CHAPTER 5

SUPPLY AND OPERATING ADVICES

Correct use of a photomultiplier calls for observance of certain rules and circuit techniques. Those described here are indicative of present-day practice and sufficient to serve as a working guide.

5.1 Applying voltage

5.1.1 Polarity

There are two ways of applying the high voltage to a photomultiplier:

– *positive polarity*, with the cathode earthed and the anode at high positive potential (Fig.5.1)

– *negative polarity*, with the anode earthed and the cathode at high negative potential (Fig.5.2).

The choice depends on the application.

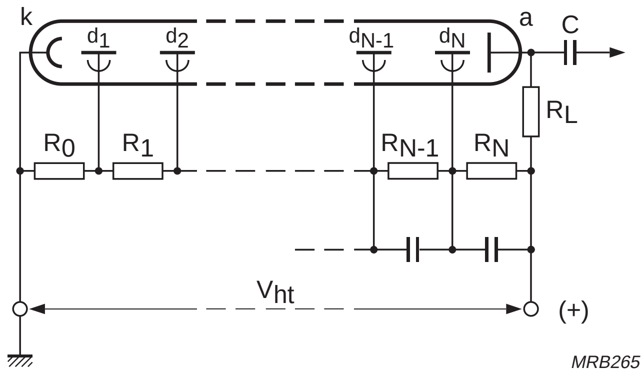

Fig.5.1 Positive-polarity voltage supply

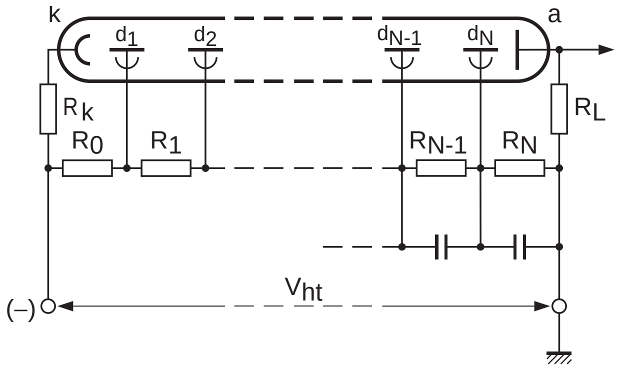

Fig.5.2 Negative-polarity voltage supply

For pulse counting, positive polarity is usually preferred; the capacitor C isolates the measuring circuits from the high voltage. Positive polarity is convenient because it does not necessitate insulating the tube from its surroundings. Magnetic screens and the coatings of scintillators should, however, be kept at photocathode (earth) potential.

For applications involving detection of continuous flux or very brief pulses, where use of a coupling capacitor would be unpractical, negative polarity has to be used. This necessitates special precautions to minimize its effect on dark current and to guard the tube against the potentially destructive effect of voltage gradients across the glass.

With tubes whose dark current is low (those with bialkali cathodes, for example), the time required for the dark current to settle after switch-on may delay measurements for half an hour or more. With all tubes, insulation defects or capacitive currents between the cathode and nearby earthed parts may make the dark current high and unstable. There are two ways to guard against this:

- mount the tube free-standing, supported only by its socket and out of contact with surrounding parts. At normal atmospheric pressure a few millimetres of air space is enough to afford adequate insulation.
- If the tube must be in contact with surrounding parts (e.g. for cooling), insulate it from them (Fig.5.4). Coat the envelope with a conductive layer (e.g. metallic paint) to neutralize any potential gradients on the glass that might give rise to leakage currents, and connect the conductive coating to cathode potential via a safety resistor of, say, 10 MΩ. Enclose the conductive coating in an insulating coating with a high insulation resistance. If good heat transfer through the coating is required a material with high thermal conductivity should be used.

If the environmental humidity is very low and the resistance of the insulation (available) is high enough, the conductive coating can be dispensed with. This requires an insulation resistance of at least $10^{15}\Omega$ (e.g. a few tenths of a millimetre of Teflon or a few millimetres of Silastene, curve 4 in Fig.5.3).

Figure 5.3 illustrates the dark-current behaviour of a magnetically shielded tube (Fig.5.4) under various conditions of insulation. Polarity was negative but the voltage was increased gradually to its nominal value to limit the initial dark-current transient. It is clear that the better the insulation, the better the dark current stability.

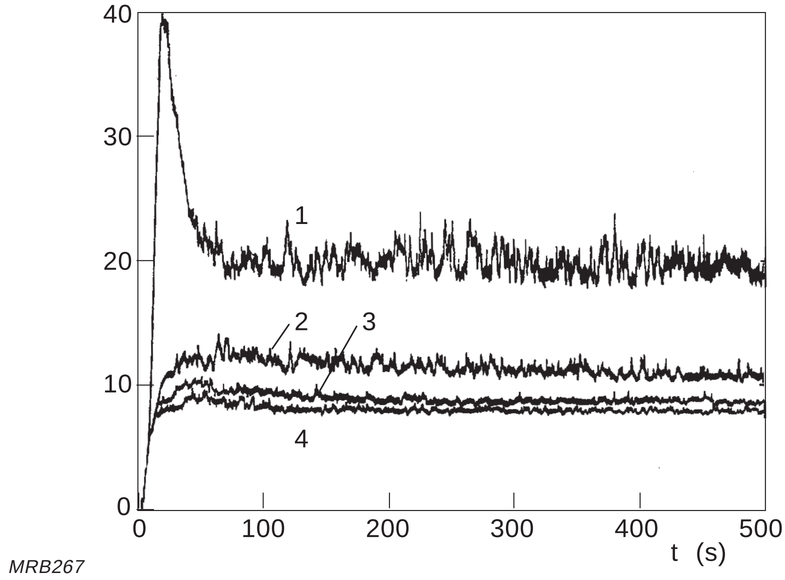

Fig.5.3 Stabilization of anode dark current I_{ao} (arbitrary scale) following gradual application of voltage in negative polarity, showing the effect of different ways of insulating the tube from its magnetic shield.

1: tube without conductive coating, wrapped with insulating tape

2: tube with conductive coating connected to cathode potential, wrapped with insulating tape

3: tube with conductive coating connected to cathode potential, insulated with 2 mm of Silastene

4: tube insulated with 4 mm of Silastene (with or without conductive coating)

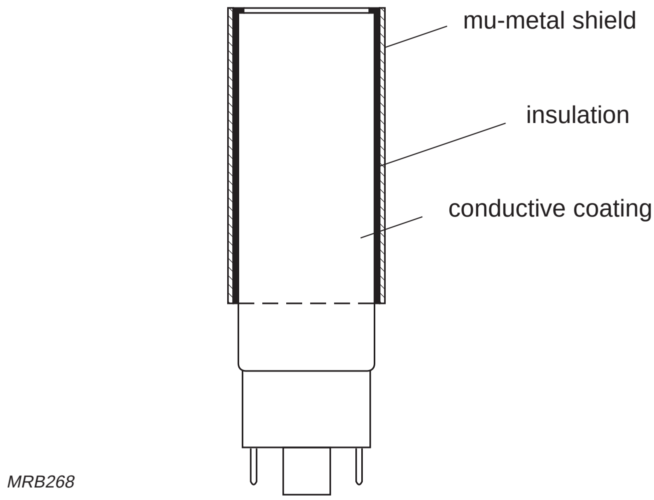

Fig.5.4 Magnetic shielding, insulation, and conductive coating

Figure 5.5 shows the dark current behaviour of a negatively-connected tube with differing provisions for heat transfer to the magnetic shield. Heat transfer coefficients in both cases were approximately equal.

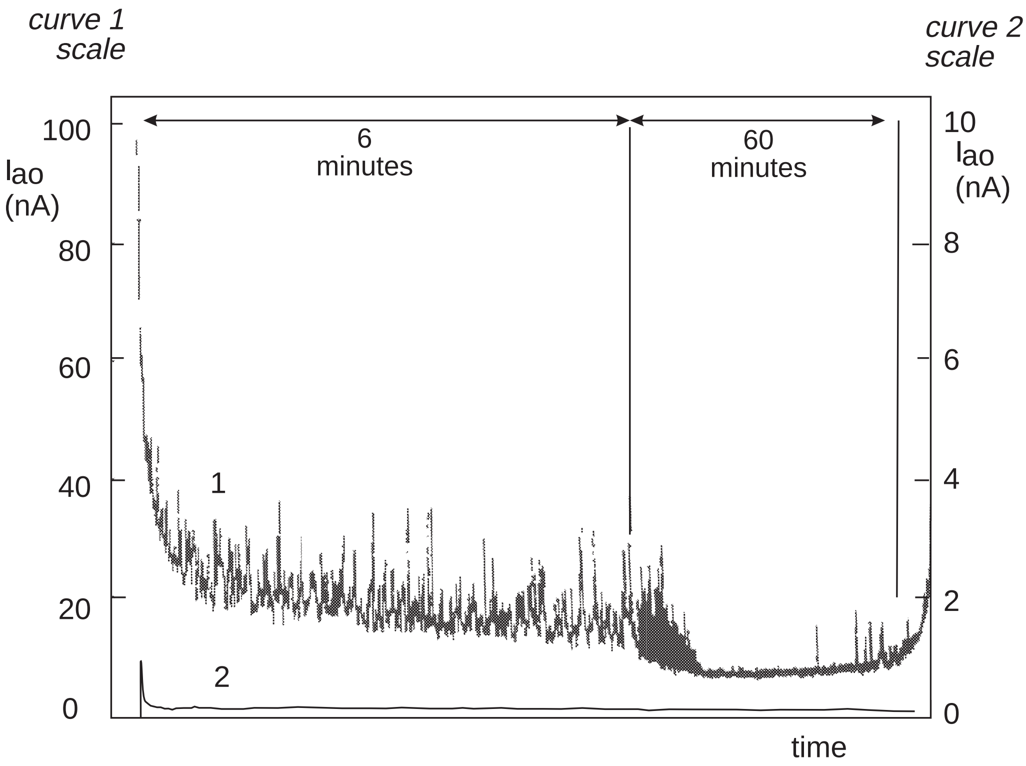

Fig.5.5 Dark-current stabilization of a negative-polarity connected tube in thermal contact with its magnetic shield, showing the effect of different ways of achieving thermal contact and electrical insulation.

1: tube without conductive coating, thermal contact between tube and shield via crumpled aluminium foil

2: tube with conductive coating connected to cathode potential, thermal contact between tube and shield via 3 mm of thermally conductive insulation

5.1.2 Rate of voltage application

Positive polarity If the high voltage is applied abruptly to a photomultiplier connected in positive polarity (Fig.5.6), an initial pulse of amplitude $V_{ht}R_E/(R_L + R_E)$ is coupled through the capacitor to the preamplifier input and causes damage. To avoid this it is advisable to use a circuit decoupling network (shown dotted in Fig.5.6) with a time constant R_tC_t of at least one second. An alternative is to shunt a protection diode across the resistor R_E to eliminate all positive-going pulses at the amplifier input.

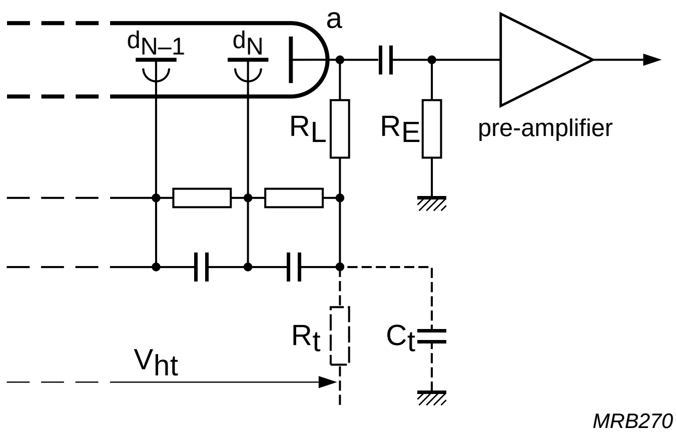

Fig.5.6 Network for decoupling the high-voltage switch-on transient from the amplifier input

Negative polarity If voltage is applied abruptly to a tube connected in negative polarity, the amplitude of the initial dark-current transient (Fig.5.5, trace 1) may be high enough to damage sensitive measuring apparatus. Applying the voltage gradually reduces the transient or may even eliminate it. The RC time constant should be a few seconds.

5.2 Voltage dividers

The choice of voltage divider, including the type of voltage distribution, depends on:
- the application of the tube; e.g. continuous, pulse, or high-current operation, etc.
- the performance required; e.g. gain, linearity, timing, stability, etc.

5.2.1 Types of voltage distribution

Recommended voltage distributions are given in the data sheets for each tube. There are three main types, designated A, B and C.

Type A (*equal steps*, Fig.5.7(a)) Voltages between all the iterative dynodes are equal. This distribution gives maximum gain for a given supply voltage and is particularly suitable for photometry and nuclear spectrometry applications.

Type B (*progressive*, Fig.5.7(b)) Interdynode voltages increase progressively in the anode direction, becoming 8 to 10 times as high in the last stages as in the first. This distribution makes it possible to obtain anode pulses of several hundred milliamperes peak with good linearity. Gain, however, is much lower than with type A distribution for the same total high voltage.

Type C (*intermediate*, Fig.5.7(c)) Interdynode voltages increase only in the last stages. Time characteristics are optimized; gain and pulse linearity are also satisfactory. So, type C distribution is particularly suitable for fast response photomultipliers.

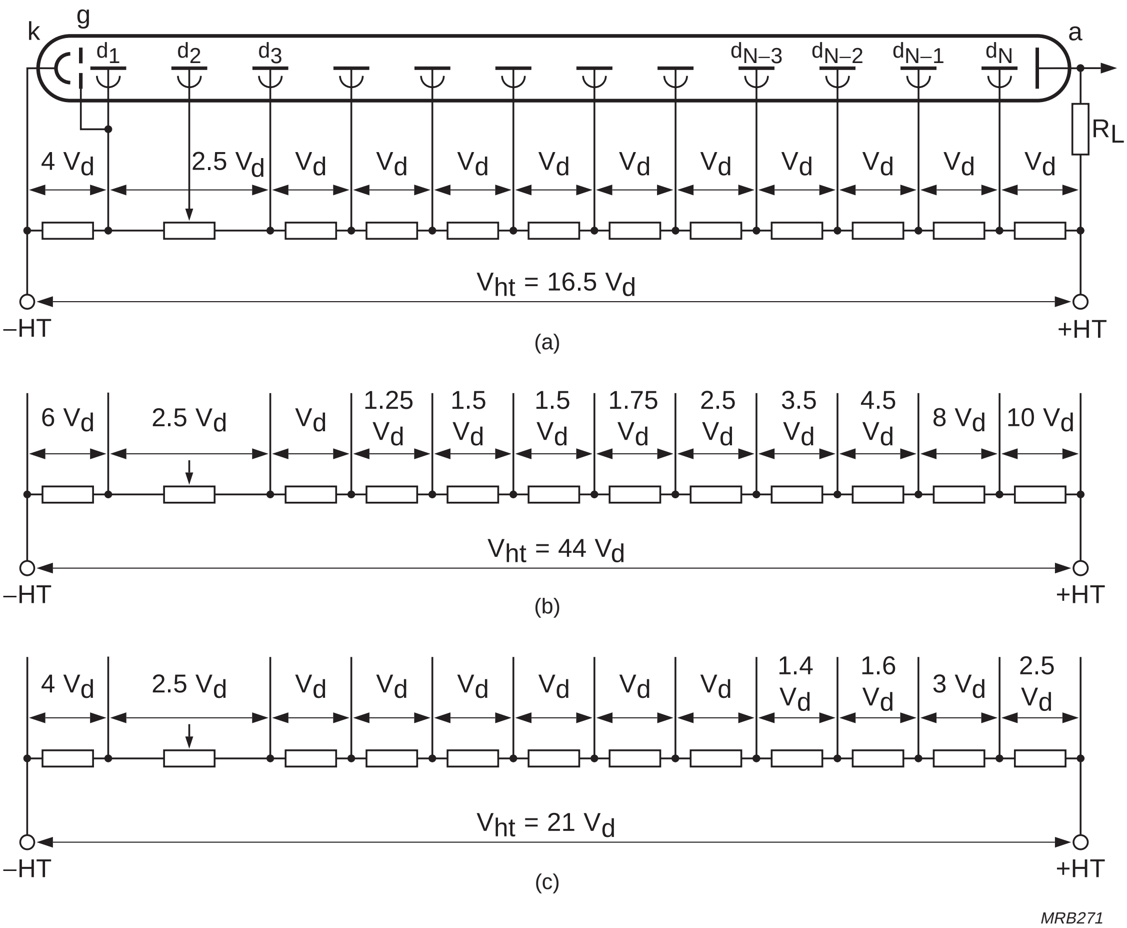

Fig.5.7 Voltage dividers: (a) type A equal step, (b) type B progressive, (c) type C intermediate

Table 5.1 compares the gain and pulse linearity of a fast response photomultiplier for each type of distribution.

Table 5.1
Gain and pulse linearity of a fast response photomultiplier, with three types of voltage distribution

Type of voltage distribution	Gain V_{ht} = 2500 V	Linear within 2% for current pulse amplitudes up to
A	1.2×10^8	40 mA
B	0.7×10^6	250 mA
C	2×10^7	100 mA

Other types of distribution are sometimes offered. The type A1 distribution, for example, has a higher cathode to first-dynode voltage than the type A to ensure good collection efficiency even when the tube operates at low voltage. The gain, though, is lower than with type A distribution.

5.2.2 Resistive dividers

Design of a resistive voltage divider depends on the supply voltage, the voltage distribution, and the anticipated mean anode current I_a. To ensure that voltage variations due to anode current variations are negligible, the nominal divider current I_p must be much larger than I_a; a good rule is

$$\frac{I_p}{I_a} \geq 100 \tag{5.1}$$

Consider a type C voltage distribution with V_{ht} = 2500 V, (Fig.5.7(c)) an interdynode voltage increment V_d, and a division ratio such that there are precisely 21 such increments. Let the maximum anticipated mean anode current I_a be 10 µA. The divider current I_p must then be at least 1 mA, which means a total divider resistance of 2.5 MΩ and an incremental resistor value of $2.5 \times 10^6/21 \approx 120$ kΩ.

The resistors must be properly rated for power and voltage. The latter is important because some of them must withstand several hundred volts continuously. Tolerances should be not greater than 5%.

The ratio specified by Eq.5.1 represents a minimum for maintaining good linearity. Two other considerations limit the maximum value of the ratio.

– Heat due to dissipation in the divider (I_p^2R) can cause an increase in the dark current, especially if the divider and tube are housed close together or with the tube vertical, cathode uppermost.
– Low divider current gives some protection against accidental overexposure of the cathode; as soon as anode current rises proportionately, gain drops abruptly and prevents the anode current from becoming excessive.

5.2.3 Zener-diode dividers

Zener diodes can be substituted for some of the resistors in the divider to keep certain inter-electrode voltages constant.

They can be used between cathode and first dynode to keep input collection efficiency constant regardless of supply voltage and gain setting; and between the dynodes of the first two multiplier stages to keep the gains of those stages constant (Fig.5.8(a)). This is useful where certain minimum voltages are necessary in the first stages because S/N or PHR is critical but the overall gain need not be high. It cannot be done, however, in those fast-response tubes that incorporate an accelerating electrode internally connected to one of the higher ranking dynodes; the potential of such an electrode must be in constant proportion to the voltage between cathode and first dynode.

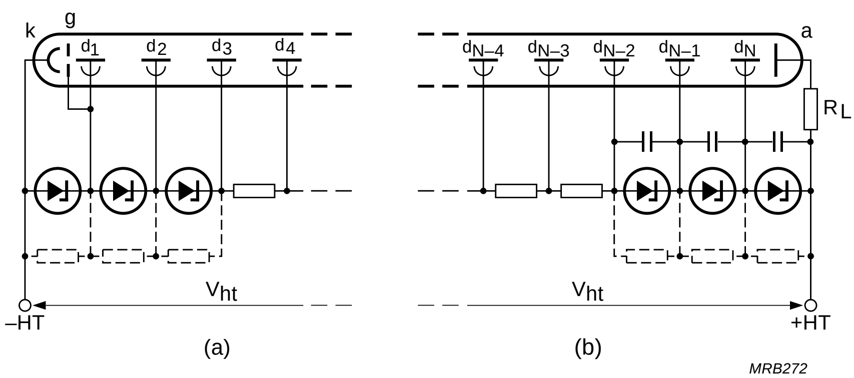

Fig.5.8 Inter-electrode voltage stabilization with zener diodes: (a) in first stages, (b) in last stages (dashed lines, protection resistors)

Another place to use zener diodes is in the last stages of the divider (Fig.5.8(b)), to stabilize the voltages there throughout a wider range of anode current variations. This also makes it possible to accommodate smaller ratios of I_p/I_a than with a purely resistive divider (§4.5.1).

In certain applications a drawback of using zener diodes is that they limit the freedom of gain adjustment. Altering the supply voltage to adjust the gain would also alter the overall voltage distribution, for it would affect the voltages across the resistor stages but not the zener-stabilized ones. As linearity is very dependent on the overall voltage distribution, a divider with zener diodes should be designed for a specific supply voltage and that voltage should be adhered to as closely as possible. Departure from it invites the risk of either overlinearity or premature saturation. The risk is considerably less if only the very last stage is zener stabilized.

Whether zener diodes are used in the higher or lower stages of the divider, they should be shunted by resistors to protect those stages from receiving the full supply voltage in the event of a diode going open-circuit. The values of the resistors should be 2 to 3 times what they would be in a purely resistive divider.

The temperature coefficients of the zener diodes is an important consideration. Variation of zener voltage with temperature can cause variation in gain.

Note, though zener-diode dividers offer some advantages in special situations, these advantages are sometimes over-stressed and often the best solution is a specially tailored resistive divider.

More elaborate voltage dividers including active components such as transistors are also used to cope with high mean anode currents (details can be found in the Photonis Photomultiplier catalogue).

5.2.4 Multiple power supplies

If negative-polarity connection is not objectionable, using more than one supply can ease some of the constraints on a resistive divider mentioned earlier. It enables good linearity to be obtained over a wide anode current range without excessive dissipation in the resistors. The supplies may be either in series (Fig.5.9(a)) or parallel (Fig.5.9(b)). In either case, the one connected to the cathode is a high-voltage low-current (<1 mA) supply. The other, feeding the anode and last three or four dynodes, is a low-voltage high-current supply. The terminal common to both is usually earthed and, in fact, has to be if the supplies themselves are not isolated from earth.

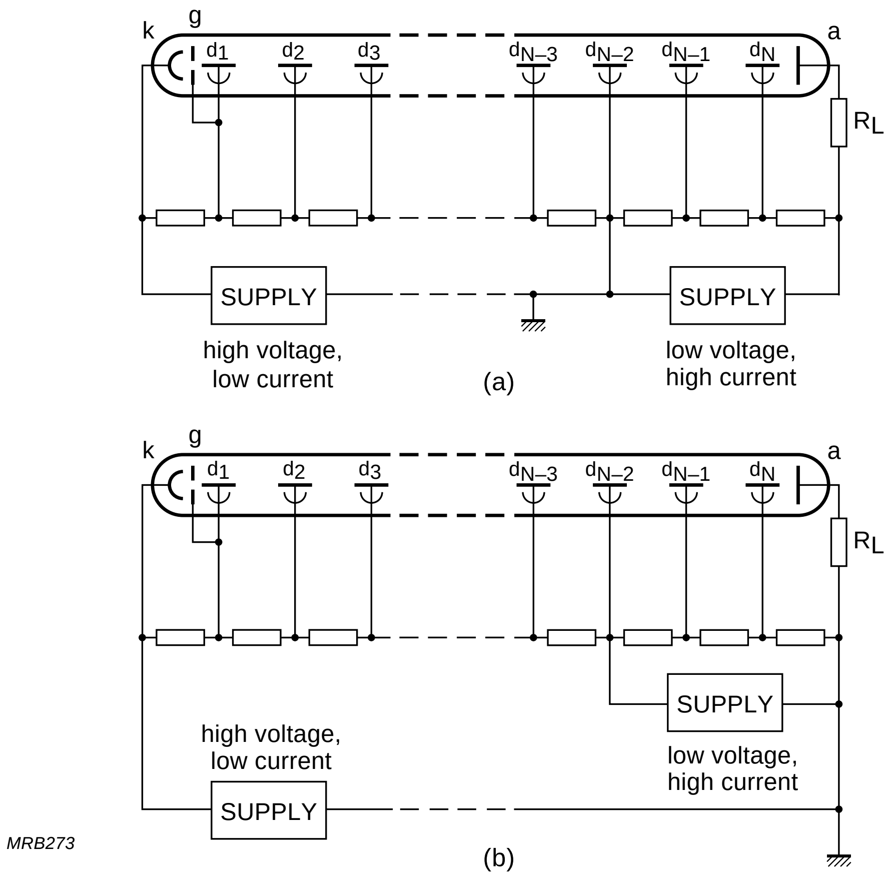

Fig.5.9 Dual voltage supplies: (a) series, (b) parallel

A parallel supply as in Fig.5.9(b) can also be used to provide additional current through zener diodes fitted to the last stages of a voltage divider. In this way the operating point of those stages can be set independently of the current through the resistive part of the divider, which can be kept low. In high-energy physics each of the last four or five stages is often provided with its own parallel supply, making it possible for the photomultiplier to accept high count rates without loss of linearity.

5.2.5 Decoupling

Provided the dynodes are adequately decoupled, instantaneous values of current in pulse operation may greatly exceed the mean value of the divider current. In observing the rule of Eq.5.1, the value taken for I_a should be the mean anode current based on the anticipated pulse amplitude and duty factor. Decoupling may be either parallel (Fig.5.10(a)) or series (Fig.5.10(b)); the former requires high-voltage capacitors which are larger and more expensive.

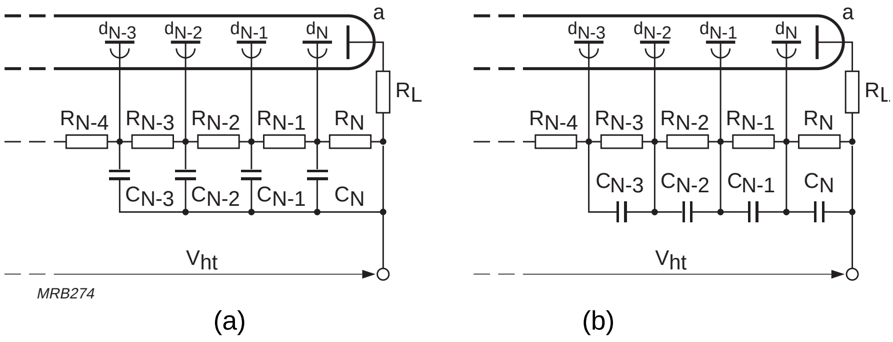

Fig.5.10 Dynode decoupling: (a) parallel, (b) series.

The decoupling capacitors act as reservoirs to restore the charge transferred by pulses passing through the tube. Let Δq_N be the pulse charge supplied by the last dynode, and ΔV_N the maximum voltage change that can be tolerated on that dynode. The required value of capacitor C_N is then given by

$$C_N \geq \frac{\Delta q_N}{\Delta V_N} \tag{5.2}$$

The charge added per stage increases in direct proportion to the gain; therefore, the required value of C_{N-1} is

$$C_{N-1} = \frac{C_N}{g_N}$$

where g_N is the gain due to dynode d_N. If all stages have equal gain, the general expression for the required capacitance in the i-th stage is simply

$$C_i = \frac{C_N}{g^{N-i}} \tag{5.3}$$

Working backwards from the last stage to the first, this rule should be applied until the value it gives for C_i is comparable to the stray capacitance of the stage (usually about 20 pF).

The foregoing is based on the assumption that the voltage changes ΔV_i on the dynodes occur independently of each other, but that is not always so. Consider the case where the pulse duration is large compared with the overall transit time, and the dynodes are series decoupled. The voltage changes can then be fed back down the decoupling network in such a way that

$$\Delta V_i = \sum_{j=i}^{N} \Delta V_j$$

For adequate decoupling the capacitance in each stage except the last must then be increased by a safety factor about equal to the stage gain. Equation 5.3 then becomes

$$C_i' = g \frac{C_N}{g^{N-i}} = \frac{C_N}{g^{N-i-1}} \tag{5.4}$$

Example A photomultiplier adjusted for an overall gain of 3×10^5 detects NaI(Tl) scintillation pulses due to γ-radiation from a ^{60}Co source. The full width at half maximum of the resulting anode pulses is about 0.3 µs, and their amplitude about 1 mA; the pulse charge is therefore

$$\Delta q_N \approx I_a t_w = 0.3 \times 10^{-9} \text{ C} \tag{5.5}$$

If the maximum tolerable change in last-dynode voltage is $\Delta V_N = 1$ V (i.e. about 1% of the interdynode voltage), the minimum decoupling capacitance required at that dynode is

$$C_N \geq \frac{\Delta q_N}{\Delta V_N} = 0.3 \text{ nF}$$

The decoupling capacitors recharge with a time constant determined by the divider resistances. It is not necessary for each capacitor to recharge fully before every pulse but only for it to recover the charge lost during the preceding pulse. What is important is that the amplitude of the capacitor voltage variation in each stage should not exceed the tolerable value of ΔV_i in that stage. This condition is satisfied when the rule given in Eq.5.1 is observed; that is, when

$$I_a \leq \frac{I_p}{100} = \frac{V_{ht}}{100 \sum_{i=1}^{N} R_i} \tag{5.6}$$

where V_{ht} is supply voltage and R_i the voltage divider resistance at each stage. Together, Eq.5.2 and (5.3) define the minimum capacitance needed to decouple each stage, and Eq.5.6 the minimum divider current to restore the capacitor charges. Apart from considerations of bulk and cost, there is no strict upper limit to the decoupling capacitances.

The examples that follow illustrate the calculation of decoupling capacitance. For the sake of comparison the following conditions are common to all three:

– peak anode pulse amplitude, $\hat{I}_a$	1 mA
– full width at half maximum, t_w	0.3 μs
– charge per pulse, $\Delta q_N = \hat{I}_a t_w$	0.3×10^{-9} C
– number of pulses per second, n	10^4
– multiplier gain per stage, g	3
– tolerable voltage variation at last dynode, ΔV_N	1 V

The mean anode current $I_a = n\Delta q_N$ and the conditions to be satisfied are those of Eqs 5.1 and 5.2.

Fixed-frequency pulses From Eq.5.1

$$I_p \geq 100 I_a = 100 n \Delta q_N = 100 \times 10^4 \times 0.3 \times 10^{-9} = 0.3 \text{ mA}$$

From Eq.5.2

$$C_N = \frac{\Delta q_N}{\Delta V_N} = 0.3 \times 10^{-9} = 0.3 \text{ nF}$$

whence, from Eq.5.4

$$C_{N-1} = 0.3 \text{ nF} \quad C_{N-2} = 0.1 \text{ nF} \quad C_{N-3} = 33 \text{ pF}$$

and $C_{N-4} = 11$ pF, which is comparable with the stray capacitance and can be neglected, as can all lower stages. For C_N to C_{N-3} four 1 nF capacitors can be used for simplicity.

Random-frequency pulses. Let $\bar{n} = 10^4$/s be the mean about which the frequency fluctuates, and assume that the maximum instantaneous frequency is $3\bar{n}$. Two methods of approach are possible.

1. As the charge per pulse is the same as before, the same decoupling capacitances will suffice. If the calculated values (0.3 nF, 0.3 nF, 0.1 nF, 33 pF) are chosen, however, the divider current will have to be tripled to allow for the instantaneous excursions of pulse repetition frequency to $3\bar{n}$
2. The mean anode current is the same as before, so it should be practical to work with the same divider current. It is, but the capacitors will have to supply three times as much charge during the instantaneous frequency excursions. This leads, as before, to the choice of 1 nF capacitors and is preferable to tripling the divider current; in this case five are required, one each for C_N to C_{N-3} and another for C_{N-4} (calculated value 0.33 nF).

Pulse bursts Consider bursts of 10 ms containing 10^4 pulses and recurring once a second. For simplicity, each burst can be regarded as a single 10 ms pulse conveying a total charge

$$\Delta q_N' = 10^4 \, \Delta q_N = 3 \times 10^{-6} \text{ C}$$

and the anode current is, as before, 3 μA. The required divider current is therefore 0.3 mA. However, the decoupling capacitance C_N becomes

$$C_N = \frac{\Delta q'_N}{\Delta V_N} = 3 \ \mu\text{F}$$

Applying Eq(5.4), this leads to a requirement for capacitors in twelve stages before the indicated decoupling capacitance becomes comparable with the stray capacitance of the stage. Fortunately, though, it is not necessary to decouple stages in which the dynode current is less than 1% of the anode current. In the present example (gain per stage, $g = 3$), this is the case for all stages lower than $N - 4$. Using standard values, the actual capacitor requirement is therefore

$$C_N = 3.3 \ \mu\text{F} \quad C_{N-1} = 3.3 \ \mu\text{F} \quad C_{N-2} = 1 \ \mu\text{F} \quad C_{N-3} = 330 \text{ nF}$$
$$C_{N-4} = 100 \text{ nF}$$

Even so, some of these are inconveniently large for capacitors that must have a high working voltage.

One practical alternative is to use a higher current (e.g. $I_p \approx 3$ mA) in a divider with zener diodes in the last four stages. Five 10 nF capacitors would then suffice.

Another alternative, which makes it possible to accept even large pulse burst, is to use a divider with separate supplies for the last four stages.

The networks of Fig.5.10 are not the only practical ones. Figure 5.11 shows a 2-and-2 series arrangement that is often used when many dynodes have to be decoupled and capacitor lead lengths must be kept to an absolute minimum (alternate dynode pinning).

5.2.6 *Wiring precautions*

When the tube is connected in positive polarity, observe the usual wiring rules for high-voltage/low-current; take especial care over insulation of the output stage. Keep decoupling capacitor leads short to minimize stray inductance. (Voltage divider resistor leads are not so critical; a printed wiring board may be used).

When the tube is connected in negative polarity take especial care over insulation of the cathode connection and, if the best possible time characteristics are required, observe the following additional rules:

- use the 2-and-2 decoupling network of Fig.5.11.
- connect the last two decoupling capacitors to the braided sheath of the coaxial cable and earth the sheath there as well as the output end. If other coaxial cables are used (e.g. for double anode output, dynode output), earth them at the same point. Terminate coaxial cables in their characteristic impedance.
- some fast response tubes incorporate integral damping resistors, in the base (see Fig.5.12). If the type used does not, connect such resistors externally between the last two dynodes and their decoupling capacitors. Use non-inductive 50 Ω resistors.

If the socket used has pin contacts, it can be mounted direct to a printed wiring board. This considerably simplifies wiring of the voltage divider.

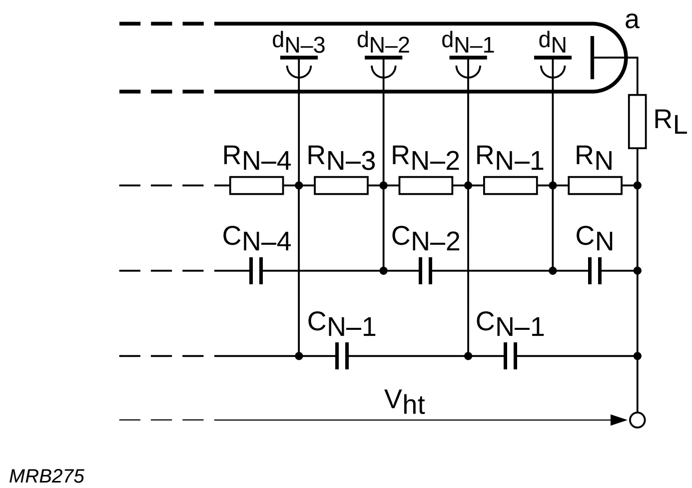

Fig.5.11 Series decoupling of alternating dynode pairs

5.3 Output connections

5.3.1 Anode resistor

Whether the tube is connected in positive or negative polarity, the anode potential must be fixed.

If the tube is connected in negative polarity and direct coupled to the measuring apparatus (Fig.5.13(a)), the anode potential is clamped by the internal resistance of the apparatus. However, if the output is disconnected even briefly while the high voltage is still applied, the anode will acquire a negative charge which may damage the apparatus when connection is restored. Therefore it is advisable to fit a protection resistor (dotted in Fig.5.13(a)) between the anode and earth. As it is shunted across the high internal resistance of the photomultiplier, the protection resistance must also be high. The value chosen depends mainly on the load circuit and is typically ≥10 kΩ.

If the tube is connected in positive polarity and capacitively coupled to the measuring circuitry (Fig.5.13(b)), a resistor between the anode and the positive terminal of the high voltage supply is essential; this resistor can also constitute the anode load. Once again, the resistance must be reasonably high and depends mainly on the input impedance of the measuring circuitry.

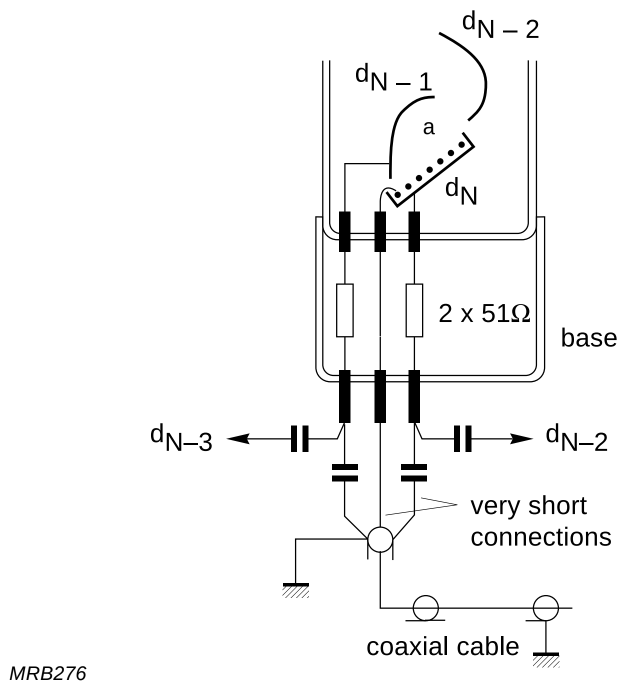

Fig.5.12 Connection of damping resistors and decoupling capacitors to the last two dynodes

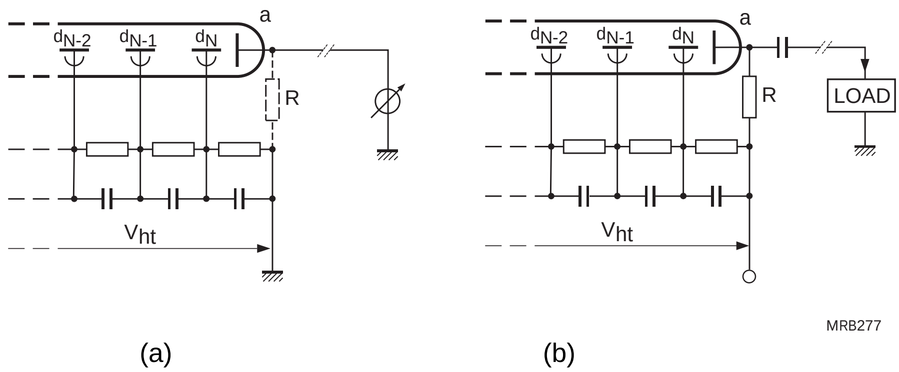

Fig.5.13 Fixing the anode potential with the tube connected in (a) negative, (b) positive polarity

5.3.2 Output cable high-voltage connection

Figure 5.14 shows two ways of using the output cable for connection of the high voltage supply. In Fig.5.14(a) resistors R_1 and R_2, together with the capacitance of the cable, form the anode load; if the signal is to be integrated, this is acceptable. If the cable is long, the alternative shown in Fig.5.14(b) can be used, with the step-down transformer matched to the characteristic impedance of the cable.

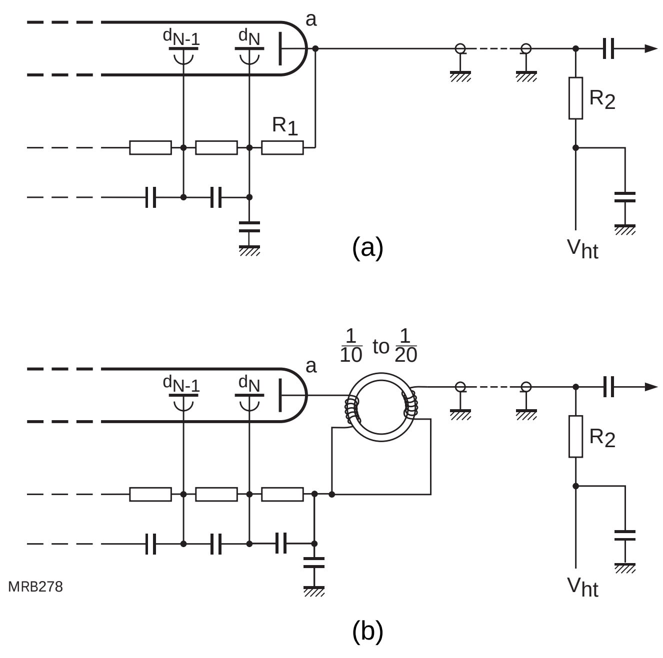

Fig.5.14 Voltage supply via the output cable

5.3.3 Signal taken from dynode

There are two cases in which a signal may have to be taken from a dynode:

- when the measurement to be made requires a signal synchronous with the anode signal
- when it is desirable to limit gain by using fewer stages.

Procedures applicable to the two cases are different.

Synchronous dynode signal. The requirement for a synchronous signal from a dynode usually arises in connection with the detection of very short pulses. The signal may be required either for synchronizing an instrument or for supplying additional charge or amplitude information. The problem is to obtain the required signal without disturbing the anode signal. Figure 5.15 shows two solutions. To obtain an amplitude comparable with that at the anode, the signal is usually taken from the last dynode but need not be if amplitude is not a governing consideration.

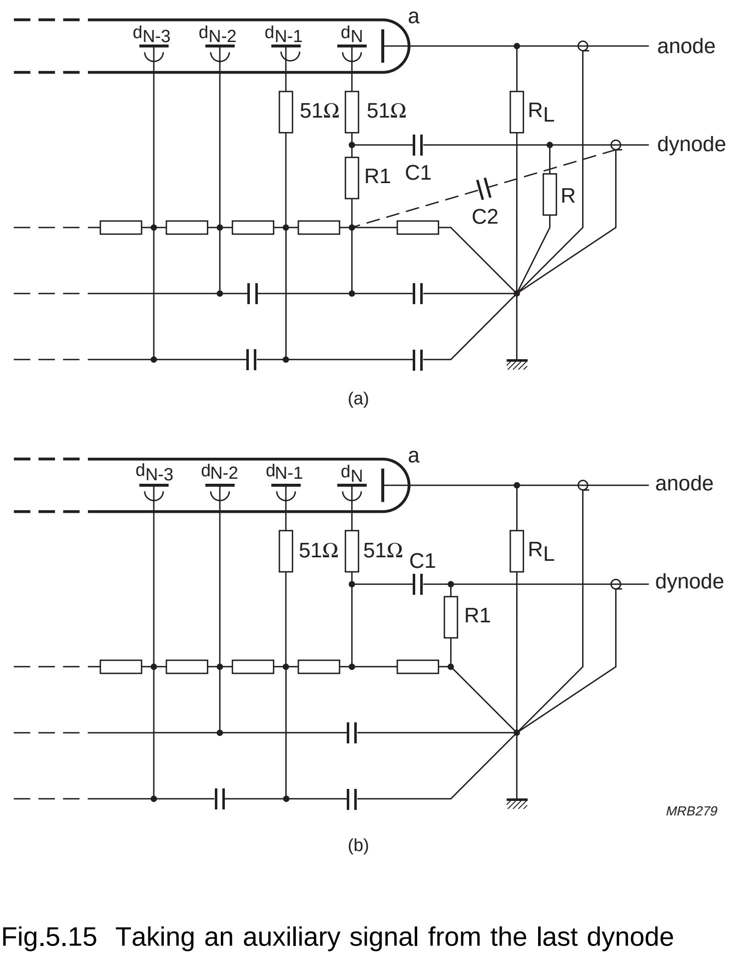

Fig.5.15 Taking an auxiliary signal from the last dynode

The coupling capacitance C_1 must be such that the product C_1Z_c (where Z_c is the characteristic impedance of the coaxial cable) is much greater than the expected pulse duration, otherwise the pulses will be differentiated. Using a single earth point, as indicated, minimizes inductive effects. If, in spite of that, oscillation does occur, connecting a second capacitor, $C_2 = C_1$, as shown in Fig.5.15(a), will minimize it.

Resistance R_1 must be large compared with Z_c but not too large. If the voltage across it exceeds a few volts there is a risk of disturbing the anode signal. For $Z_c = 50\ \Omega$, the usual value of R_1 is 200 Ω to 300 Ω. When the dynode output is not in use, terminate the cable with its characteristic impedance.

Pulses taken from a dynode are positive-going. If negative-going pulses are required, as is standard in fast nuclear instrumentation, a coaxial cable transformer can be used.

Dynode output. If the incident flux is large and high gain is not required, fewer multiplier stages can be used. The voltages at the still active dynodes need not be changed but the overall high voltage can be reduced. The problem is to obtain a linear output signal from dynodes that have not been designed for that. Figure 5.16 shows two solutions.

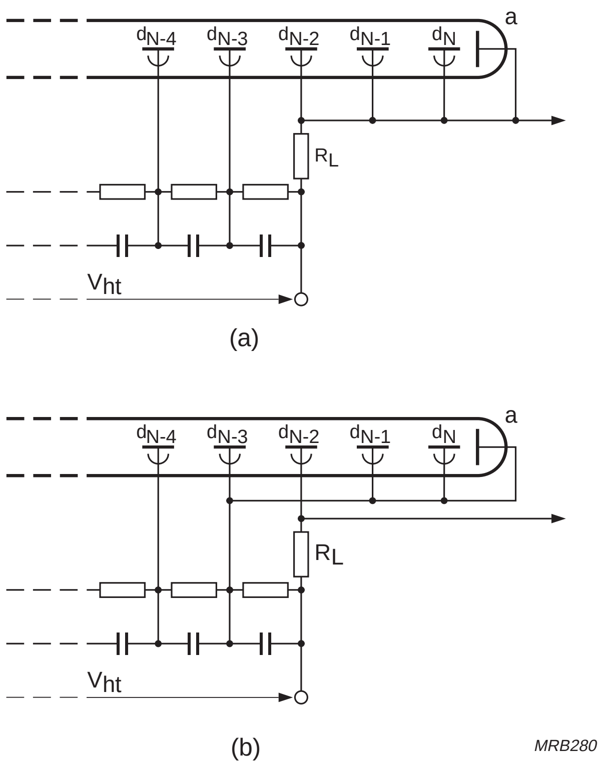

Fig.5.16 Reducing the number of active stages

In Fig.5.16(a) the inactive electrodes and the last active dynode are wired together for collecting the signal. In Fig.5.16(b) the signal is taken from one dynode, and the inactive higher-ranking dynodes and anode are connected together and taken to the potential of the next lower ranking dynode. This is an attempt to simulate the normal output geometry, in which the anode is situated between the last two dynodes. The (b) circuit is preferable to the (a) circuit for two reasons:

- it puts less stray capacitance in parallel with the load; the larger the load and the greater the bandwidth required, the more important this is.
- it gives somewhat better pulse linearity. Even so, the linearity scarcely extends to more than a few milliamperes, for dynodes, unlike the anode, are not designed for linear operation at high peak currents.

In neither circuit should unused electrodes be left floating or simply earthed; that would disturb electron trajectories elsewhere in the multiplier.

5.4 Anode load

5.4.1 Continuous operation

As long as the operating point is in the saturated current region, (Fig.5.17) a photomultiplier behaves as an almost perfect current generator; the anode current depends only on the incident flux and is completely independent of the load. Nevertheless, even in the ideal case the current through the load does lessen the voltage between the last dynode and anode. It is therefore important to check that this voltage remains high enough throughout the intended operating range to ensure complete electron collection at the anode. Figure 5.17 shows two load lines superimposed on a set of current/voltage characteristics. The extent of the linear operating range depends on the load line slope.

In practice the characteristics in the saturation region do have a slight slope (which may be either positive or negative). This accentuates the effect of the load on linearity and must be taken into account if the tube is operated in such a way as to cause excursions of more than, say, 10 V in the last-dynode-to-anode voltage.

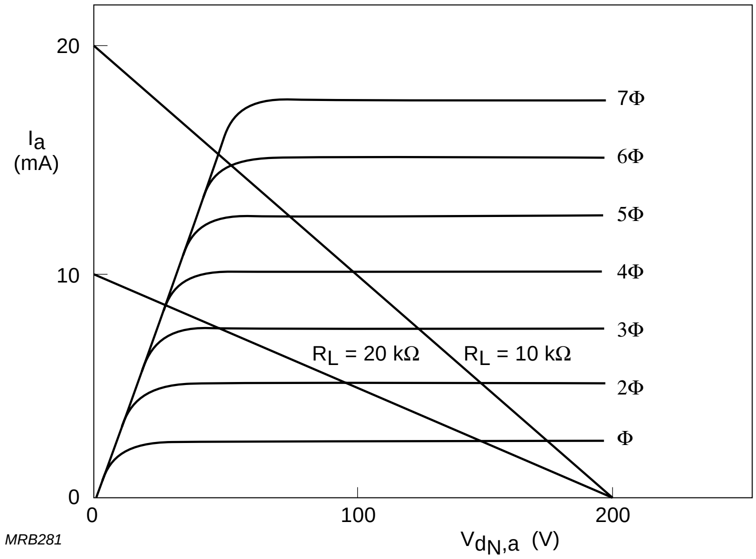

Fig.5.17 Effect of anode charge on linearity (idealized current/voltage curves)

5.4.2 Pulse operation

There are three modes of response, depending on the RC time constant of the load.

Charge mode. When the load time constant is much greater than the anode current pulse width, the pulse charge is integrated. The voltage across the load resistor is then proportional to the total charge in each pulse.

If the mean number of photons per light pulse is $\overline{n}_{p,i}$, the mean charge supplied to the anode is

$$\overline{q}_{a,i} = \overline{n}_{p,i} \, \rho e \; G \qquad (5.7)$$

where ρ is the quantum efficiency of the photocathode, e the electron charge, and G the overall gain of the tube. The amplitude of the resulting voltage pulse at the anode is

$$V_a = \frac{\overline{q}_{a,i}}{C} \qquad (5.8)$$

Thus, if maximum output is required at a specified value of RC, it is desirable for C to be as small as possible.

Current mode. When the time constant is much less than the anode current pulse width, the voltage across the load resistance varies as the anode current. A typical application of this mode is in the detection of very short pulses in time spectrometry (§6.3.2); the anode load is usually the 50 Ω impedance of a coaxial cable. The amplitude of the voltage pulse is

$$V_a = \frac{\bar{q}_{a,i} R}{t_w} = \frac{\bar{n}_{p,i} \rho e G R}{t_w} \tag{5.9}$$

where t_w is the full width at half maximum of the anode current pulse. Thus, for a tube operating under single-electron conditions with $\bar{n}_{p,i}\rho = 1$, $t_w = 2.4$ ns, $G = 3 \times 10^7$ and $R = 50\ \Omega$,

$$V_a = 100 \text{ mV}$$

That is, emission of a single photoelectron gives rise to a mean anode pulse of 100 mV into 50 Ω.

Intermediate mode. When the time constant is comparable with the anode current pulse width, the response is intermediate between charge mode and current mode. A typical instance is in nuclear spectrometry, where the scintillator light pulse usually has an exponential decay. The shape of the resulting voltage pulse depends on the ratio of the time constant RC and the scintillator pulse decay time τ. When RC >> τ, the pulse amplitude is as given by Eq.5.8; and when τ >> RC, it is as given by Eq.5.9. Between those extremes, the response has to be determined with the aid of graphs (see Fig.6.9 to 6.13).

5.5 Operating range

5.5.1 Gain and dark current characteristics

The gain and dark current characteristics plotted in the data sheets for each type are merely typical; there is some variation from tube to tube. However, each tube is accompanied by a measurement certificate stating the gain and dark current measured at a specific applied voltage and with a specific voltage distribution. To obtain the actual characteristics of that tube it is sufficient to transfer the measured points to the corresponding data sheet graph and translate the graphed characteristics so that they pass through the transferred points. Figure 5.18 shows an example. Note that in the example the measured gain is given only for type A voltage distribution. To obtain the corrected gain characteristics for type B and C distributions the corresponding lines are shifted by the same amount as the line for type A.

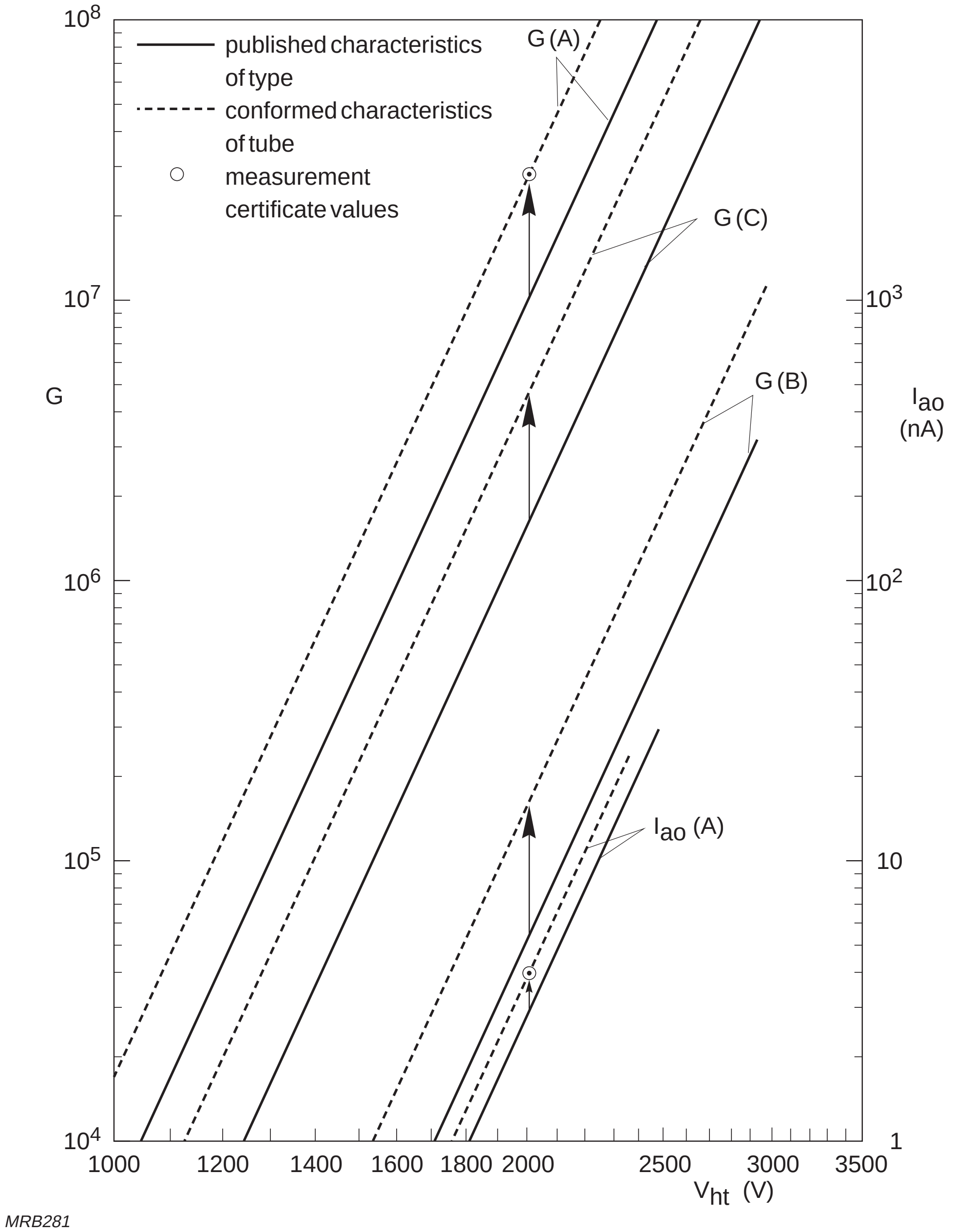

Fig.5.18 Conforming the measured gain and dark current characteristics of an individual tube to the published characteristics of the type. G(A), G(B)) and G(C) are the gain characteristics with the recommended type A, B and C voltage dividers; I_{ao}(A), the dark current with the recommended type A voltage divider

This procedure yields a good approximation; in reality the slopes of the characteristics also differ from tube to tube, but only by a few percent. The procedure is equally applicable to anode sensitivity characteristics when those, instead of gain, are given in the data sheets and measuring certificate.

Because it is a factory measurement taken after only a short stabilization period, the value of the dark current given in the measurement certificate tends to be pessimistic.

5.5.2 Choice of operating conditions: continuous operation

If the illumination level can be set at will, there is a temptation to set it high to obtain a good signal-to-noise ratio. How high it actually can be set depends on the cathode current level at which the effects of cathode resistivity become significant; for bialkali SbKCs cathodes that level is about 1 nA.

With the cathode current determined by the working conditions, the required gain and anode current depend on the characteristics of the measuring or signal processing circuits used. The minimum practical gain is that which corresponds to the minimum electrode voltages specified in the data sheets of the tube; these are based on considerations of linearity and minimum gain fluctuation. The maximum practical gain is usually determined by dark-current and signal-to-noise ratio considerations or pulse linearity limits.

For good gain stability with time, the mean anode current should be kept as low as possible.

5.5.3 Choice of operating conditions: pulse operation

In pulse operation the following factors affect the choice of operating range:
- detection efficiency
- energy and time resolution
- pulse linearity
- instrument triggering threshold
- maximum count rate.

Their significance varies according to whether the illumination is about constant (small dynamic range) or widely varying (large dynamic range).

Small dynamic range. This is often encountered in spectrometry applications, where optimum time or energy resolution is required. The important thing is to minimize response fluctuations by:
- optimizing the voltages at the electron-optical input system and the first two multiplier stages to minimize transit-time and gain fluctuations
- optimizing the collection of light at the photocathode.

Beyond the threshold set by the first of these recommendations, the practical minimum for the gain depends on the sensitivity or detection threshold of the signal processing circuits. The practical maximum is set either by dark current considerations or by linearity limits that come into play at pulse peaks. Beyond a certain value of applied voltage, the dark pulse rate increases faster than the gain (§3.1.3). The linearity limit may be due to either the photomultiplier or the circuits.

In pulse counting applications, the upper and lower gain limits can be found experimentally by plotting the variation of count rate as a function of applied voltage. This reveals a distinct counting plateau (see Fig.6.16) within which to set the operating point.

In other low dynamic-range applications, the choice of operating point may depend on other criteria, such as the bandwidth and gain of the signal processing circuits.

Large dynamic range This commonly applies in high-energy physics, where light pulses often vary over a wide range. The same fundamental rules apply as for low dynamic range, but four additional criteria outlined below are now decisive.

- The minimum number of photoelectrons per pulse that has to be detected.
- The sensitivity threshold of the electronics, which determines the amplitude or charge of the minimum detectable anode pulse.
- The maximum allowable anode pulse charge, which depends on the linearity limit of the tube at the chosen operating voltage.
- The ratio of maximum to minimum pulse amplitude or number of photoelectrons per pulse (dynamic range). If this is referred to the anode, and the minimum number of photoelectrons is very small, it should be expressed in terms of pulse charge rather than amplitude. So long as the interval between successive photoelectrons is shorter than the response pulse width of the tube, the anode pulse resembles a unique multi-electron pulse. But when the interval is longer than the response pulse width, the tube resolves the individual electrons into discrete single-electron pulses. Under these conditions it makes no sense to speak of amplitude.

Figure 5.19 illustrates how these criteria determine the practical boundaries of operation. The figure shows the log-log relation between the mean number of photoelectrons per pulse, $\overline{n}_{k,i}$, and the gain G

$$\bar{n}_{k,i} = \frac{\bar{q}_{a,i}}{Ge} \tag{5.10}$$

with the mean anode charge per pulse, $\bar{q}_{a,i}$, as parameter. The lines corresponding to each value of $\bar{q}_{a,i}$ are called *isocharge lines*. The practical boundaries of operation are:

- A horizontal line corresponding to the minimum number of photoelectrons per pulse, $n_{k,i(min)}$.
- An isocharge line corresponding to the charge sensitivity threshold of the electronics, $q_{a,i(min)}$.
  If the threshold is given in terms of anode pulse voltage, $V_{a(min)}$, rather than charge, the conversion is

$$q_{a,i(min)} \approx \frac{V_{a(min)}\ t_w}{R} \tag{5.11}$$

  where R is the anode load resistance and t_w the FWHM of the anode pulse, i.e. introducing time as an additional parameter.

- An isocharge line corresponding to the charge linearity limit of the tube, $q_{a,i(max)}$. This is usually given as a pulse current, $\hat{I}_{a(max)}$, at a specific operating voltage; the conversion is

$$q_{a,i(max)} \approx \hat{I}_{a(max)}\ t_w \tag{5.12}$$

  once more introducing the FWHM of the anode pulse as an important additional parameter. This boundary cannot be found until the operating voltage is known or at least estimated. Its estimation depends on the fourth of the listed criteria: the dynamic range. How to arrive at a realistic estimate can best be explained by example.

Consider a photomultiplier in pulse operation under the following conditions:

– minimum number of photoelectrons per pulse	10
– sensitivity threshold of the electronics	4 pC
– required dynamic range	100
– anode pulse full width at half maximum	5 ns
– pulse-peak linearity limit at V_{ht} = 2500 V	250 mA

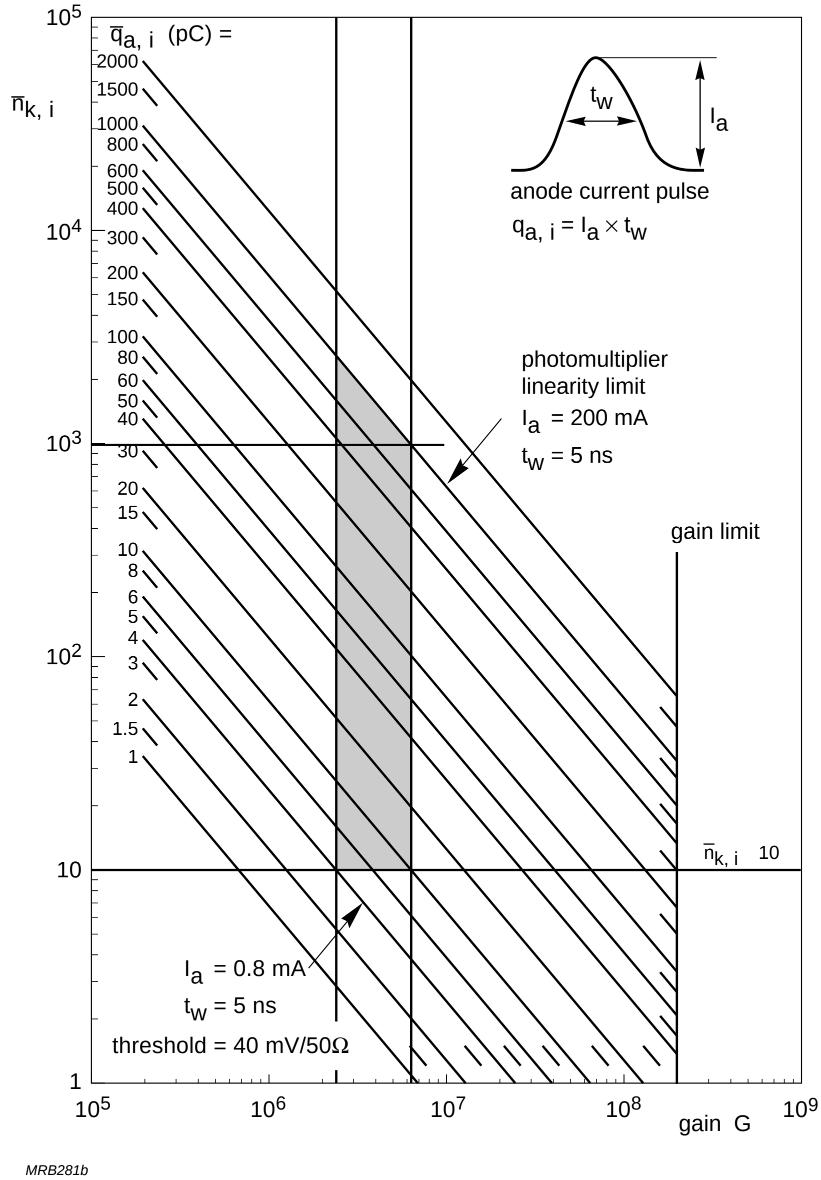

Fig.5.19 Photomultiplier working diagram in terms of photoelectrons per pulse, $n_{k,i}$, gain, and mean anode charge per pulse

In Fig.5.19 the first of these conditions corresponds to the bold horizontal line at $n_{k,i} = 10$; the second corresponds to the isocharge line $q_{a,i} = 4$ pC (third from bottom). Their intersection defines the minimum acceptable gain, about 2.5×10^6. Assume that, with the type of tube and voltage divider used, this is obtained at $V_{ht} = 2200$ V. At that voltage the pulse-peak linearity limit, $\hat{I}_{a(max)}$, given above is reduced by the ratio

$$\frac{\hat{I}_{a(max)}}{250} = \left(\frac{2200}{2500}\right)^n \tag{5.13}$$

where the exponent n is between 2 and 3, depending on the type of tube. For the present case, assume it is 2; then $\hat{I}_{a(max)} \approx 200$ mA and, from Eq.5.11 and $t_w = 5$ ns,

$$q_{a,i(max)} = 1000 \text{ pC}$$

which is indicated by the second isocharge line from the top. This is the third boundary of the practical operating region.

The fourth boundary is set by the required dynamic range which implies a maximum of 1000 photoelectrons per pulse. The $q_{a,i} = 1000$ pC isocharge line crosses the $n_{k,i} = 1000$ ordinate at a point corresponding to $G = 6.25 \times 10^6$. Thus, the practical boundaries of operation are:

$$\left.\begin{array}{l} n_{k,i} = 10 \\ q_{a,i} = 4 \text{ pC} \end{array}\right| \Rightarrow G_{min} = 2.5 \times 10^6 \qquad \left.\begin{array}{l} n_{k,i} = 1000 \\ q_{a,i} = 1000 \text{ pC} \end{array}\right| \Rightarrow G_{max} = 6.25 \times 10^6$$

If $G \leq 2.5 \times 10^6$, pulses containing less than 10 photoelectrons do not exceed the sensitivity threshold of the electronics. And if $G > 6.25 \times 10^6$, the required dynamic range of 100 cannot be accommodated.

Defining these boundaries gives a first approximation of the required operating point. It is then necessary to check that a gain in the indicated range can be obtained at the assumed $V_{ht} = 2200$ V. If not, a new value of V_{ht} will have to be assumed and the boundaries redetermined.

Uncertainty about the value of n in Eq.5.13 reflects on the accuracy with which the upper gain limit can be determined; however, this is seldom significant except when the gain operating range is narrow. Also, the current linearity limit given in the data sheets is a nominal value from which individual tubes may deviate; allowance should be made for this in the calculations. The pulse peak linearity mostly departs very slowly from the 2% linearity limit and higher pulses may still be linear to within say 3 to 4%.

In some cases of large dynamic range operation, other criteria have to be taken into account; especially, the high mean anode current due to a high count rate may affect stability (§4.6.1), necessitating reconsideration of the initial parameters by, for example, reducing the tube gain and compensating for this with additional gain in the preamplifier.

5.6 Gain adjustment

Gain characteristics differ from tube to tube. Sometimes, though, it is necessary to ensure that a number of tubes working together operate at equal gain. There are two ways to do this.

Supply voltage adjustment. Gain can be adjusted by adjusting the high voltage supplied to each tube. If the tubes do not have separately adjustable supplies but are fed from a common supply, their voltages can still be adjusted by ballast resistors connected in series with their respective voltage dividers. Even though the current from the supply is practically constant, the ballast resistors should be decoupled.

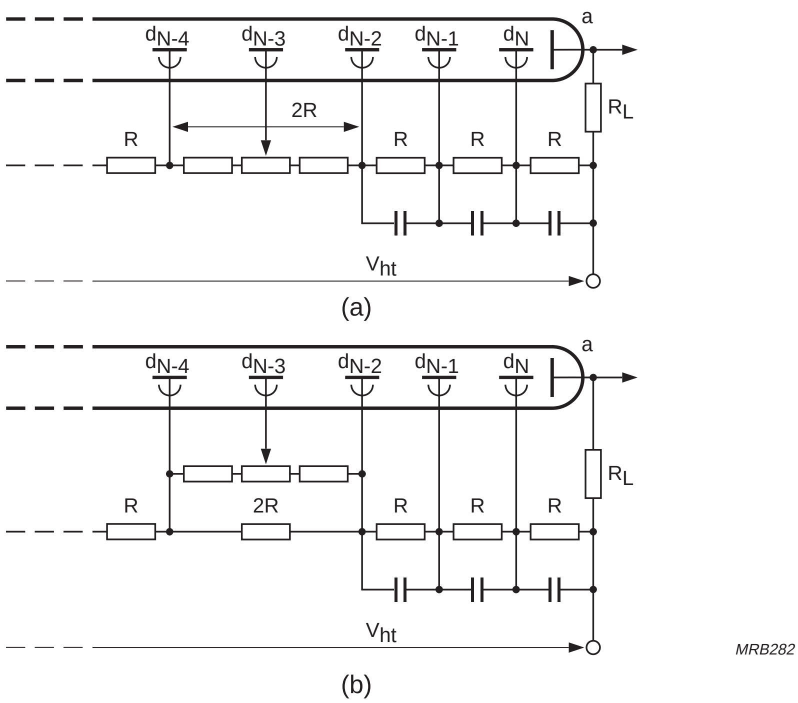

Fig.5.20 Alternative circuits for adjusting gain by adjusting the voltage of one dynode

Dynode voltage adjustment. This is often used when it is not practical to adjust the high voltage supply to each tube. Gain can be altered by altering the voltage of any dynode (see Fig.4.10), but an intermediate one is always chosen to avoid interfering with the collection efficiency of the electron-optical input system or the output stage. Of the two adjustment circuits shown in Fig.5.20, the (b) version is preferable if the divider current is high; it makes it possible to use a high-value potentiometer (about 1 MΩ) with a low power rating (≤0.75 W). In both the (a) and (b) versions, resistors should be connected on both sides of the potentiometer to limit its working voltage; in practice, the range of control variation required is usually far less than the maximum possible. As all terminals of the potentiometer are at a fairly high voltage, the potentiometer must be well insulated.

Dynode voltage adjustment is more effective with focusing than with venetian-blind dynodes. A disadvantage of it is that it can impair stability and increase susceptibility to magnetic fields; on the other hand, time characteristics are relatively unaffected.

5.7 Supply for multiple tubes

When many photomultipliers are used together (as in hodoscopy, scintigraphy, tomography) the high voltage can be supplied either separately to each or by a single supply common to all.

Separate supplies are preferable. They prevent any reaction between tubes and, if they are adjustable, facilitate individual gain adjustment. Compact, adjustable and non-adjustable, individual supply modules are marketed, as well as supplies with several, separately adjustable output channels. The latter, though, are usually bulky and expensive.

Common supply to a number of tubes is often used when the current required from each is low. Each then has its own voltage divider to minimize reaction between tubes and its own provision for gain adjustment, which may be a potentiometer either in series with the divider (Fig.5.21) or controlling the voltage of one of the dynodes (Fig.5.22).

If the application requires the voltages at the electron-optical input system or the output stages to be zener stabilized, it may be advantageous to use a single zener diode (or series-connected group of diodes) for all the parallel-connected tubes. This ensures sufficient current through the diodes even when the current through the individual voltage-divider chains is low.

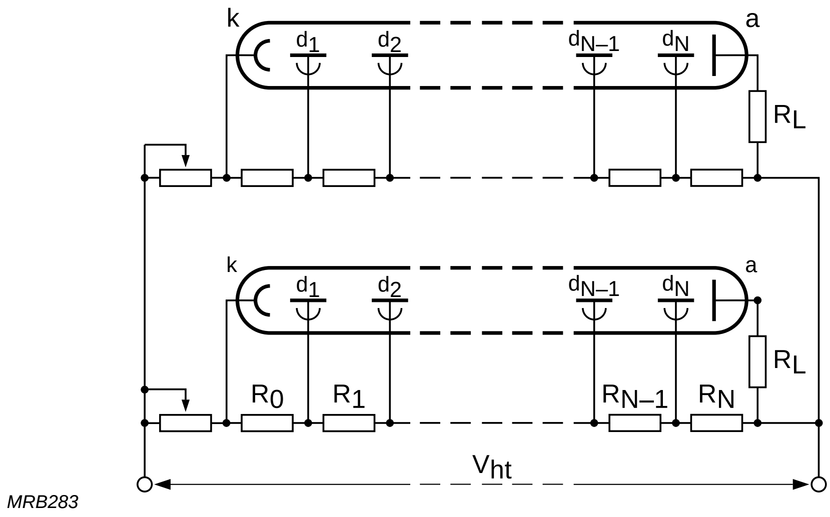

Fig.5.21 High-voltage supply to photomultipliers in parallel

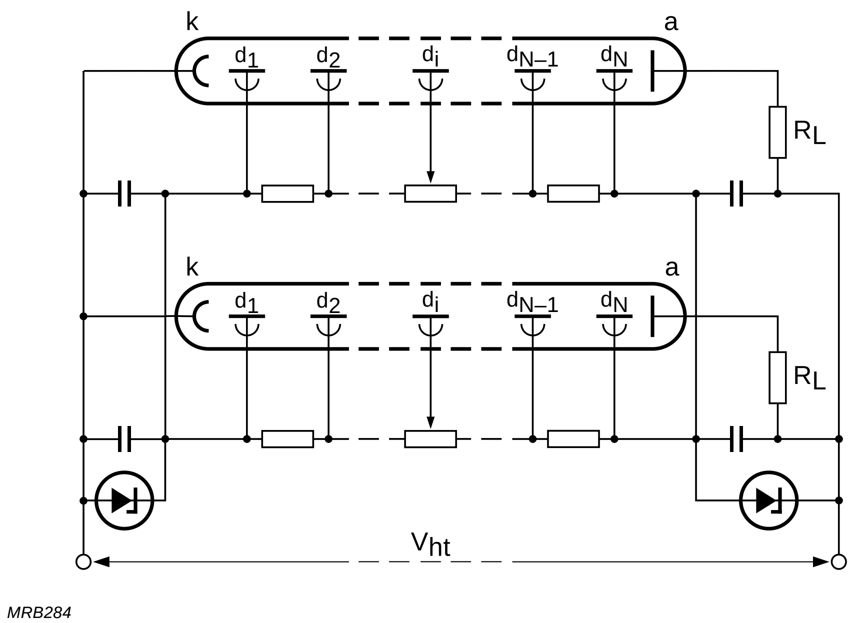

Fig.5.22 High-voltage supply to photomultipliers in parallel, with common zener-diode stabilization

5.8 Dark current reduction

5.8.1 Cooling

As the dark current is partly thermionic it can be reduced by cooling the tube, but only down to a certain temperature depending on the photocathode composition. Below that temperature leakage current, field emission, and other temperature-independent components of the dark current predominate. Depending on the requirements of the application, cooling can be in a cold chamber, by refrigerating coils, or by Peltier coolers. Whichever is used, precautions against condensation should be taken by enclosing the tube either in a partial vacuum or a dehumidified atmosphere. Moisture condensed on the tube impairs insulation, increases leakage current, and can affect the transmission of the input window.

It is preferable to connect an artificially cooled tube in positive polarity so that no special precautions need be taken as regards insulation. If the tube does have to be connected in negative polarity, a thermally conductive electrical insulator must be interposed between it and the cooling provision.

5.8.2 Reducing effective photocathode area

If less than the whole area of the photocathode is to be used (as when working with collimated light, for example), dark current can be reduced by rendering the unused area inactive. This eliminates the thermionic component of the dark current from the inactive area and is therefore the more rewarding the larger that component is. It is particularly advantageous with S20, S20R, and S1 cathodes which, being more sensitive toward the long wavelengths, have relatively high thermionic dark current components.

An outer zone of the cathode can be made inactive by defocusing the electrons from it so that they are not collected by the first dynode. This can best be done magnetically, either

- by means of an axially-magnetized toroidal permanent magnet concentric with the cathode and slightly in front of it. Figure 5.23 illustrates the effective-area reduction, plotted along one cathode diameter, that can be obtained in this way.
- or by means of a solenoid surrounding the cathode. Figure 5.24 illustrates the area-reduction effect, and Fig.5.25 compares this with the effect on dark current. The dark current decreases less rapidly than the effective area because not all of it originates from the cathode. A potential disadvantage of the electromagnetic method is that heat dissipated in the solenoid may raise the cathode temperature.

All these methods are more difficult to use with small-diameter tubes. Such tubes seldom have an accessible focusing electrode, and the input to the electron multiplier is often off-centre; a centred magnet or solenoid therefore upsets the electron-optical input system.

Dark current can also be cancelled or compensated electronically (§5.11.2), and dark pulses can be excluded by coincidence techniques. However, neither these nor any of the methods described above has any effect on signal noise, which is fundamentally irreducible (Chapter 3).

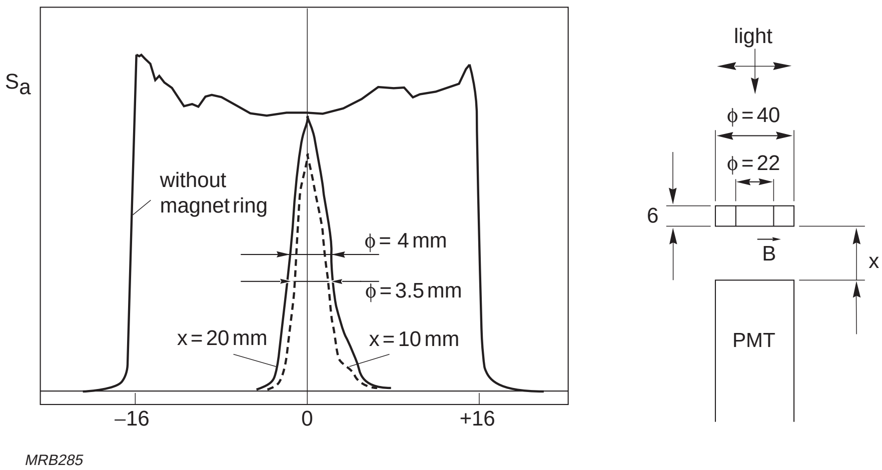

Fig.5.23 Effective-cathode-area reduction due to a permanent-magnet ring in front of the cathode (32 mm diameter S1 cathode)

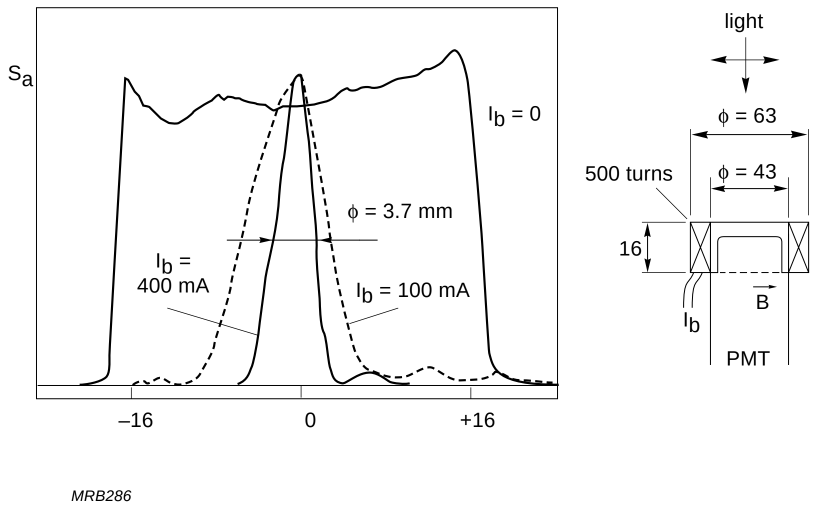

Fig.5.24 Effective-cathode-area reduction due to a solenoid surrounding the cathode (32 mm diameter S1 cathode)

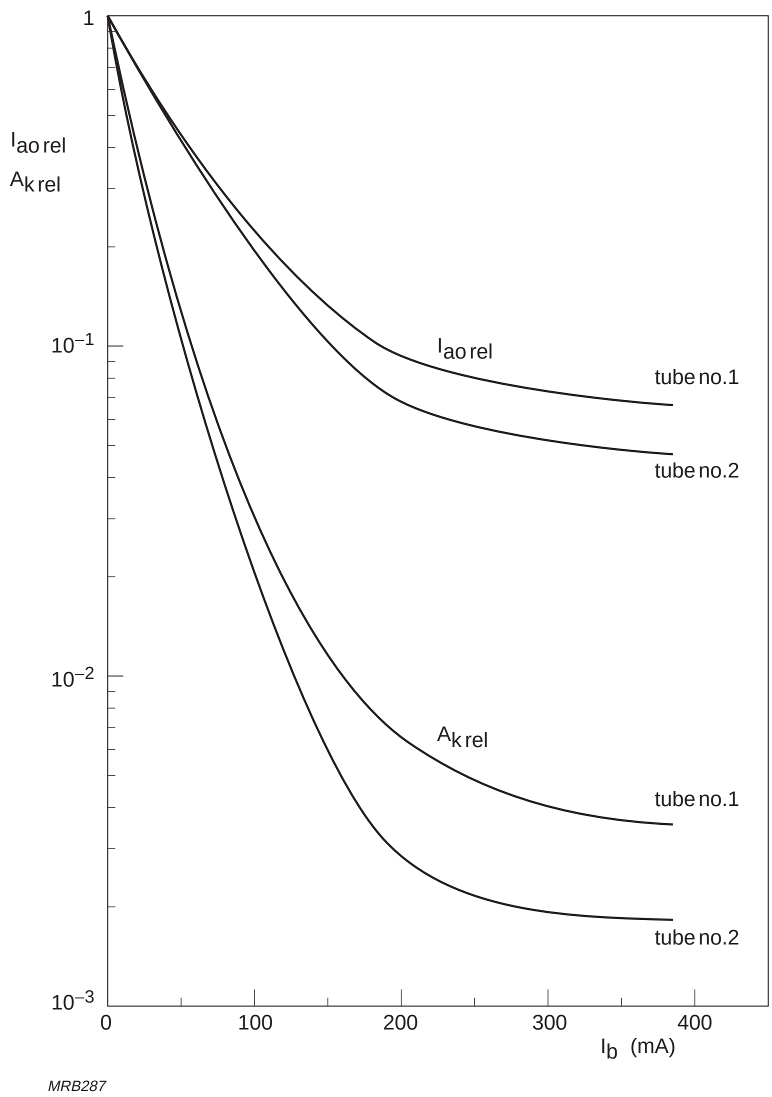

Fig.5.25 Relative dark-current (I_{a0}) and cathode-area (A_k) reduction as functions of solenoid current I_b in two tubes with 32 mm diameter S1 cathodes, using the method of Fig.5.24. Note that the dark-current reduction is much less than the effective-cathode-area reduction

5.9 Magnetic shielding

Since fields as weak as the earth's can affect sensitivity (§4.8.2), a mu-metal shield is always desirable. At flux densities of more than a few milliteslas, however, such a shield saturates and becomes ineffective. It must then be surrounded by a supplementary shield, usually of soft iron. Shielding effectiveness, based on material and dimensions, can be determined from Fig.5.26. For example, a mild-steel shield with an inner diameter of 70 mm and thickness 5 mm ($r_{in}/r_{ext} = 0.875$) in a flux density of 100 mT gives an attenuation of about 30. An inner mu-metal shield with an inner diameter of 57 mm and thickness of 0.8 mm ($r_{in}/r_{ext} = 0.973$) gives an additional attenuation of about 300, leaving a residual flux density that is well below the 0.05 mT to 0.1 mT sensitivity threshold of most tubes.

Magnetic shields should extend about one diameter beyond the cathode plane, especially if the magnetic flux density is high. If the tube is operated in negative polarity, the shielding must be completely insulated from the glass (§5.1.1) or connected to the cathode potential via a protective resistor.

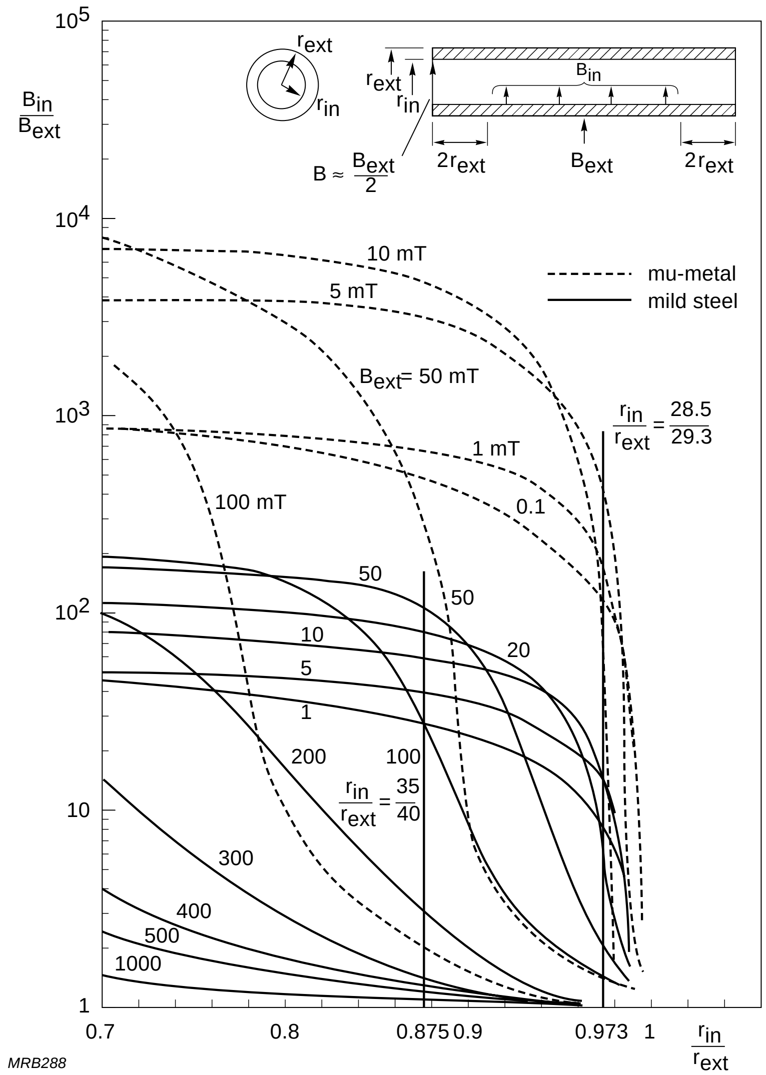

Fig.5.26 Comparative effectiveness of mu-metal and mild-steel magnetic shields

5.10 Gating

In applications in which there is intermittent exposure to high illumination, it may be desirable to guard the tube or its circuits against overloading. This can be done by gating the tube via the high voltage supply, the cathode, one or more of the dynodes, or the cathodes and dynodes together.

High-voltage supply gating. The tube can be gated at any desired frequency by chopping the high-voltage supply but, as this requires a low-impedance source of high-voltage pulses, it is usually not practical. Moreover, the anode sensitivity only reaches its maximum some 10 μs to 50 μs after the high voltage does. Finally, high-voltage supply chopping aggravates dark-current and gain instabilities.

Positive cathode pulse gating. Applying a positive pulse to the cathode has similar disadvantages to high-voltage supply chopping. Moreover, the photo-sensitivity of the first dynode limits the effectiveness of the gating.

Negative focusing-electrode gating. Provided the focusing electrode is independently accessible, applying a 50 V to 100 V negative pulse to it is the easiest method of gating. However, the effectiveness of this too tends to be limited by first-dynode photoemission.

Negative-dynode gating. Applying a negative pulse via a capacitor or coaxial-cable transformer to one or more of the dynodes (usually the 3rd to the 5th) can reduce gain several hundredfold, depending on the amplitude of the pulse.

Combined cathode and dynode gating. During negative-dynode gating, photoemission continues. If the illumination is intense the recovery time is long and there is a risk of cathode fatigue. There is also a risk of impaired gain stability due to scattered charges built up on the glass and insulators by electron bombardment. A way out of these difficulties is to gate a dynode and the cathode together, simultaneously lowering gain and suppressing photoemission. This can be done by applying a pulse between the cathode and the third dynode, using a separate voltage divider earthed at the third dynode to supply the first two stages. Although the amplitude of the pulse has to be several hundred volts, gain can be reduced by a factor of 10^{-4} in this way.

The tube can be permanently blocked and then unblocked whenever an event to be counted or measured is expected. This is difficult, however, the unblocking pulse amplitude, on which the gain of the tube depends, has to be very precisely controlled.

To protect the tube against excess illumination an external coincidence circuit is preferable, such a circuit might make use of a gate in the anode circuit.

Another gating technique described in the literature makes use of a cross-bar grid fitted to the input window to localize the area of photoemission.

5.11 Drift compensation

5.11.1 Anode sensitivity drift

Anode sensitivity drift can be compensated by
– varying the gain of the tube or the preamplifier of the measuring circuit
– varying the illumination.

Varying the gain. Figure 5.27 illustrates the principle. A signal due to a reference illumination of the tube is compared with a fixed reference level to derive an error signal. The error signal is fed to a control circuit which regulates
– the high-voltage supply
– the potential of one dynode
– or the gain of an amplifier in the anode circuit
to equalize the illumination signal and the reference signal so as to make the error signal zero.

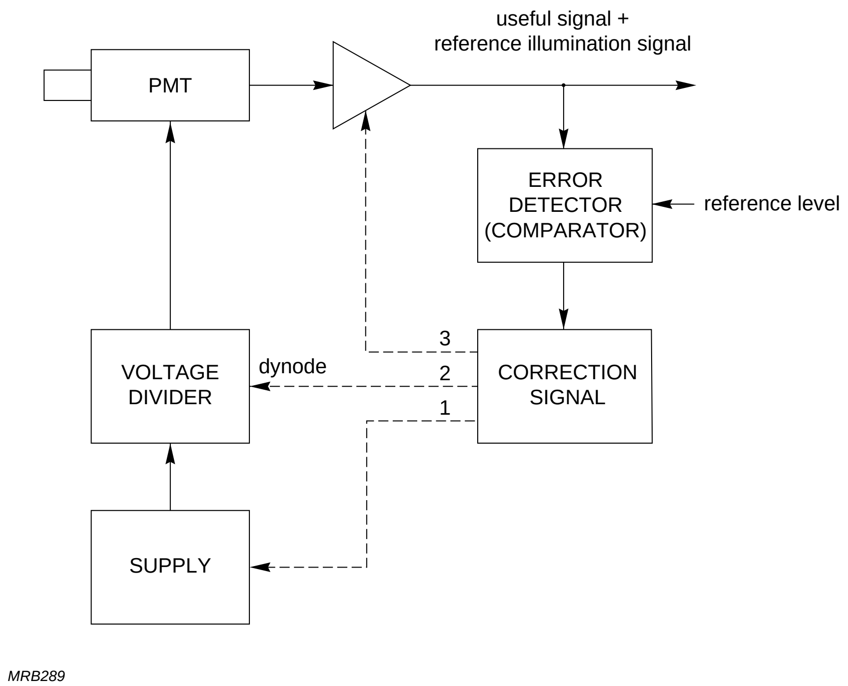

Fig.5.27 Correction of anode-sensitivity drift by varying the gain. The signal due to a reference illumination of the cathode is compared with a fixed reference level to derive an error signal. The gain can be altered by any one of the three routes shown

Depending on the type of application, the reference illumination may be a calibrated radioactive (α or γ) source/scintillator combination or a LED. A calibrated radioactive source integral with the scintillator normally used with the tube makes it possible to compensate drift in both the anode sensitivity of the tube and the luminous efficiency of the scintillator at the same time.

Systems incorporating microprocessors for compensating the drift of several photomultipliers at once are described in the literature.

Varying the illumination. Gain shift due to wide variation of illumination can be reduced with the aid of a constant source of supplementary illumination, or compensated by means of a variable one.

Constant supplementary illumination. When the mean anode current is only 20-30 nA but the relative variation is large, gain shift due to the variation can be reduced by using a constant source of supplementary illumination (e.g. a LED) to increase the mean anode current. A capacitor preceding the signal processing circuits eliminates the resulting DC component.

This method has the disadvantage that operating the tube at a higher mean anode current may introduce another causes of gain instability – long-term drift.

Variable supplementary illumination. In pulse applications the gain shift due to variations of mean anode current can be compensated by a supplementary light source (LED) which can be varied in opposition to the mean anode current variation. Figure 5.28 illustrates the principle. Integrated anode pulses are compared with a fixed reference to generate an error signal which controls the current to the LED; the greater the difference between the integrated pulses and the reference, the more light the LED emits. Once again, a coupling capacitor preceding the signal processing circuits eliminates the DC component due to the LED.

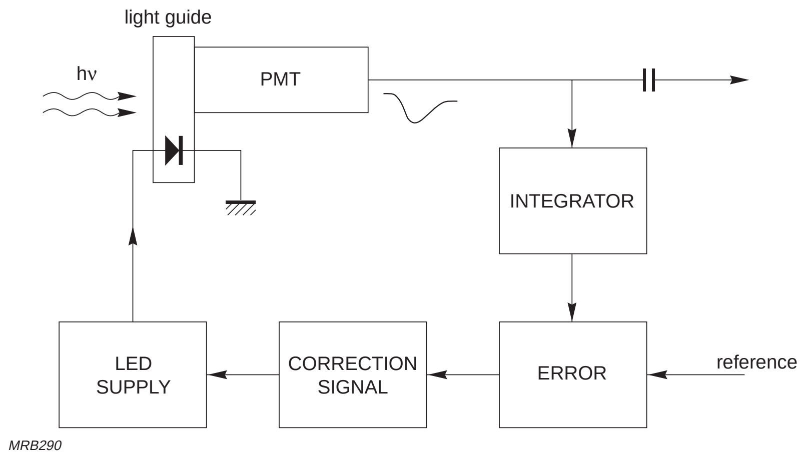

Fig.5.28 Correction of gain drift by varying a supplementary light source

5.11.2 Temperature-dependent dark-current drift

Temperature-dependent dark-current drift can be compensated by means of a current source (e.g. a selected diode or transistor) obeying the same temperature law as the dark current. Figure 5.29 shows a typical circuit. When there is no signal from the photomultiplier the output of the differential amplifier is $V_s = V_0 - RI_{a0}$, where I_{a0} is the dark current. V_0 is adjusted to make $V_s = 0$ at the prevailing temperature. Provided the transistor from which V_0 is derived follows the same temperature law as the dark current, the voltage V_s due to the dark current remains zero at other temperatures. Diodes or transistors that follow exactly the same law as the dark current of a given photomultiplier are not easy to find, but a selection can usually be made that gives no more than a small discrepancy over a practical temperature range.

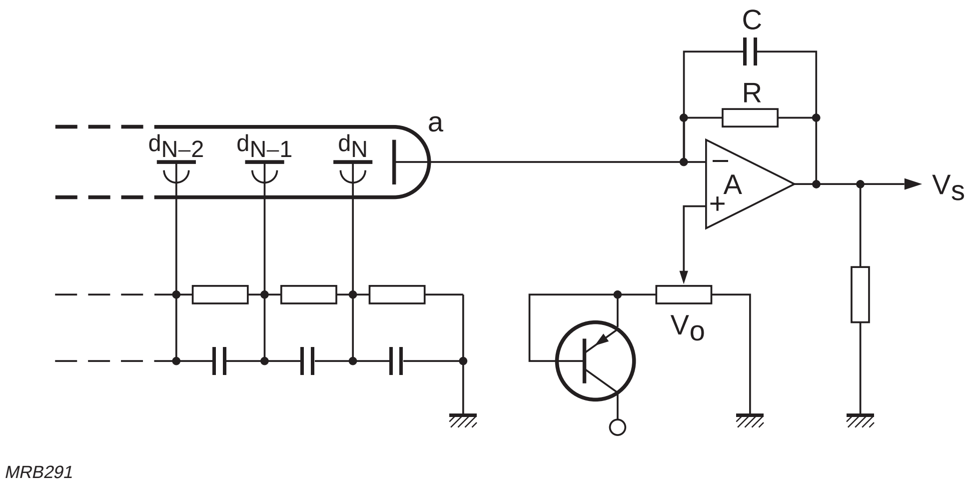

Fig.5.29 Circuit for compensating temperature-dependent dark-current drift

5.12 Pin connections and safety precautions

To save space it is sometimes necessary to make connections direct to the pins of a tube with glass base, instead of using a socket. Do so with contact clips if possible; if it is not possible, spot-weld the connections. *Do not solder them*; soldering heat can crack the glass.

For such applications it is also possible to order tubes with flying leads which can be safely soldered.

Internally connected pins (marked 'i.c.' in the data sheet) must never be connected. They are used only during manufacture of the tube. If used, the applied potential may distort the field distribution within the tube.

Non-connected pins (marked 'n.c.' in the data sheet) should not be used either, even as intermediate wiring points. They are provided only to facilitate interchangeability of similar tubes (i.e. an electrode that is externally connected in one type may be internally connected in a similar type using the same socket and operating voltages). Connections made to an n.c. pin could cause leakage currents or local insulation breakdown.

Ensure that there are adequate safeguards for those working near high voltage. Voltages as high as 3 kV are used with some tubes, and some power supplies can deliver more than 10 mA.

Handle photomultipliers with due regard for the fact that they contain a high vacuum, glass is fragile, and the flat input window is highly stressed. Take particular care with tube of 100 mm diameter or more, because of risk of implosion.

5.13 Signal processing

In a few applications involving steady-state operation the anode current of a photomultiplier can be read from a galvanometer. In most applications, however, and especially those in which the tube forms part of a control system, amplification is necessary, and the amplifier must have stability and noise characteristics comparable with those of the tube. Integrated-circuit operational amplifiers that meet this requirement are available nowadays fairly inexpensively. Such an amplifier is usually a direct-coupled differential amplifier with very high input impedance (>10 MΩ) and high open-loop gain (>10^5) and near-zero output impedance. The transfer function of the amplifier is determined solely by the input and feedback networks associated with it.

5.13.1 Operational amplifiers

Figure 5.30 shows an operational amplifier connected as a voltage amplifier. If A is the open-loop gain of the amplifier, $V_s = A(V_2 - V_1) = -AV_1$. Because of the very high input impedance of the amplifier, the current I_3 is negligible; therefore,

$$I_1 = \frac{V_e - V_1}{R_1} = I_2 = \frac{V_1 - V_s}{R_2}$$

and the gain with feedback is

$$G = \frac{V_s}{V_e} = -\frac{\dfrac{R_2}{R_1}}{1 + \dfrac{1}{A} + \dfrac{R_2}{AR_1}}$$

which, since A is very large, reduces to $G = -R_2/R_1$.

To ensure stability the resistances R_1 and R_2 must be very stable. The same holds for the feedback resistance R_f in the current/voltage converter of Fig.5.31, a circuit which finds wide application in view of the fact that a photomultiplier is an almost ideal current generator.

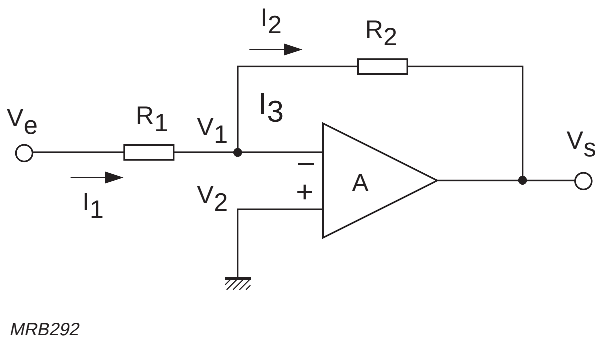

Fig.5.30 Voltage amplifier (inverting)

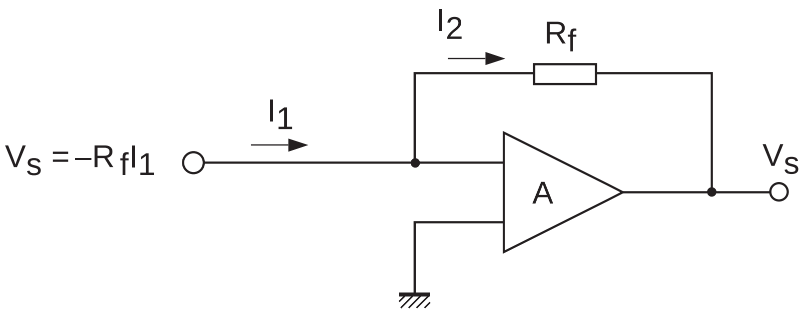

Fig.5.31 Current-voltage converter

Operational amplifiers can also be used for such mathematical operations as addition, subtraction, and integration; see Fig.5.32 to 5.34. A practical example of subtraction, for instance, is the arrangement for compensating temperature-dependent dark-current drift shown in Fig.5.29. Integration is required in some scintillation counting applications, where it is necessary to measure the quantity of light contained in individual scintillations, whatever their pulse shape or duration. The anode charge q_a due to a single scintillation is proportional to the quantity of light it contains. Integrating the anode current over the duration of the scintillation gives a voltage V_s proportional to q_a and, hence, to the quantity of light. An electronic switch (represented by S in Fig.5.34) discharges the integrating capacitor between scintillations.

In applications in which the photomultiplier output can vary by several orders of magnitude it is often desirable to convert it to logarithmic form. This can be done by incorporating a diode or transistor in the feedback loop (Fig.5.35) so as to exploit the logarithmic characteristic of a forward-biased p-n junction. A transistor has an advantage over a diode in that earthing its base establishes an exact zero reference for the output voltage.

When the photomultiplier output is a very low-level, varying direct current, the zero-drift of a DC amplifier can be a significant source of error. An alternative (Fig.5.36) is then to chop the signal at a frequency must higher than its own range of variation, amplify the resulting modulated squarewave signal in a stable AC amplifier, demodulate the amplifier output, and restore the original but now amplified DC signal by means of a low-pass filter.

$$V_S = \left(\frac{R_f}{R_1}V_1 + \frac{R_f}{R_2}V_2 + \frac{R_f}{R_3}V_3 + \cdots\right) \qquad V_S = R_f(I_1 + I_2 + I_3 + \ldots..)$$

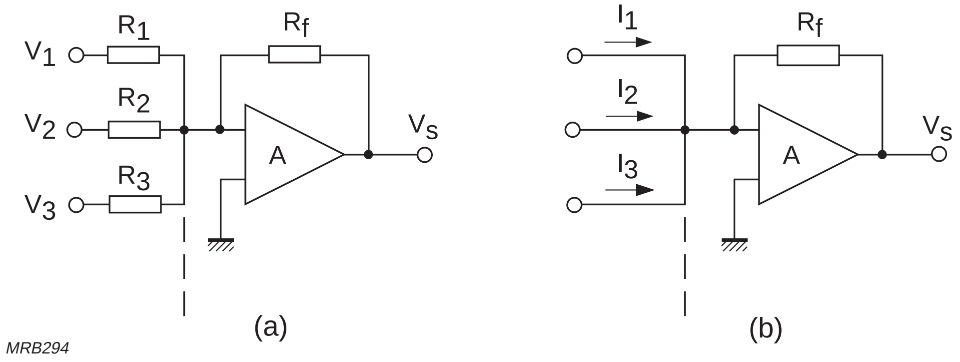

Fig.5.32 (a) Voltage adder, (b) current adder

$$V_S = \frac{R_f}{R_1}V_1 - \frac{R_f}{R_2}V_2 \qquad V_S = R_f(I_1 - I_2)$$

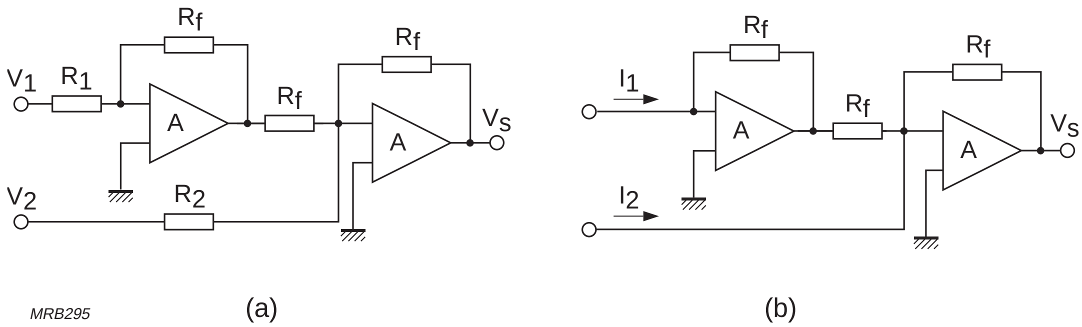

Fig.5.33 (a) Voltage subtractor, (b) current subtractor

$$V_S = -\frac{1}{C}\int_0^{t_{min}} I_a(t)dt \simeq -\frac{\bar{q}_{a,i}}{C}$$

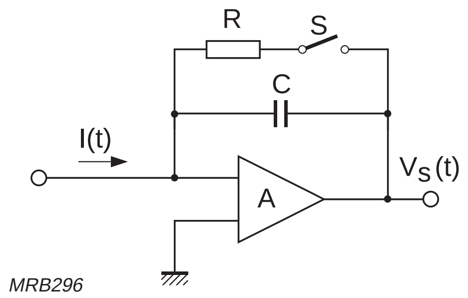

Fig.5.34 Integrator. For integrating the charge due to light pulses, t_{min} must exceed the pulse duration. Switch S discharges the capacitor between pulses

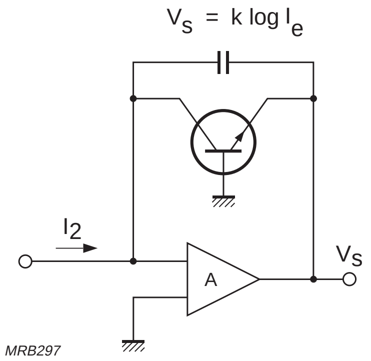

Fig.5.35 Logarithmic amplifier

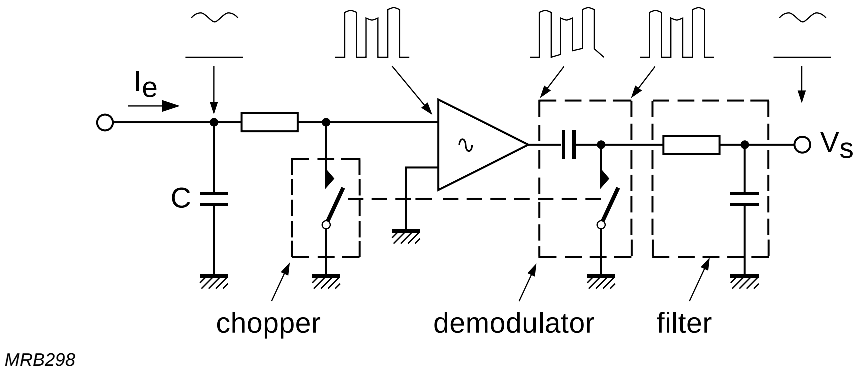

Fig.5.36 Chopper-stabilized amplifier

5.13.2 Separating the signal from the noise

Synchronous detection. A principle similar to that of the chopper stabilized amplifier is often used in photometric instruments when the signal information is known to occupy only a narrow bandwidth; it is based on reducing the bandwidth of the measurement to increase the signal-to-noise ratio (see §3.3.3). The light flux to be measured is first chopped at a frequency at least twice its own highest frequency of variation (Fig.5.37) so that the resulting photomultiplier output is itself a squarewave carrier modulated by the signal information. The modulated carrier is amplified and then demodulated with the aid of a reference signal synchronous with the chopper that switches the gain of a mixer between +1 and −1. A low-pass filter at the mixer output eliminates carrier frequency components and leaves an amplified continuous signal proportional to the input.

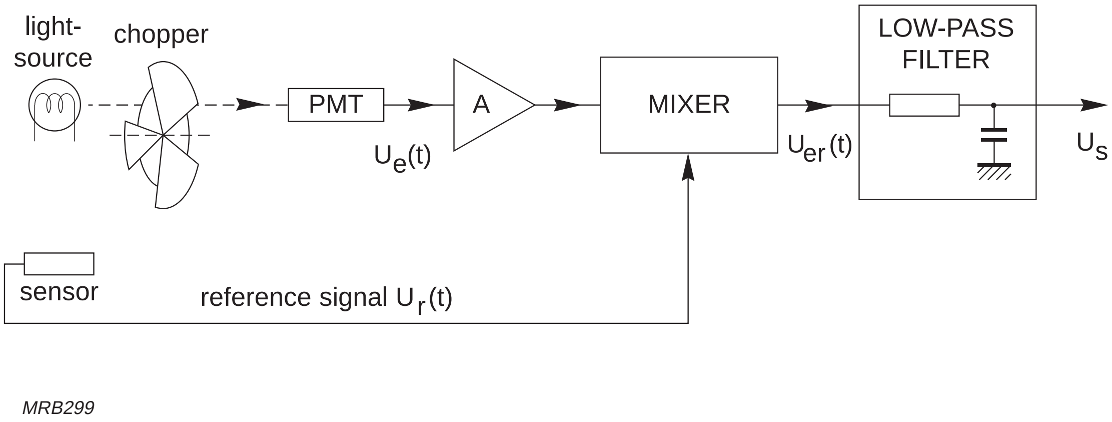

Fig.5.37 Synchronous detection

Autocorrelation. Another method for separating the signal from the noise is based on the autocorrelation function

$$\gamma(\tau) = \frac{1}{T}\int_0^T U_1(t)U_2(t - \tau)\,dt$$

where U_1 and U_2 represent signals to be discriminated. When $U_1 = U_2$ the autocorrelation function $\gamma(\tau)$ attains a maximum at $\tau = 0$. If the signal is periodic, $\gamma(\tau)$ is also periodic and has the same period. If the signal is aperiodic, $\gamma(\tau) = 0$ except when $\tau = 0$, in which case $\gamma(\tau)$ corresponds to an impulse function. Therefore, in a signal-processing system incorporating a delay channel for discriminating signals from noise by autocorrelation, aperiodic noise is quickly suppressed and periodic signals become clearly distinguishable as soon as the delay τ is made significantly greater than zero.

Figure 5.38 shows an autocorrelation system for extracting a low-level periodic signal, $U(t) = U_o\,(1 + \cos\omega t)$ from a photomultiplier output which is corrupted by a noise u(t). After being bandpass filtered around the frequency ω, the photomultiplier output is applied to two equal-gain amplifier channels, one incorporating a delay τ and the other undelayed. The outputs of the two channels are multiplied together and integrated over a time T, giving

$$\gamma(\tau) = \frac{1}{T}\int_0^T U(t)U(t-\tau)\,dt = \frac{A^2}{T}\int_0^T \{U_o(1+\cos\omega t) + u(t)\}\,[U_o\{1+\cos\omega(t-\tau)\} + u(t-\tau)]\,dt$$

The discrimination principle rests on finding the least value of τ which is an integral multiple of the signal period and for which the fluctuating component, corresponding to terms containing the noise u(t), becomes negligible. The equation then simplifies to

$$\gamma(\tau) \approx \frac{A^2U_o^2}{T}\int_0^T (1+\cos\omega t)\,\{1+\cos\omega(t-\tau)\}\,dt \approx \frac{A^2U_o^2}{2}\,(2+\cos\omega\tau)$$

which, since τ is chosen as an integral multiple of the signal period, reduces to

$$\gamma(\tau) \approx \frac{3}{2}A^2U_o^2$$

Thus, the output signal is proportional to the square of the input signal.

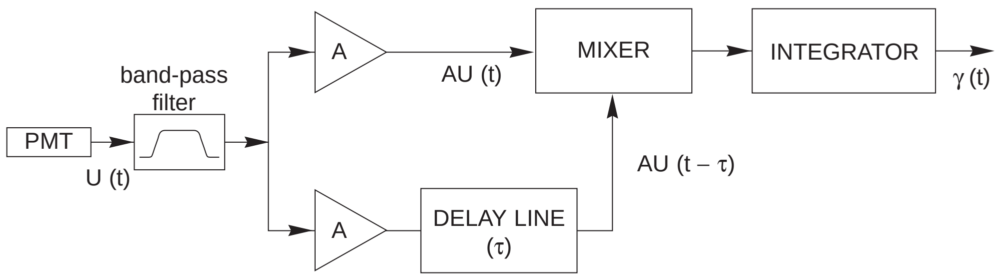

Fig.5.38 Autocorrelation detector

5.13.3 Detection at very low light levels

Quadratic detection. When a photomultiplier is exposed to a constant or slowly varying flux the resulting anode current consists of a DC component, which varies as the flux, and a fluctuating ‘shot-noise’ component which varies as the square root of the DC component. If the flux is very low the energy transferred by the fluctuating component can exceed that transferred by the DC component, and in that case it may be more expedient to measure the fluctuating than the DC component.

An advantageous way to do this is by quadratic detection, which gives an output proportional to the square of the fluctuation – and, hence, to the incident flux.

In the quadratic detector shown in Fig.5.39 the DC component of the voltage across the load resistor R_L is too low to be useful and the fluctuating component u(t) is amplified by a matched pair of wideband amplifiers. The amplifier outputs are multiplied by each other to produce a voltage $A^2u^2(t)$ which is then low-pass filtered to obtain its mean value, the output voltage U_s, which is thus proportional to the incident flux.

As the noise contributions of the two amplifiers are not correlated with each other, they do not significantly contribute to the output voltage U_s. However, the photomultiplier dark-current noise and the load-resistor thermal noise, which are applied equally to both channels, do contribute to the output voltage and are thus a factor limiting the sensitivity of quadratic detection.

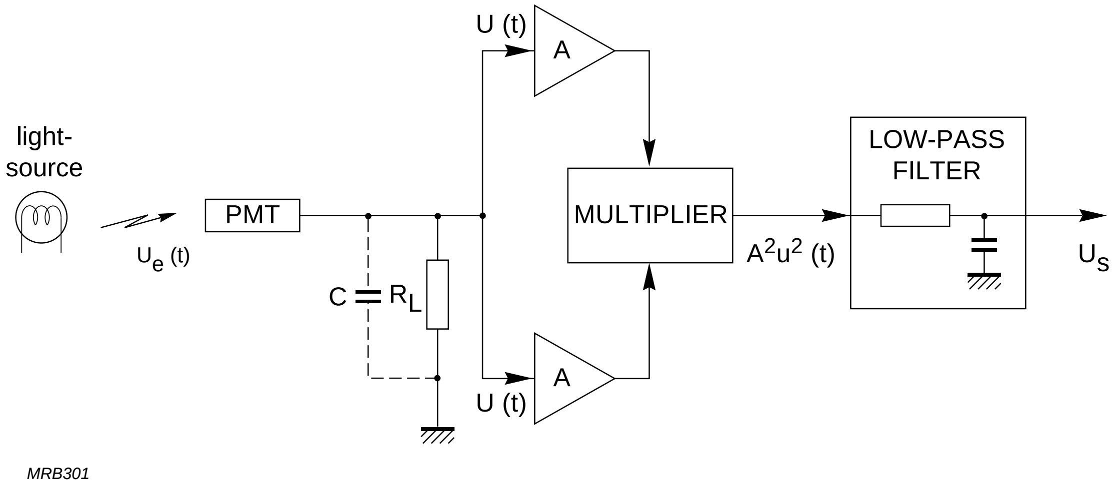

Fig.5.39 Quadratic detector

Photon counting. In photon counting the photomultiplier is operated under single-electron conditions. Flux levels as low as a few tens of photons per second can be measured – and the method has the advantage of making it possible to eliminate several otherwise disturbing factors from the measurement. Among these are the DC component of the dark current, low-amplitude pulses originating in the electron multiplier, and high-amplitude pulses of other than photoelectric origin. Photon counting can also be used for determining the shape of fast, low-intensity light pulses, as in certain applications of scintillation counting.

Low flux levels. In the photon-counting system of Fig.5.40, the level of the incident flux is such that the cathode emits only single electrons. The individual anode charges due to these are integrated to produce proportional voltage pulses which are passed by a discriminator to a pulse counter whose output over a set time is a measure of the incident flux.

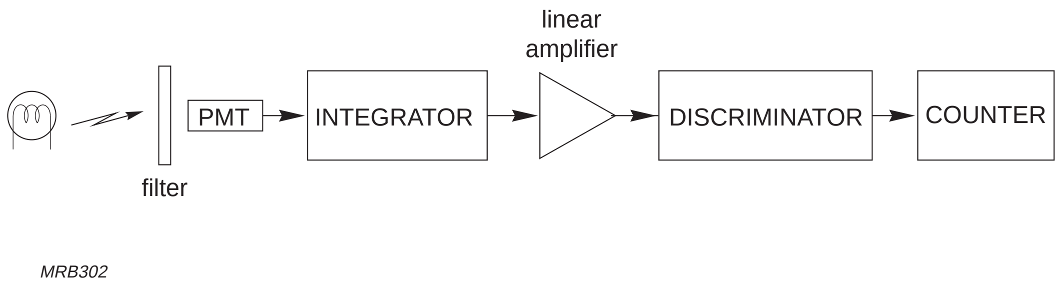

Fig.5.40 Photon-counting detector

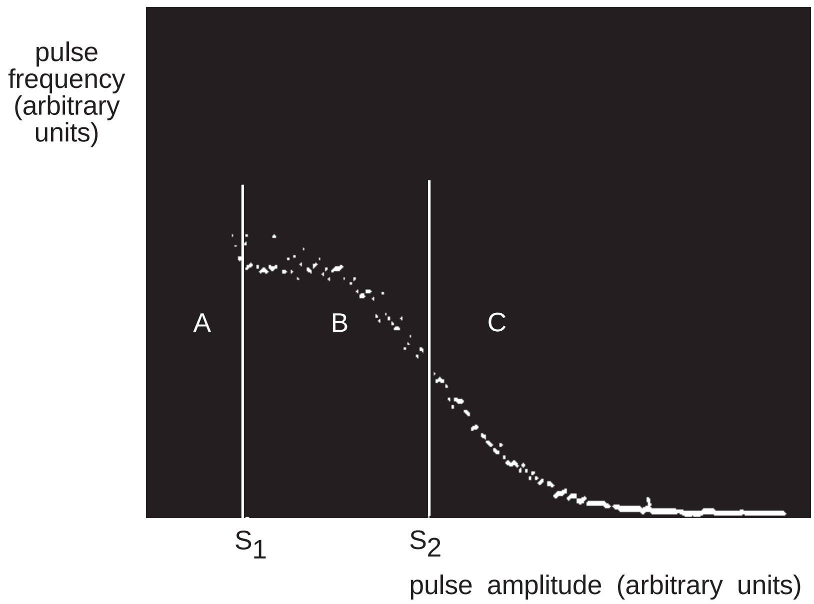

Fig.5.41 Pulse-amplitude spectrum of photomultiplier operating under single-electron conditions, showing lower and upper thresholds for photon counting (defining counting window B)

Because of statistical fluctuations in the electron multiplication, the amplitudes of the single-electron pulses are distributed according to the SES (§2.1.6). The dark noise pulses are distributed according to a spectrum whose general shape is often quite different (Fig.5.41) with a large quantity of very small pulses but also some very large pulses (cosmic rays, afterpulses). To optimise the S/N (signal/dark-noise) ratio, two thresholds S_1 and S_2 are adjusted to give the best pulse-amplitude 'window'.

However, there are also some dark pulses within the discriminator window, and the count must be corrected for them. If $\bar{n}_p$ is the mean number of incident photons and $\bar{n}_d$ the mean number of non-excluded dark pulses per unit time, the total count during a period τ will be $\tau(\bar{n}_p\rho\eta + \bar{n}_d)$, where ρ is the quantum efficiency of the cathode and η the collection efficiency of the electron-optical input system. Subtracting the number of dark pulses counted during a like period without illumination gives the correct count

$$n_e = \tau(\bar{n}_p\rho\eta + \bar{n}_d) - \bar{n}_d\tau$$

which (from §3.2) has a variance

$$\sigma_n^2 = \bar{n}_p\rho\eta\tau + 2\bar{n}_d\tau$$

At very low flux levels n_d is usually much larger than n_p and the signal-to-noise ratio becomes

$$\frac{S}{N} = \frac{\bar{n}_p\rho\eta\tau}{\sqrt{2\bar{n}_d\tau}}$$

which increases as the square root of τ. Thus, improving the signal-to-noise ratio by increasing the counting time is analogous to the improvement that can be obtained in charge-integration methods by narrowing the bandwidth.

The signal-to-noise ratio can also be improved by cooling the cathode (§5.8.1) to reduce the number of single-electron pulses of thermionic origin, and by magnetically reducing the cathode area from which thermionic electrons can be collected (§5.8.2).

Determining the shape of fast, low-intensity pulses. Repetitive pulses detected by a photomultiplier can be displayed in real time on an oscilloscope, provided their duration is more than a few tens of nanoseconds and their intensity corresponds to more than a few thousand photons. However, if their duration is short compared with the pulse response of the photomultiplier (a few nanoseconds), this is not possible, even if the number of photons per pulse is small enough for them all to be resolved. Then, a sampling method must be used and the pulse shape reconstructed statistically. For this it is essential for the transit-time fluctuations of the tube to be small.

Figure 5.42 shows a set-up for reconstructing (delineating) fast, low-intensity pulses by sampling. (Note the similarity to the transit-time spread measuring set-up of Fig.4.8). The pulses received by photomultiplier PMT1 are so attenuated that the tube operates under single-electron conditions. Those received by PMT2 are not attenuated but are delayed sufficiently to ensure a measurable time difference between the outputs derived from the two tubes. Discriminators 1 and 2 produce standardized timing pulses coinciding with chosen reference points on their respective input pulses. From the time differences between the discriminator pulses the time-to-amplitude converter and the multichannel analyser construct a histogram that reflects the photon distribution in the original light pulses.

The auxiliary channel at the bottom of the diagram minimizes statistical errors due to the effect of pulse-shape fluctuations and walk errors in the discriminators. The auxiliary channel gates the multichannel analyser only when the pulses from both photomultipliers satisfy predetermined amplitude criteria.

Let L(t) be the probability density of photon-emission instants during the light pulse, and R(t) the probability density of the single-electron current-pulse transit time of PMT1. The probability density of the corresponding anode-current pulse is then given by the convolution

$$L^*(t) = L(t) * R(t)$$

If L(t) and R(t) are both gaussian with variances σ_L^2 and σ_{tt}^2, then $L^*(t)$ is also gaussian and has a variance

$$\sigma_L^{*2} = \sigma_L^2 + \sigma_{tt}^2$$

For a fast-response tube $\sigma_{tt} \leq 0.4$ ns. If this is small in comparison with σ_L, then $L^*(t)$ is practically identical to L(t).

The time reference in the set-up of Fig.5.42 is derived from a second photomultiplier PMT2 which also has transit-time variations. To take this into account, let $\bar{n}_{k,i}$ be the mean number of photoelectrons per pulse at the PMT2 cathode. Then, if R(t) for PMT2 is gaussian, the probability density $(L')^*(t)$ of the corresponding anode current pulse is also gaussian and has a variance

$$[(\sigma_L')^*]^2 = \frac{\sigma_L^2 + \sigma_{tt}^2}{\bar{n}_{k,i}}$$

which is $\bar{n}_{k,i}$ times smaller than the corresponding variance for PMT1.

The probability density distribution $L^{**}(t)$ of the histogram constructed by the multichannel analyser is given by the convolution.

$$L^{**} = (L')^{*}(t) * L(t) * R(t)$$

which, provided its standard deviation is large compared with $(\sigma_L')^{*}$ and σ_{tt}, is practically identical to the illumination function L(t).

Thus, this method is a good way for reconstructing brief, low-intensity light pulses provided the transit time fluctuations in the two photomultipliers are low compared with the light pulse duration.

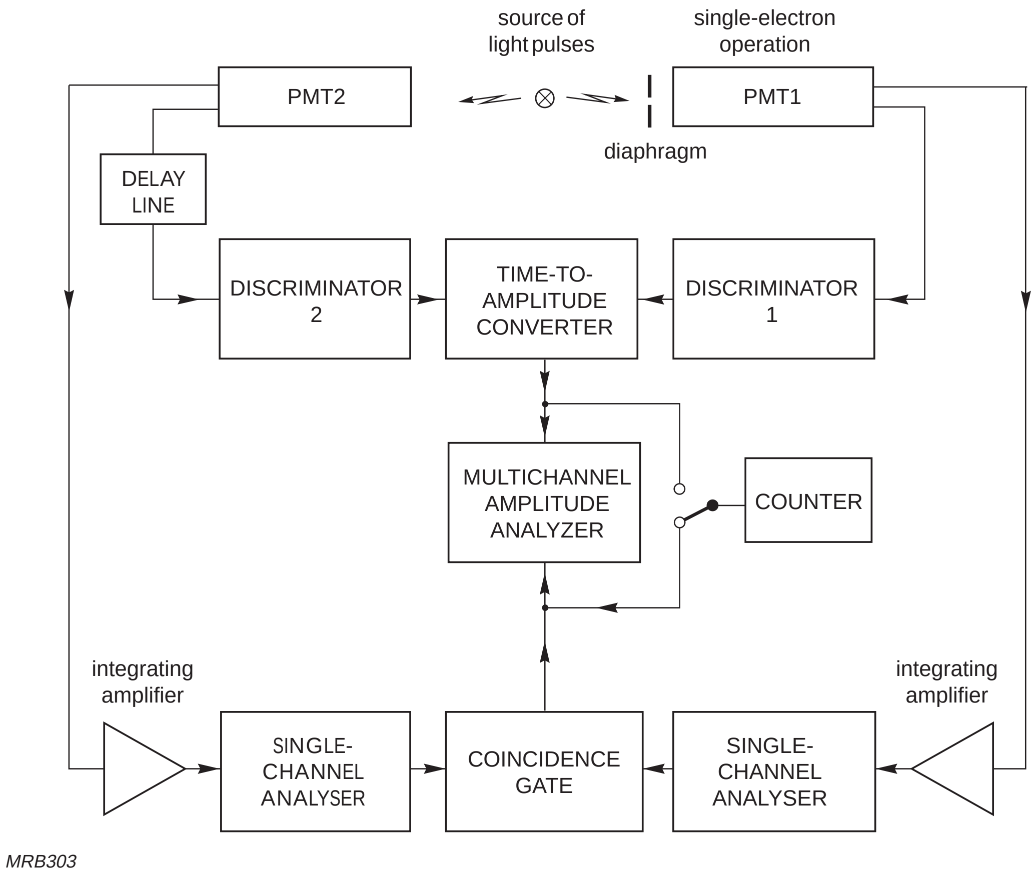

Fig.5.42 Set-up for reconstructing fast, low-intensity pulses by sampling