NUCLEAR INSTRUMENTS AND METHODS 96 (1971) 3–8; © NORTH-HOLLAND PUBLISHING CO.

AUTOMATIC GAIN-CONTROLLED AMPLIFIER FOR BEAM PROFILE MONITORS*

P. ROGERS, D. POTTER and V. RADEKA

Brookhaven National Laboratory, Upton, N.Y., U.S.A.

Received 10 May 1971

An amplifier for beam profile scanners of the moving-wire type has been developed in which light controlled resistors vary the gain so as to provide constant peak amplitude. The peak output signal is independent of the intercepted beam current over the range 4×10^{-3} to 4×10^{-10} A. The rms noise is less than 8×10^{-12} A, and the rise time at maximum gain is 20 μsec. The control loop maintains approximately constant gain during a single scanner period, yet can change the gain over several decades in one second. At any given gain the amplifier response is linear. With this amplifier in operation, it is not necessary to change the gain manually, as the beam intensity varies over the range of beam currents commonly encountered with particle accelerators.

1. Introduction

The operation of modern accelerators at high beam currents of accelerated ions requires an accurate knowledge of both beam profile and position at various critical points of the accelerator system. In order to observe the beam intensity at the target while simultaneously studying its profile and position at other locations, a low duty-cycle high-speed scanning system is used because it does not change appreciably the intensity or character of the beam at the target. The speed of the scanner should be sufficient to allow dynamic observation of the tuning conditions of the various beam transport elements, such as quadrupoles and steering magnets, and its corresponding display device should be relatively free of flicker so that eye fatigue is minimized. An oscillating-loop scanner was described by Wegner and Feigenbaum[1]). In this scanner a small water-cooled tube is arranged in the form of a loop so that both X and Y beam profiles are scanned with a single scanner. The scanner has an aperture arrangement with a light source and a photodiode that generates fiducial marks on the display corresponding to the X and Y position of the axis of the beam pipe.

The problem in observation of the beam profile and position is that the beam current may change by several orders of magnitude when the accelerator operating conditions are changed. In order to accommodate these large current changes, an operator would have to select the gain of a current amplifier manually every time a parameter is changed that alters the beam current. The range of currents intercepted by the scanner extends over some six orders of magnitude, from about 10^{-9} to 10^{-3} A. The purpose of this paper is to describe a new type of current amplifier with automatic gain control (AGC) that will operate over this range while maintaining a constant output amplitude without distortion of the observed beam profile. The response time constant of the automatic gain control is from 0.1 to 0.2 sec over the whole range of input currents.

2. Basic considerations

The basic approach is to sample the waveforms corresponding to beam profile at their maximum and to control the gain of the current amplifier so that the waveform amplitude remains nearly constant. The requirements on the time response, signal range, baseline stability and noise make the implementation of this basic approach difficult. The automatic gain control should respond in a time short compared to the response time of the human operator. The rise time of the current amplifier must not contribute to the distortion of the beam profile. Noise and baseline shifts should be small compared to a signal at the lower end of the current range.

The AGC response is limited by the sampling frequency, f_s, which is determined by the scanner oscillation frequency, f_o. The scanner[1]) oscillation frequency is about 20 Hz. There are two X profile and Y profile scans each per cycle. Since X and Y profiles can be different in shape and amplitude, two samples of the profile amplitude can be used for automatic gain control for one period of the scanner. This sets an ultimate lower limit on the AGC response time constant, approximately determined by

$$\tau_s \approx 2/f_s = 1/f_o,$$

with $f_o = 20$ Hz, $\tau_s = 50$ msec.

* Work supported by the U.S. Atomic Energy Commission.

The large dynamic range requires that the transimpedance of the AGC current amplifier span the range from 10^3 to about 10^9 Ω. For minimum baseline drift and noise it would be best if this range were covered by the feedback resistor of one operational amplifier. This is not possible because of signal rise time requirements. The current amplifier should not contribute to the distortion of the observed beam profile more than the scanner contributes due to the finite thickness of the scanner wire. The rise time of the amplifier should be equal to or less than the time the scanner takes to move by one wire diameter. For the described scanner[1]) this is about 60 μsec at an oscillation frequency of 20 Hz. This imposes an upper limit on the value of the feedback resistance because of unavoidable parasitic capacitance. Therefore, at least two stages are required to which automatic gain control is applied.

Large dynamic range leaves little choice with regard to the gain control element. Light-controlled resistors based on variation of conductivity of certain semiconductors with light appear to be the most suitable gain control elements in this case. A principal disadvantage of light-controlled resistors is their slow response, which presents additional time constants in the control loop and a limitation on the overall time response. Continuous automatic gain control over the whole range was preferred to the discrete (switched) automatic gain control because we concluded that the former would result in a simpler solution. The complete current amplifier with automatic gain control is shown in fig. 1.

Current signal amplification is accomplished in two gain-controlled stages. The third stage serves as a fixed gain, second-order, low-pass filter. An inverter is used at the output. Input can be of either polarity corresponding to positive or negative ions in the beam. Signals at the input and the output of the inverter are mixed (network with diodes D_5 and D_6) so that a negative signal arrives in both cases at the control circuit.

A peak stretcher with a constant discharge rate was chosen as a simple approximation to the zero-order, sample-and-hold circuit. There are two unavoidable lag elements in the gain control loop. One is presented by the finite sampling frequency, and the other is the light-controlled resistor. The light-controlled resistor has a rather complicated response which varies over the range of light and resistor values. Its very simplified small signal response can be represented as a single pole network with a time constant varying from 20 to 50 msec over the range of resistance used here. If an integrator is included in the loop, a closed loop response with an equivalent time constant varying from 0.1 to 0.2 sec over the whole range can be achieved.

Due to a large signal dynamic range, the error signal is saturated except for a very narrow region around equilibrium. There are two approaches to assure loop stability in such a nonlinear feedback loop. One is to reduce the small signal gain near equilibrium, while keeping constant the maximum slew rate of the control variable (integrator output). This could be accomplished by a simple diode-resistor network between emitters of T_5 and T_6. The other approach is to apply a correction proportional to the rate of change of the signal in addition to the absolute value of the error. This quantity is not available since the output of the stretcher is either zero or saturated for large signal changes. An alternative way, applied here, is to subtract from the error a signal proportional to the rate of change of the control variable. This is accomplished by the network C_f, R_f in the error amplifier-integrator circuit, as shown in fig. 1. One of these approaches is essential to achieve the stability and minimum response time. Without such a network the stability could be achieved only at the expense of a very much increased response time, by reducing the gain of the error amplifier and the integrator slew rate.

3. Light-controlled resistor

The resistance of light-controlled resistors varies over several orders of magnitude with light intensity. The current-voltage relationship of the resistance element is linear up to several volts, so that the signal dynamic range for constant resistance can be large. Electronic devices, like bipolar transistors and field-effect transistors, can be used as gain varying elements in analog multiplier configurations. Their response is much faster, but the range over which the gain can be varied is limited, as well as the signal dynamic range. With these devices a number of stages with gain control would have to be used, each with a small gain range. The output signal from the first stage would have to be small to prevent saturation at larger input signals when the gain is decreased to its lowest value. Noise and drifts would then be high compared to the lowest value of the signal.

The light source in commercially available light-controlled resistors used here is a small incandescent lamp. The resistance element is a semiconductor (e.g. cadmium selenide) plate with ohmic contacts. Resistance as a function of lamp voltage for a typical light-controlled resistor (Hewlett-Packard 5082-4510) is

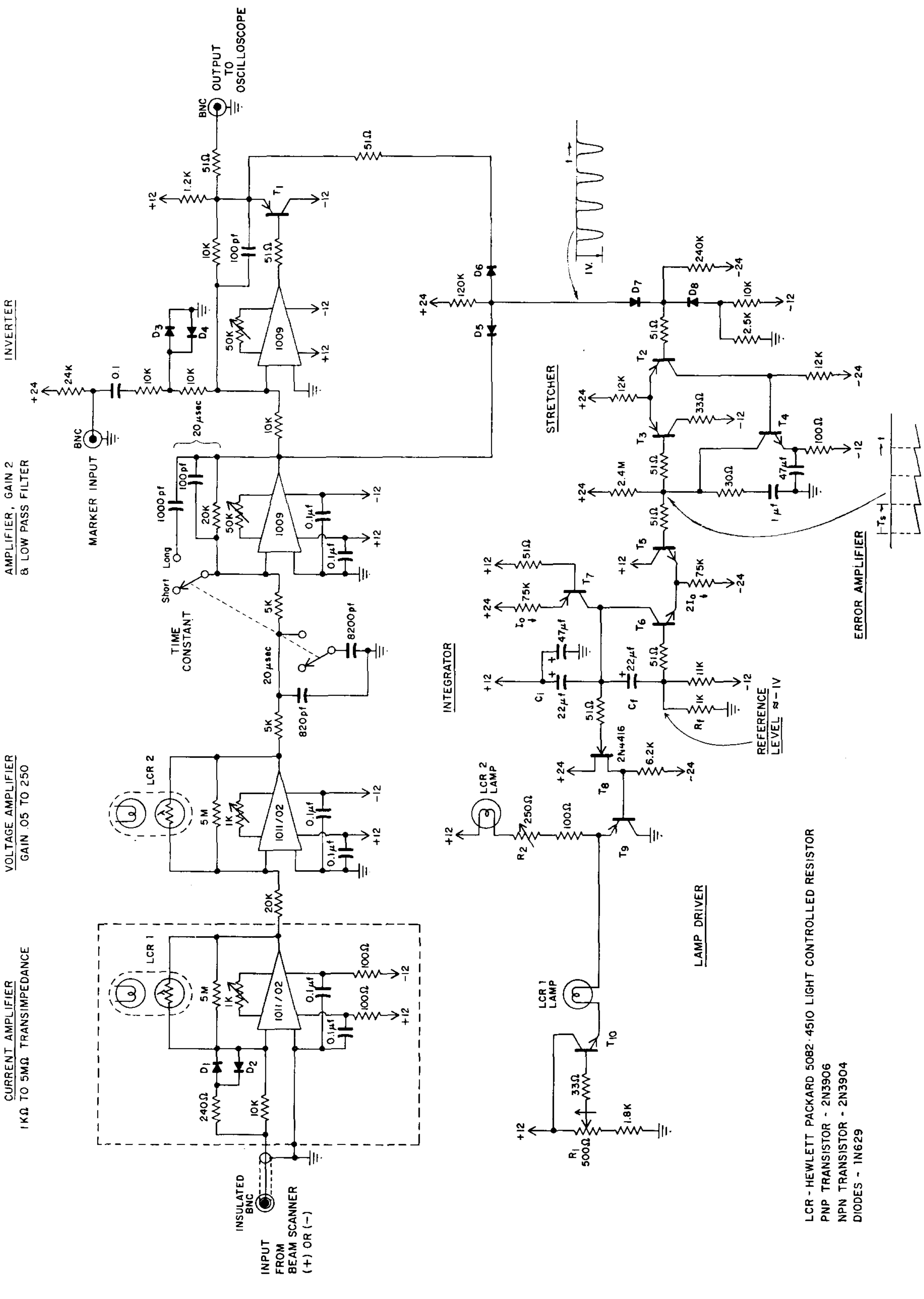

Fig. 1. Current amplifiers with automatic gain control.

shown in fig. 2. The capacitance across the resistance element is about 2 pF. This capacitance limits the maximum value of the resistance for a given amplifier response time. Assuming the maximum value of the feedback network time constant of 10 μsec, the maximum value of the resistance is 5×10^6 Ω. The minimum value is about 10^3 Ω at a lamp voltage of 10 V.

The coupling between the control variable (lamp voltage) and the resistance element is very low. The capacitance between the lamp and the resistance element is about 10^{-2} pF, and the insulation resistance is greater than 10^{12} Ω.

The time response of the device is quite complex. It consists of two components, the lamp filament thermal time constant and the charge recombination time constants in the resistance element. The latter dominates at higher resistance values. Devices with a dominant time constant of less than 50 msec should be used in this circuit. The details of the transient response are not particularly important since the gain control loop response is also limited by the low sampling frequency.

4. Remarks on circuit design and adjustments

The overall transimpedance of the current amplifier is 2.5×10^9 Ω at its maximum value. This corresponds to full scale sensitivity of 4×10^{-10} A with an output amplitude of 1 V. The gain of a single current amplifier stage (assuming a very high impedance source) for the low frequency (series) voltage noise and drifts is unity. Consequently, the transimpedance of the first stage should be as high as permitted by the signal rise time requirements, so that the subsequent voltage gain can be low. Operational amplifiers with field-effect transistors are used to minimize the input current. Philbrick 1011/02 amplifiers with low voltage drifts are used in the first and second stage. The gain control is arranged so that as the signal increases by about 2 decades from the low end of the range the gain decreases in the second stage, while it remains constant in the first stage. This means that the signal-to-noise (drift) ratio improves linearly with the signal. This is achieved by subtracting an adjustable constant voltage from the lamp LCR1. (If the feedback resistor in the first stage were to decrease, thermal noise current from that resistor would increase, and signal-to-noise ratio would be lower for small signals.) The adjustment is performed at a signal of 10^{-9} A by raising the voltage at the base of T_{10} (resistor R_1) until there is a perceptible increase in noise (generated by the feedback LCR in the first stage). For most LCR units base-emitter voltage on T_{10} was sufficient and a separate adjustment was not necessary. At the high end of the signal dynamic range

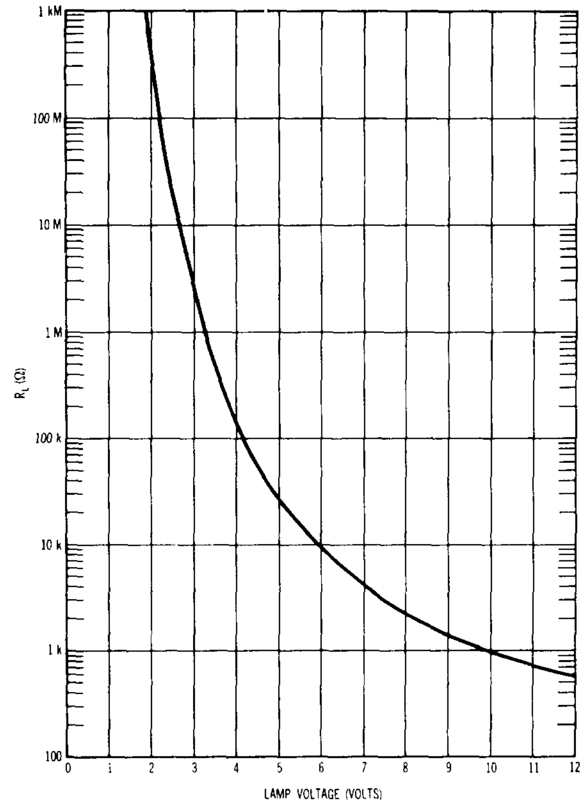

Fig. 2. Resistance as a function of lamp voltage for a light controlled resistor HP 5082–4510.

LCR1 should be at its minimum value so that saturation of the first stage is prevented. This is achieved by reducing the lamp voltage for LCR2 at high lamp currents. An adjustment is made using resistor R_2 in series with the lamp. At a signal level of 10^{-3} A, the output of the first stage is mimimized, while the output of the whole amplifier is maintained at 1 V. A smaller signal dynamic range results with preselected values for R_1 and R_2 and no adjustments.

The resistor-diode network (resistors 10 kΩ and 240 Ω, diodes D_1 and D_2) in the input stage serves to maintain aperiodic response for various signal amplitudes and input capacitances, while keeping the change of potential of the scanner low.

The channel thermal noise of the field-effect transistor in the input stage becomes dominant at low signal levels with the feedback resistor values chosen. This noise source can be represented as an equivalent series voltage generator in the input of the amplifier. Its contribution is determined (as is the voltage gain) by the ratio of the feedback network impedance to the input network impedance. Since the input network is mainly the capacitance of input connections, long cables between the scanner and the amplifier would result in increased noise.

The noise spectrum of current amplifiers increases with frequency beyond the cutoff frequency of the feedback network as the gain for signal current decreases. The noise spectrum eventually decreases as the unity-gain frequency of the operational amplifier is approached. Most of the noise power for wide-band current amplifiers is outside the frequency region in which signal gain is constant. Therefore, a large improvement in signal-to-noise ratio can be achieved without a significant increase in signal response time by placing a low-pass filter after the current amplifier. A second order filter is used here, as shown in fig. 1.

The input stage should be shielded to prevent feedback from the output stages. Due to a high transimpedance and a short response time, a small degree of coupling would increase the response time (for negative feedback) or cause oscillations (for positive feedback). For example, negative feedback via a capacitance of 4×10^{-2}pF with transresistance of 2.5×10^{9} Ω, at maximum gain, would increase the response time constant to 100 μsec.

5. Results

A current pulse generator with rise time of less than about 10 μsec is required to test the automatic gain control and the signal response. A simple generator is shown in fig. 3. Current switching is used at higher current levels, and voltage switching through an IRC HF resistor (5 MΩ) is used at lower current levels.

The gain control response is illustrated in fig. 4 for various input amplitude changes. The response time for a step decrease in signal by six decades is about 2 sec, and it is limited by the slew rate of the control

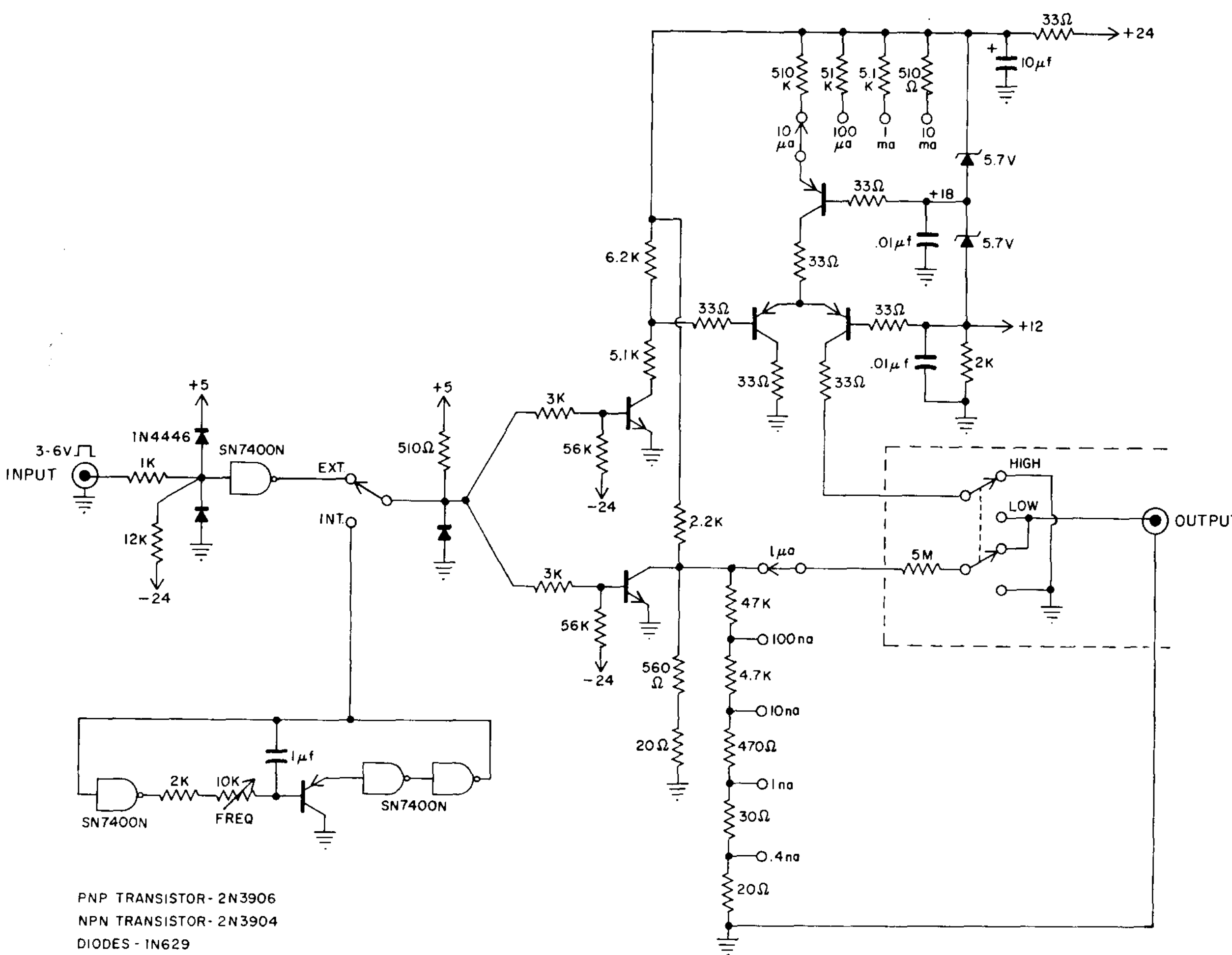

Fig. 3. Current pulse generator.

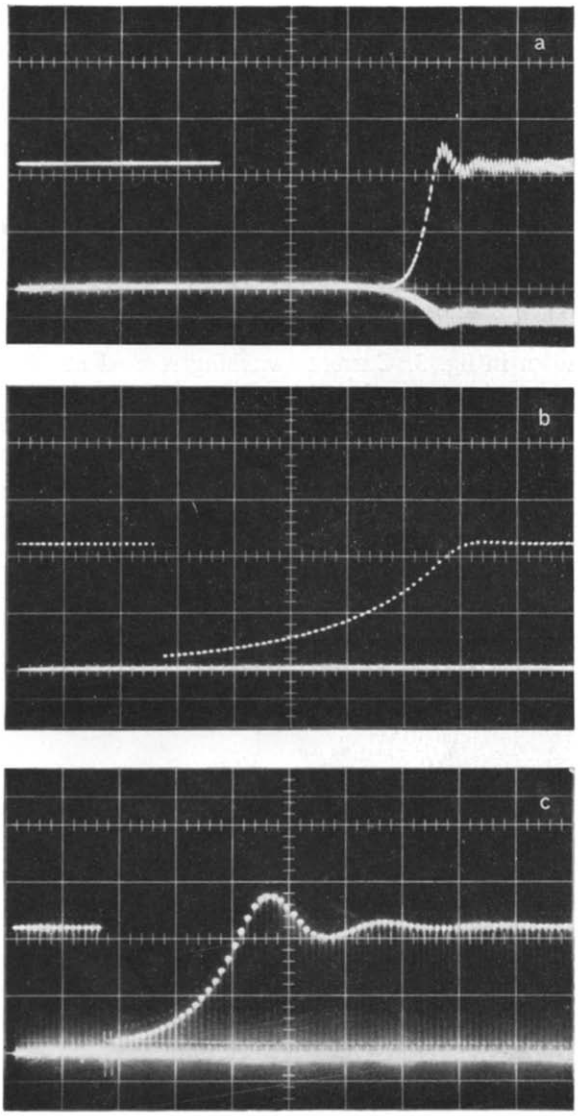

Fig. 4. Gain control loop response to step amplitude changes. (a) From 1 mA to 1 nA; horiz. scale 0.5 sec/div. (b) From 1 mA to 100 μA; horiz. scale 0.1 sec/div. (c) From 100 nA to 10 nA; horiz. scale 0.1 sec/div.

circuit, fig. 4(a). The response at higher signal levels, fig. 4(b), is somewhat slower than at lower levels, fig. 4(c). This is because of the smaller rate of change of resistance with lamp voltage (fig. 2) at lower resistance

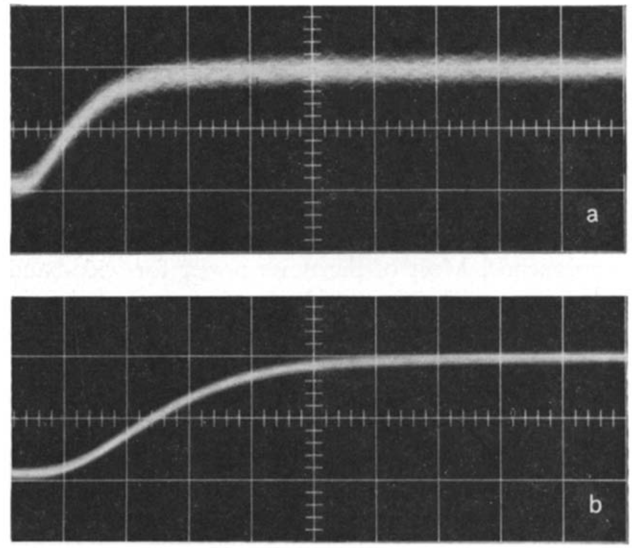

Fig. 5. Current amplifier response to individual pulses at maximum gain. (a) Step response with short time constant. (b) Step response with long time constant. Horiz. scale 20 μsec/div; vert. scale 150 pA/div.

values, which corresponds to a lower loop gain. Fig. 4(c) shows the response at a signal level at which the periodic response is most pronounced, due to higher loop gain and a longer response time of LCR's. Fig. 4(a) shows the effect of input amplifier voltage offet on the baseline as the gain increases to its maximum value.

Fig. 5 shows step response of the current amplifier and the effect of filtering on noise and on the signal. Fig. 5(a) represents the shortest response time for this amplifier. Fig. 5(b) is with the low-pass filter chosen so as to make the response time of the amplifier equal to that of the scanner. The noise in fig. 5(b) is about 40 pA peak-to-peak.

The authors wish to thank H. E. Wegner for stimulating and supporting this work.

Reference

[1]) H. E. Wegner and I. L. Feigenbaum, IEEE Trans. Nucl. Sci. NS-**14**, no. 3 (1967) 1099.