillouin zone, with the magnitude of the wave vector being about 0.85 times the magnitude at the zone boundary. The optical and acoustic phonons with this wave vector are the ones involved in the one-phonon aided transitions; the TO phonon has energy 57 meV, and the TA phonon has energy 18 meV. It is possible to have two, and higher order, phonon aided processes, provided the sum of the wave vectors of the phonons is equal to the above value. As

pointed out by McLean³ one can have transitions with the simultaneous absorption of the above TA or TO phonon and an O phonon (i.e. a zero wave vector optical phonon). The energy of the O phonon is 63 meV. Two phonon aided transition thresholds would be expected at 0.991 eV (for TO + O) and at 1.031 eV (for TA + O); these agree approximately with the data of Fig. 3. The observed threshold at 0.911 eV could be attributed to absorption of three phonons (O + O + TO). The temperature dependence of the absorption in the region from 0.925 eV to 0.983 eV is in agreement with this assumption. The probability of the simultaneous absorption of three phonons is given by the product of the occupation numbers n_i of each of the phonons, where

$$n_i = [\exp (h\omega_i/kT) - 1]^{-1}$$

and k is Boltzmann's constant. From this the ratio of the absorption at 331 K to that at 296 K is calculated to be about 2.17 as compared to the ratio of 2.15 between Eqs. (1) and (2). Two phonon structure has been previously reported^{10,11}. These interpretations were concerned with identifying peaks in derivative type experiments with combinations of TO, TA, O and S phonon energies, where S is the phonon connecting different conduction band minima. The detailed nature of the structure in our results shows that choosing phonons from only some high symmetry points and directions is inadequate. Calculations of two phonon effects using a wider sampling of phonons has been done only for Raman scattering, where the sum of the momenta of the

phonons is zero. The two phonon spectra have been observed both in Raman effect¹² and electron energy loss measurements¹³. Calculations along those lines for the absorption edge might be useful in analyzing the present experimental results.

B. TO Phonon Absorption Region

It was pointed out in the introduction that there is some disagreement in the literature with regard to the fine structure in the region of TO phonon aided transitions and its interpretation. The data obtained in the present measurements in this region are shown in Fig. 6 where $\Delta R/\Delta\lambda$ has been plotted against decreasing wavelength. The solid curve in the figure is drawn through those features that are seen with the same shape and relative amplitude in all the results, i.e. different runs on different samples. At higher temperatures these features were present, correctly shifted towards lower energies.

Following the interpretation reported by Shaklee and Nahory⁶, we identify the peaks at (a) and (b) as the thresholds for transitions to the ground state of the exciton with the absorption of a TO and an LO phonon respectively. The peaks (d) and (e) correspond to the TO and LO phonon aided transition to the first excited level of the exciton. The energy separation between the ground and first excited level of the exciton is found to be 11.9 ± 0.5 meV. The energy of the LO phonon is calculated to be 55.9 ± 0.5 meV. The values for the phonon energies and exciton levels obtained are in excellent agreement with the values found by Shaklee and Nahory in the phonon emission region.

Fig. 6. The derivative of photocurrent with respect to wavelength for a silicon p-n junction, plotted against decreasing λ . The points are measured values of $\Delta R / \Delta \lambda$, the line is drawn through the features seen consistently in all runs. The peaks denoted by long arrows correspond to features seen by previous investigators.

But there is a discrepancy between theory and experiment. From the theory² it would be expected that the derivative of α would have a $-\frac{1}{2}$ power dependence on the energy from the threshold at (a) to about 12 meV away; and a $\frac{1}{2}$ power dependence on the energy from a threshold 14 meV away from (a). Clearly this is not what happens. Instead at (c), about 6.4 meV away from (a), there is a step increase in the derivative and, neglecting the fine structure, there is an almost linear increase commencing about 10 meV away from (a). This behavior is in good qualitative agreement with the results of MMQR. They interpreted a step-like increase 5.5 meV away from the TO phonon peak as the onset of TO phonon aided transitions to the first excited state of the exciton, and the almost linear increase in the derivative commencing about 10 meV away from the TO peak as due to band transitions. Dean et al⁵ found a step increase about 7.5 meV from the TO peak with band transitions commencing about 4.5 meV after the step. They also interpreted the step increase as the onset of TO phonon aided transitions to the first excited state of the exciton. Finally, in Fig. 2 presented by Shaklee and Nahory⁶ the derivative in the region of TO phonon emission shows a definite increase, though not step-like, commencing about 7 meV away from the TO peak. It is clear from their figure and Fig. 6 of the present work that the increase is not due to background contributions.

Shaklee and Nahory's interpretation of the fine structure is probably the correct one, as it leads to good agreement between the experimental and theoretical values

of the exciton levels, and gives the right energies for the peaks in Fig. 6. There are some features in Fig. 6 which do not correspond to any peaks noted by Shaklee and Nahory.

Some of these, like the structure near (f), could be due to multiphonon effects. These effects would be more prominent in the present measurements which were done at elevated temperatures and in the phonon absorption region. However, the increase in the derivative 7 meV away from the TO peak cannot be due to two phonon effects or impurity effects. If it were it would appear symmetrically placed with respect to the exciton indirect gap. But, as has been pointed out, it has been observed on the high energy side of the TO peak in all the experiments, at low and high temperatures, and both in the phonon emission and phonon absorption regions. It thus appears to be related to the TO peak, and consequently to the exciton, but at the moment there is no explanation for it.

5. Footnotes

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$$\alpha \sim [h\nu - \epsilon_g + \epsilon(n) \pm h\omega_i]^{\frac{1}{2}} U(h\nu - \epsilon_g + \epsilon(n) \pm h\omega_i)$$
 where $h\nu$ and $h\omega_i$ are the photon and phonon energies, ϵ_g is the energy gap, and U the unit step function. The upper sign corresponds to the absorption and the lower to the emission of the phonon. The sharp rise from a threshold will be smoothed out by thermal broadening of the exciton levels. The absorption coefficient due to the creation of unbound hole-electron pairs has the form

$$\alpha \sim [h\nu - \epsilon_g \pm h\omega_i]^{3/2} U(h\nu - \epsilon_g \pm h\omega_i)$$
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INTRODUCTION AND REVIEW OF THE PROBLEM

Solid state image sensors consist of a matrix of integrated light sensing elements. Each element may contain one or more basic solid state devices and is coupled to a number of electrodes. Unlike the electron beam scanned pick-up tubes, in solid state image sensors the signal created by the light at each matrix site is read from the corresponding element by applying appropriate voltages on the electrodes with external circuits. Compared to conventional TV cameras, these image sensors offer the potential advantage of reduced weight, volume and power consumption; low voltage level operation; compatibility with other integrated circuits and devices; digital scanning and accurate registration; reduced cost and long life.

Significant success in developing these sensors came rapidly in the 1960's. A group at RCA Laboratories used all thin film techniques to fabricate 180×180 element¹ and 256×256 element^{2,3} photoconductor arrays scanned by thin film circuits. A research group at Westinghouse Defense and Space Center developed silicon mosaics, making 50×50 , 100×128 and 200×256 phototransistor element arrays⁴.

The discovery⁵ of the charge storage mode of operation for the p-n junction aided greatly in understanding the operation and improving the performance of phototransistor arrays.

It also led to the design of a number of different types of silicon image sensors including the photodiode-bipolar transistor (PBT)⁶ structure. Area arrays utilizing the PBT element have many advantages compared to phototransistor arrays. In Manuscript I a detailed theoretical and experimental study is presented of PBT area arrays.

A big advance in the design of solid state image sensors was made by Sangster⁷ who suggested that, instead of a switch at each element, the signal could be handed bucket-brigade fashion down the line⁸. The invention by Boyle and Smith⁹ that this signal could be minority carrier charge stored in the inversion layer of an MOS capacitor started the development of charge coupled devices (CCD's)^{10,11}.

Shortly afterward the charge injection device (CID) was developed by Michon and Burke¹². The CID image sensor is similar to the CCD in that it employs the charge transfer principle of the charge coupled devices and in that it involves only MOS technology. It differs from CCD arrays, however, because it is X-Y and randomly addressable as are the PBT, phototransistor and photodiode arrays.

Considerable experimental work has been reported on CID image sensors¹³. However, no theoretical work has been published. In Manuscript II, the efficiency by which photo-generated minority carriers are collected is investigated theoretically using numerical techniques. The output signal

as function of illumination level is also studied.

In CID as well as CCD and PBT image sensors, in which the light is incident from the front or electrode side, polysilicon electrodes are used¹¹. If the polysilicon layers are sufficiently thin they will be fairly transparent to the visible radiation even though silicon absorbs strongly in that region. In Manuscript III theoretical calculations are presented of the losses due to interference effects in air/SiO₂/polysilicon/SiO₂/Si structures and of the losses due to absorption of the incident radiation by the polysilicon electrodes.

In Manuscript IV experimental results are presented of the absorption of light by silicon near the absorption edge of silicon.

The closed circuit TV system that was built to test the PBT arrays is described in Appendix I-A.

In Appendix II-A is presented the numerical analysis for the solution of the 2-dimensional diffusion equation. A design for a CID type image sensor that can be operated in the back illumination mode is discussed in Appendix II-B.

Additional calculations on the transmittance of air/SiO₂/polysilicon/SiO₂/silicon structures are given in Appendix III-A. The Reflectance and Transmittance formulae¹⁴ for a j-film structure and derived from the application of the matrix method are shown in Appendix III-B.

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