TOELESS PULSE SHAPING WITH A SINGLE DELAY-LINE NETWORK*

K. Abe, L. Tauhata and D. C. Binns
Centro Brasileiro de Pesquisas Físicas
Rio de Janeiro, Brazil

(Received 3rd June 1975)

ABSTRACT

New unipolar delay-line clippers producing negligible cancellation remnant have been developed. Near perfect clipping is achieved using a combination of several types of coaxial cable transformers working as a phase inverter, a new pulse adder, or an impedance transformer. Only passive elements are used in the bridge network. The construction is simple and the performance is extremely stable and wide in dynamic range and frequency band width. Completely symmetrical bipolar pulses are also easily obtained using this technique.

---

* Submitted for publication to Nucl. Instr. & Meths.

1. INTRODUCTION

Many papers discussing requirements for nuclear pulse shaping and filters to compatible with high counting rate and good energy resolution have been reported. The upper limit of permissible counting rate is set mainly by the pulse pile-up that occurs before the first differentiator. The recovery time of the shaped waveform to the baseline is very important in the reduction of pulse pile-up which depends on both the pulse width (peak pile-up) and the base-line restoration (tail pile-up).

The linear lumped element network proposed by Blankenship and Nowlin[1-2] has good performance in terms of both signal to noise ratio and pulse duration. The pulse waveform is nearly an equilateral triangle and its trailing edge shows a perfect monotonic return to the baseline without undershoot. The noise figure for unipolar pulses is comparable to that of a single delay-line and RC integrator filter. In detail, the optimum unipolar noise factor is 1.135 for the lumped element filter network and 1.098 for the single delay-line and RC-filter[3-4], compared with 1.00 for the cusp response filter[5].

However, with regard to the permissible counting rate , the single delay-line clipper is still superior[6] to RC differentiators, including the above network, if no usual undershoot nor cancellation remnant is to follow the clipped pulse. This is especially true when the clipper is coupled to a fast linear gate and stretcher system. In a pulsed beam accelerator experiment, for example, only relatively large pulses are picked up and analyzed precisely over small but rush background pulses with such a system. A flat top pulse shape is usually prefered there.

In this paper simple forms of single delay-line clippers will be described which are exceedingly good in terms of signal to noise ratio, pulse duration, flat top, counting-rate independent baseline, stability, linearity and wide band of frequency response. We restrict ourselves to the case that the rise time of the signal is always sufficiently short compared with the resultant pulse width.

2. GENERAL

a. Undershoot

The undershoot following single delay-line-clipped pulses are generally due to two causes; the first is overcancellation ocurring when both the attenuation of pulse amplitude along the delay line and the time constant of exponential decay of the original pulses are small in scale, relative to the clipped width (in Fig. 1, the line A_W is above D).

In this case a simple clipping, as indicated in Fig. 2, can be adopted and the sweeping-out of undershoot is easily completed by reducing the amplitude of delayed pulses by a factor , $\exp(-T_D/\tau)$, including the proper attenuation along the delay line in order to get coincidence of the two lines, where T_D and τ stand for the delay time (i.e. afterwards, pulse width) and the decay time constant of original pulses, respectively. The reduction is obtained simply by insertion of finite resistors at one and/or both ends of the delay-line.

The seventh column in Table 1 shows time constants of amplitude attenuation vs. delay time. Although we can get only a rough estimation of the conditions to match both lines, A_W and D in Fig.1, without knowledge of the pulse waveform, we can understand from the time constants that the overcancellation is limited to appear only in the case of the input pulses of relatively fast decay, of a few microseconds or shorter in time constant, unless inconveniently thick cables are used. Therefore, if perfection is necessary in the cancellation, the simple clipping is limited in use only to the cases where the appreciable tilt of pulses is not idsturbing.

The second source of undershoot arises from baseline shift depending on counting rate, that appears in amplifier systems employing interstage capacitive coupling. This baseline shift is eliminated in totally direct coupled systems[8] or by employing DC restorers[9], the latter suffering increase of system noise under some conditions[8].

That kind of baseline shift is, however, another problem to the pulse shaping discussed here. This effect is ignored from now on.

b. Rear Pedestal

The finite "toe"[7], i.e. rear pedestal or cancellation remnant following the clipped pulses due to under-cancellation, appears when both the decay time constant of input pulses and the amplitude attenuation of the pulses travelling along the delay line are large in scale, relative to the clipped width (the line D is above A_s in Fig. 1). This case is common to fast electronics where helically wound delay lines cannot be used owing to their relatively narrow band-width.

(i). The first solution to eliminate the toe is by shortening the decay time constant of input pulses through an RC differentiation to get the factor $\exp(-T_D/\tau)$ large enough to reduce the trailing edge back to complete zero. The differentiation, however, often produces no longer desirable flat-top pulses[10,33], even though the original pulses have slow decay and fast rise times. The situation is similar to the former discussion.

(ii). Another way is tried to eliminate such a trouble as above, where the delayed pulse is subtracted from the input pulse in a difference amplifier[10] or a phase inverter and summing amplifier bridge[11] as shown in Fig. 3, the two being equivalent in the end. The circuitry suitably fits both lines A_s and D in Fig. 1, without resorting to additional differentiation, so that the output pulse will retain its flat top. Further advantage is that the double matching of impedance at both ends enables better shaping than the simple clipping, practically accompanying no small ringing often parasitic to delay-line network. Thin cables with any degree of attenuation coefficient can be employed as long as the band width is appropriate, the relevant data being shown in the fifth columm in Table 1.

The method itself is well suited for clipping either fast or slow pulses of any decay time constant. A small problem is that the delay-line bridge needs twice as long a cable as compared with the simple clipping, since a thin cable can now be used. This technique, however, involves the use of an active element, i.e. a difference amplifier or a summing amplifier, which strongly restricts the performance of pulse shaping due to its finite common mode rejection or non-linearity in inverting and/or summing, dynamic range, response time and additional noises, although several kinds of wideband cables are commercially available and easily utilizable for this clipping.

(iii). Another solution of subtracting a suitable fraction of the unclipped signal from the clipped one with a difference amplifier has been proposed[12]. The shaped signal retains its flat top; however, the circuity also includes active elements and is further complicated[10].

(iv). The DDL (double delay line) clipping[13] is also well known for its effectiveness. The baseline recovery is good and both pileup overloading and low-frequency noise can be minimized.

However, the pulse duration with respect to the peak pileup results in twice as long the corresponding single-clipped pulses. A noise disadvantage[10] is as follows; when λ stands for the ratio of the pulse width to the integration time constant of a single RC integrator, the noise factor is respectively 1.075 or ≥ 1.37 for $\lambda = 0$ or ≥ 1 [4,14].

The higher noise factor is reduced to the lower one of the single delay-line clipping by the use of a partially gated filter[11], the circuitry being complicated.

The toe is strongly reduced, but, strictly speaking, still remains as long as the bipolar peaks are not perfectly symmetrical due to the attenuation of the first clipped pulses while travelling along the second delay line.

(v). In the case where the time constant of exponential decay of input pulses and the amplitude attenuation of the pulses travelling along the delay line are not too small nor too large, at least either of the two extremes is applicable, since the perfect cancellation after the time T_D is exactly equivalent to the perfect accordance of lines A_W and D or A_S and D in Fig. 1 and, in a final analysis, to the pole-zero cancellation[18] in the comprehensive transfer function.

c. A Choice of Development

Among the various types of proposals that have been made so far we find that the mode shown in Fig. 3 would be freer of limitations, flexible and perfect for the pulse clipping, dragging neither toe nor undershoot, if only the active elements in the bridge could be replaced by passive ones. Now a way was found. The problem is how to obtain the phase inverter and summing circuit suitable for the purpose using only passive elements.

The phase inverter is nowadays easily obtained with only passive elements, that is, twisted-wires or coaxial-cable transformers using ferrite toroids[19-22], which are improvements of the prototypes[23-26,15].

The summing circuit requires a high-fidelity operation over a wide band of frequencies, from extremely low ones to higher ones. A transmission-line transformer with an extra winding of a single wire on a ferrite toroid was developed for that purpose, and found to work very well.

A variable attenuator and a resistor potentiometer held together there serve to clear out the toe or undershoot by balancing the normal and inverted signals in a delay-line clipping bridge.

Deterioration of the phase inversion in low frequency or on long pulses is well known[15]. One channel of the new summing circuit also suffers a similar deterioration. However, weak points of that type can be naturally compensated and result in no effect as to the clipping operation, when they are absorbed into a change of the input signal waveform or set common to both the normal and inverted signals.

3. OPERATIONAL ELEMENTS

a. Phase Inverter

The phase inverter, employing a length of coaxial line and a ferrite core, has been fully discussed elsewhere[19-22,27]. The inverter proposed by I. Hayashi[28], which is similar to, but larger than, that described in ref. 29, is employed here; only a brief description will be given for the sake of completeness.

The inverter, shown in Figs. 4 and 8, is formed with a fine coaxial cable of 75 cm in length, wound 15 turns on a closed ferrite core. The two ends of the cable must be separated as far as possible and fixed with a binding agent, e.g. a type of epoxy resin. The coaxial cable used is polyethylene-filled and practically identical to a cable 1.5D-2N, which is bound with a thin cloth ja - cket of nylon in place of usual solid vinyl to reduce its external diameter as much as possible, however, the electrical characteristics are the same as those of the cable 1.5D-2V listed in Table 1. The characteristic impedance is 50Ω. Inner and outer diameters of the ferrite toroid are 16 mm and 28 mm, respectively; the thickness is 13 mm. The permeability μ_0 is about 3000. The rise time is less than 1 nsec and the decay time is about 50 μsec for a step-function pulse applied with matched impedance at both ends. The inductance of the cable sheath, i.e. outer conductor, wound on the toroid is estimated of its input impedance with lowering frequencies (See 3b(i), 4a(i)). The transit time is about 4 nsec. The attenuation is negligibly small for this length, as estimated from the sixth column of Table 1. The rise time is also consistent with the results of the fifth column and the table caption. This inverter can reliably transfer any type of pulse (excluding DC), at least in principle, apart from the disturbance given to relevant external circuits. The sheath terminal must be a little distant from the ground chassis (See 4a(vi)).

b. Transmission-Line Pulse Adder

The new summing network with integrated attenuators was formed with a fine coaxial cable, a single copper wire and a toroidal ferrite core, as shown in Figs. 5 and 8. The cable and ferrite toroid are the same as those of the phase inverter mentioned above. The cable is 60 cm in length, wound tightly 13 turns and fixed with a binding agent. The single wire of the same length as that of the cable is then wound tightly the same number of turns in a free gap on the common toroid, fixed with a small separation between the ends (a) and (d), and connected to the cable sheath as indicated in the figures. As a rule the resistance of the wire should be the same as that of the cable sheath. (The wire is of 0.5 mm in diameter and about 0.06Ω in this case).

However, considerable discrepancy in resistance adds only a small effect to the final result, so that the equality is primarily for the sake of easy discussion (See 3b(iv), 4d). The variable resistor VR is used with the resistor r for both matched termination of the cable and appropriate attenuation of the amplitude of input 1.

The resistors R_1 and R_2 of the same value are inserted with the aim to keep the impedance of point (c) low a wide range of frequencies (See 3b(ii), 4a(v)), and to make sure of an attenuation factor, i.e., one half the amplitude of input 2. The inductor potentiometer consisting of the cable and single-wire windings also shares a role with the resistor potentiometer. The factor "one half" has an important meaning (See 3b(i), (ii))and is also useful for simplicity of consideration and construction, and is applicable unchangedly up to the length of delay line corresponding to 6 dB of attenuation, e.g., about 1 km of RG-8/U cable or 5 μsec in its delay time, as seen from Table 1.

Details of the function of this puzzle ring are given stepwisely below.

(i). Channel of The Input 2

The resistance and inductance potentiometers composed of R_1, R_2, L_1 and L_2 are of equal arms over a wide range of frequencies from DC to higher ones. Therefore joint (c) always builds one half the potential of input 2 whatever are the circunstances around input 1, as to connection and applied voltage, as long as the following are valid; (A). A load impedance applied to point (f) is sufficiently large compared with the impedance of joint (c) relative to ground (See 3b(ii)). (B). DC resistance of the cable sheath is negligibly small compared with the characteristic impedance. Condition (A) is often difficult to be fulfilled. However, the bulk of relevant effect is usually a shift of the potentiometer ratio down to the ground. This effect can be cleared out, in principle, inserting a resistor R_L' into the points (f) and (d), apart from the compensation of parasistic capacitances.

It is convenient and sufficient to settle the value of resistor R_L' to be the same as the load impedance R_L so as to form an additional equal-arm potentiometer along with the load. Then point (f) also builds a half potential of input 2 and the load effect disappears.

As a matter of fact, the load impedance owing to the succeeding stage is not constant nor known exactly. There may be some discrepancy between the real impedance and the supposed one. Fortunately, if such a residual effect is small, it is automatically involved and compensated in the final adjustment of the variable attenuator VR, and practically leaves no more problems (See 3b(ii), 4a(i), (v)).

The whole inductance between points (a) and (d) is about 3.8 mH at a low frequency, which brings about a considerable decrease of input impedance of channel 2 as the relevant frequency falls down; therefore it gives an additional tilt to a response for a step-like input signal, the situation being similar to that of the phase inverter. The degree of tilt is dependent on the source impedance prior to the system, as well as on the inside conditions, and is not simple. However, the division factor of the potentiometer "one half" persists at all events for the pulses introduced through input 2. Also there is a way where the impedance variation can be absorbed into the deformation of input pulse waveform and results in no effect on the toe or undershoot (See 2c, 4a(i)).

On the contrary, the slight enhancement of tilt is put to good use in the final network (See 4d).

(ii). Channel of the Input 1

The coaxial cable from point (a) to (b) is short and terminated appropriately with the resistors VR and r in parallel, no inversion occuring.

The cable infinitively long or properly terminated is considered a resistor of its characteristic impedance, as long as the resistances of inner and outer conductors are negligibly small compared with the cable impedance and, in the network, their inductive coupling on the ferrite toroid is perfect. Here those conditions are true. Then, in principle, the short cable is fully reliable for transmission of any type of pulses (excluding DC), (See 4b(i)

(ii) as to its limitation). However, the load impedance gives some trouble here, too. A major effect of the load impedance involved in channel 1 is a reflection of the input pulse due to some degree of newborn mismatch of impedance at the endpoint (b)-(c), when a long cable is set prior to the input 1. An effect coming from stray capacitance is again ignored here.

Joint (c) properly views the ground through the two resistors R_1 and R_2, as well as the two coils L_1 and L_2 composed of the cable sheath and the single wire. The coils are all wound tightly on the common ferrite toroid, the same number of turns, with very small DC resistance and now looked contrariwise to each other from point (c). Then if input 2, point (d), is grounded and the inductive coupling coefficient k between the two coils is unity, joint (c) also looks grounded. In reality, however, the source impedance put to (d) is finite but kept normally below that of the cable. The constant k is still smaller than unity, frequency-and, strictly speaking, current-dependent. Therefore the impedance of (c) is raised by a certain amount. The amount increased does not exceed, at any rate, 0.7 times the value of R_1 or R_2 which is previously set small and comparable with the cable impedance. This is one of the reasons why such an additional R-potentiometer is inserted in parallel with the inductance one. The other reason is to suppress any stray capacitance effect around there.

Thus a slightly larger impedance than a full or a half value of load impedance,according to the presence of the compensation resistor R_L', is effectively connected in parallel across points (f) and (c). It is then necessary to correct the whole resistance of the pair of main resistor VR and supplementary terminator r by a suitable amount.

For that purpose, it is convenient to set VR several tens per cent larger than the cable impedance, and r more or less ten times as large as that, when the load impedance is previously estimated large but not sufficient. Adjusting operations of the slider positions of the two interfere somewhat with each other.

Yet, a correct match of impedance is not expected on a realistic basis, since usually the input impedance of the next stages is not known exactly nor constant. Also a situation around the impedance variation of (c) is similar. Therefore it is always a better policy to connect it to a sufficiently large enough load impedance not to need a compensation resistor (See 4a(iii)). In practice, however, it is shown unnecessary to worry much about exact matching of the impedance. For example, 50% deviation of the resistance R_L' from R_L produces at most 0.5 and 0.03% reflection for R_L of 1k and 10 kΩ, respectively. The corresponding contribution to the final summation operation appears delayed by the same time as the clipping width after the trailing edge of the clipped pulse. The estimate was made only on the basis of a simple circuit theory (See 4a(v), (vi)). Further discussion will be given in 4b(ii) as to the frequency-dependence of the characteristic impedance of cables.

(iii). It is hitherto known that there is no interference between channels 1 and 2, as long as the load impedance is sufficiently large, or its effect is small and finally cleared out (See also 4a(iv)). The factor of potentiometer "one half" persists always as to any pulse coming into channel 2. On the other hand, the pulse into 1 appears across the endpoints (b) and (c) with high fidelity. Then, in principle, we get the following relation without disturbance;

$$V_{out.} = \alpha V_1 + \frac{1}{2} V_2 \qquad (1)$$

V_1, V_2 and V_{out} are the voltages of input channels 1 and 2 and output, respectively. α is a given attenuation factor on the whole of channel 1.

The weighted summation is valid over a wide range of frequency excepting DC. When α is set to 1/2 including all effects of attenuation in channel 1, a simple summation is obtained.

In addition, an inversion in polarity of V_1 or V_2 performs a simple subtraction.

(iv). The present validity down to extremely low frequencies is vital to the construction scheme of complete delay-line clippers that work even for the ideal step-function signal.

For nuclear pulses, the validity down to DC itself is not required, since their decay time constant is usually set to approximately 50μsec or shorter, and the following is well known; in the case of a structure in which the low-frequency response falls down with a slope of 6 db/oct, accompanied by a 90^0 phase advance, 50 μsec of decay time constant of response to a stepfunction signal corresponds to about 3 kHz of low-frequency cutoff in characteristic (See, e.g., ref. 7), and, of course, practically indicates no presence of DC component.

4. SINGLE DELAY-LINE CLIPPERS COMPOSED OF PASSIVE ELEMENTS

a. Construction

Figs. 6-8 show a new construction of single delay-line clippers by means of the components mentioned above. The schemes are identical to that in Fig. 3. Generalities have been described in secs. 2b(ii) and 2c. The following are to supplement them.

(i). Together the phase inverter and channel 2 of the pulse adder distort the input waveform in such a way that the decay time constant is slightly shortened, depending upon the source impedance, since they behave partly like inductors against the preceeding stage. At any rate, however, the instantaneous absolute amplitudes of the incoming and that of the outgoing pulses are always identical at the phase inverter besides the short transit time. Therefore the same pulse amplitudes, in a sense of absolute value, apply at any instant to both the pulse adder and the appropriately terminated delay line in the network.

The pulse adder is also known to be sufficiently reliable to yield a perfect sum pulse made of the normal or inverted input and the corresponding delayed one with opposite polarity relative to the former, in accordance with eq. (1). The sum pulse here means the delay-line-clipped pulse with just one half the amplitude of the input fed to the system. When the variable attenuator is properly adjusted according to the intrinsic attenuation of the delay line used, the resultant has neither toe nor undershoot. "Just one half" is here somewhat symbolic. A slight deviation of the factor due to the variation of potentiometer components is practically absorbed into the final adjustment of VR.

(ii). The proceeding discussion itself is valid for both types of clipper shown in Fig. 6, and we can easily get either the normal clipped pulse or the inverted one dragging neither toe nor undershoot by the use of a switch indicated in Fig. 7. However the input impedance of the system and the termination condition concerning the phase inverter and delay line make a difference to both, then it may result in a slight readjustment of VR. For pulses with slow rise time the situation shows no difference. When the sum value of R_1 and R_2 is set equal to the cable impedance of the phase inverter, hence delay line, the difference also nearly disappears. On the other hand it is not desirable for the double matching of impedance of the delay line. The normal type of clipping serves in all events. Since the phase inverter and pulse adder can be constructed in common or different impedance with thin cables of 50, 75, 93 Ω, etc. commercially available or with several twisted wires of $\geq$ 85 Ω(See, e.g., ref. 7), it is usually possible to match the impedance at all points in either type of clipping. However, each cable, or twisted pair, used for the inverter or adder is shorter than 75 cm, and then perfect matching is usually not required.

As an example, the choice of $R_1 + R_2$ as 110 or 150 Ω gives the best match to the 50 Ω normal clipping system in the case of 93 or 75 Ω source impedance, respectively. The value of 200 Ω for a set $R_1 + R_2$ is practically most tolerable to the 50Ω normal clipping system from the viewpoint of the double impedance matching, since it makes an input impedance of 40 Ω and gives, against the input of the phase inverter and delay line route, an impedance of 63, 55 and 40 Ω according to the output impedance of 93, 75 and 50 Ω of the preceeding stage, respectively.

(iii). When the load impedance R_L is sufficiently large, it is not necessary to insert the compensation resistor R_L'. The supplementary matching resistor r is also unnecessary. By assuming that the inductive coupling coefficient k between L_1 and L_2 is unity, and point (d) is grounded in order to get an overestimation limit of pulse reflection at the worst condition, the undesirable change $-\delta R$ of effective termination resistance due to the connection of load, and the reflection coefficient Γ on the basis of simple circuit theory are estimated as follows,

$$-\delta R \leq R_{cf}^2 / (R_L + R_{cf}) \tag{2}$$

$$\Gamma \leq R_{cf}^2 / (2(R_{bf} + R_{cf})(R_L + R_{cf}) - R_{cf}^2) \tag{3}$$

where R_{bf} and R_{cf} stand for the resistance between the points (b) and (f), and the (c) and (f), respectively. Taking as $R_{bf} + R_{cf} = 50\Omega$ and $R_{cf} / R_{bf} = 7/3$, we get that $\Gamma \leq 1.2$, 0.6, 0.25, 0.12 and 0.06% for R_L = 1k, 2k, 5k, 10k and 20 k Ω, respectively.

The percentage of V_1 contributes to a summation delayed by the same time as the clipping width after the trailing edge of the clipped pulse. This type of contribution cannot be cleared out by readjustment of VR. Thus we find that the condition of sufficiently large resistance means usually $R_L \gtrsim 20\ k\Omega$ for the 50Ω system.

(iv). In principle, some kind of interference from channel 2 to 1 exists, whether the load impedance is sufficiently large or not.

(A). Even though R_L is sufficiently large, an additional sheath current induced by another electromotive force unrelated to the pulse transmission (hence the damping effect is ignored here) may make a discrepancy between the voltage amplitudes of the inner and outer conductors at point (a) and (b) owing to the finite conductance along the cable and imperfect inductive coupling of the windings, i.e. the inner and outer conductors. However in practice it has no effect on the result since the usual nuclear pulse has a decay form looking as if it has been transmitted through a high--pass filter with a low-frequency cutoff of 3 kHz or higher (See 3b(iv)); then extremely low frequency components contributing mainly to such a effect,i.e. those corresponding to the reactance comparable to the DC resistance, are scarcely included in that pulse.

(B). When R_L is not large, a cross talk from channel 2 to 1 appears according to the slider position of the variable attenuator VR. Provided the cable is properly terminated with Z_1 and the voltage induced on the inner conductor is equal to that of the outer one, the new voltage V_1' from channel 2 to point (f) of channel 1 is as follows,

$$|V_1'| \leq \frac{V_2}{2} \frac{R_{cf}(Z_1 + R_{bf})}{R_L(Z_1 + R_{bf} + R_{cf}) + R_{cf}(Z_1 + R_{bf})} \qquad (4)$$

The polarity of V_1' is opposite to V_2.

Let $Z_1 = 50\ \Omega$, $R_{cf} = 35\ \Omega$, $R_{bf} = 15\ \Omega$ and $R_L = 20\ \mathrm{K}\ \Omega$, then V_1' is estimated approximately 0.06% of V_2.

(C). The cross-talk pulses mentioned above appear finally at the input of system in more reduced size, taking exactly the same time as that the original pulse needs for arriving at the point (f) through the same route, i.e. the cables of phase inverter, delay line and pulse adder.

The timing of the cross talk participating to the summation coincides exactly with that of the proper delayed pulse. Thus such type of cross talks are buried in the final adjustment of the amplitude of delayed pulse. A slight deformation of the cross talk pulses, suffering anew by inductance of the inverter and adder, contributes much less.

(D). Of course, introducing the R_L' appropriately clears out the cross talk.

(v). In the case that $R_L \leq 20\ \mathrm{k}\ \Omega$, furnishing with a supplementary variable resistor r is usually preferable for correct termination of the cable, hence R_L', too. In that case, reflection in channel 1 caused by an inappropriate R_L' comes out.

Putting primarily $R = R_1 = R_2$, $\Delta R = R_L' - R_L$, $k = 0$, L_1 and L_2 are very large, then the effective load R_L'' connected across the points (f) and (c) is estimated as follows,

$$R_L'' = \frac{1}{2} W_2 + \frac{\Delta R}{2} \frac{1}{2 + \Delta R\, W_1 / W_2} \qquad (5)$$

Where $W_1 = Z_2 + 2R$, $W_2 = R + R_L$, $W_3 = Z_2R + W_1R_L$ and Z_2 is the impedance viewed from point (d) looking back towards input 2. Assuming $2R = 200\ \Omega$, $Z_2 = \pm 50\ \Omega$, all cables are of 50 Ω and the amount of scatter of R_L' from R_L is 50% of R_L, then the second term equals to approximately $10^{-1}R_L$ for $R_L \gg 20$. 50% deviation of R_L' results in less than 20% scatter of R_L''. Further putting $R_{cf} + R_{bf} = 60\ \Omega$ and $R_{cf}/R_{bf} = 7/3$ in addition, then the reflection coefficient Γ induced by 50% inequality of R_L' is obtained, namely, Γ = 0.5, 0.3, 0.1 and 0.03% for R_L = 1k, 2k, 5k and 10 k Ω, respectively. The amount decreases with better choice of R_L'. A fraction of V_1 slightly deformed and limited at most by the above factor appears at the input of system and causes dragging a delayed toe or undershoot.

This kind of effect cannot be cleared out without an appropriate choice of R_L' (See 4a(iii)) or readjustment of r.

The primary assumption on L_1, L_2 and k is in fact partly not true, especially in lower frequencies, but any fluctuation of their values results in at most 60 Ω difference on R_L''. The contribution to the reflection is limited respectively within about 50, 25, 10 or 5% of each reflection coefficient Γ mentioned above in the case of 50% inequality of R_L'. The contribution decreases rapidly with increase of R_L'. Although this kind of effect can be reduced with a lower value of R, the use of R_L over 2 k Ω is usually agreeable with the above parameters.

(vi). The above treatment on pulse reflection based only on a simple circuit theory represents an oversimplification.

Leaving aside the fabrication accuracy of the cable,there are still many causes of microscopic pulse reflection, such as : (A) the use of rumped resistor across the cable, (B) stray or equivalent capacitance (See the following section) around the terminator or junction, (C) simple joint of the delay line and the cables of phase inverter or pulse adder of very different diameters, (D) the finite damping of energy travelling outside the coaxial sheath in the coil mode, etc. Such particulars involving higher mode propagation have been given in, e.g., refs. 15, 20 and 30.

b. Suitable Cables

Strictly speaking, all the above discussions are not always true as far as the fidelity of transmission line and constancy of its impedance with a frequency, from extremely low to higher , are concerned.

(i). Fidelity of Transmission Cables

Phase-frequency distortion and amplitude distortion due to frequency-dependent attenuation, are the largest limiting factors at the usual nuclear pulse clipping from 50 nsec to 1 μsec in width.

For the construction of the phase inverter and pulse adder only a few cables are commercially available, but twisted wires can also be used. Their necessary length is short and then they easily fulfill the usual requirements over the frequency region of interest for the processing of nuclear pulses, as already mentioned above. Practically they do not deteriorate the rise time of the re sultant pulse. On the other hand, the delay line cable used to define the clipping width is usually very long; therefore the distor tion of pulse waveform suffered during the passage violates to a nonnegligible extent the good similarity expected in waveform between the normal and delayed pulses put into the pulse adder. The cable effects cannot be compensated completely by ordinary networks and lead to a short drag of deformed toe or undershoot. An amount of damping itself has no interest in this clipping network. Utilization of thin cables isn't kept at a distance any more because of its loss in contrast to former times. Small-sized construction of the delay line unit enables good shielding against external noises. Rise time response or band width of frequency is now one of the most important measures in the cable selection from the viewpoint of sharp-cut trailing.

A uniform rectilinear line shows such a peculiar step-function response, as a relatively rapid initial rise is followed by a very slow approach to the final amplitude of the pulse[7,15]. The relevant description has been given briefly in the caption of Table 1. Notice its dependence on the length of cable used. In practice the rise time is to be taken as approximately twice the specified value, due to the cable deficiencies and the equivalent capacitance lumps associated with entering the cables[20,30,35]. Details on transient analysis of coaxial cables considering skin effect have been given in refs. 16 and 17.

The utilization of a distortionless cable mentioned below solves the whole problem, since the attenuation and velocity do not increase with frequency and are still constant for all frequencies.

(ii). Characteristic Impedance

The general definition of the characteristic impedance of cable is

$$Z_o = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \tag{6}$$

where R, L, C and G = total series resistance, inductance, shunt capacitance and leakage conductance of line per unit length, respectively.

At sufficiently high frequencies (e.g., for RG-58 A/U or 1.5 D-2N, f >> 3.5 kHz or 120 kHz, respectively), the characteristic impedance is a constant pure resistance equal to $\sqrt{L/C}$ ohms. At low frequencies where R is comparable with ωL and G comparable with ωC, the characteristic impedance varies with frequency, and approaches $\sqrt{R/G}$ as the frequency approaches zero. On the other hand, only if RC = GL, the characteristic impedance is independent of frequency and always the real constant $\sqrt{L/C}$ (See, e.g., refs. 34-36), which is the condition for a distortionless cable.

A rectilinear cable for which G, but not R, is negligibly small is commonly available. In this case, the presence of conductor loss has the obvious effect of attenuating the signal, as mentioned in the above section. Another and less obvious effect of the conductor resistance, is that Z_o is no longer a real quantity in lower frequencies of nuclear interest;

$$Z_0 = \sqrt{\frac{L}{C}\left(1 - j\,\frac{R}{\omega L}\right)} \qquad (7)$$

The form of this expression shows that the impedance of the infinite cable has a capacitively reactive component[35]. Thus, for a sending-end generator of step-function pulse, the usual cable of finite length, properly terminated with a resistor, behaves as a time-variable load impedance which increases gradually but clearly from the so-called characteristic impedance (i.e. the limiting value in higher frequencies) to higher during twice the one-way transit time (N.B. the usual matching of impedance using a resistor fails in low frequencies due to the frequency-dependence of characteristic impedance). When the source impedance is finite, mainly the delay line used for definition of clipping width distorts the common input waveform, for example, of a simple step function to a folded curve consisting of an initial intrinsic rise, relatively long slow slope and succeeding nearly flat top. In the clipping , it gives rise to a limited drag of toe or undershoot with a small bump at its end, in addition to the degradation of sharp-cut trail caused by the peculiar rise behaviour of transmitted signal. The dragging time is, in this method, very nearly the same as the pulse clipping width. It is impossible to eliminate the toe or undershoot with an ordinary network due to such a complicatedly deformed input. Some reduction, however, can be attained by insertion of, e.g., an inductance-resistance shunt across its input or by use of a preceeding amplifier with source impedance as low as possibly allowed by the frequency-dependent input impedance of the delay--line clippers with sacrifice of the top flatness of the clipped pulse or two-way matching of impedance, respectively.

Although, in general, the use of lower attenuation cable gives less troubles of that kind, if the distortionless cable[34-36] is available, the problem entirely disappears. The distortionless cable could be made by using poorer insulation and thus increasing G, but only at the expense of large, if uniform, attenuation[35]. An amount of damping itself, however, has no interest in this clipping network. Thus the distortionless cable is much preferable.

(iii). Use of Impedance Transformers

In the case of the use of helical line the termination practice requires some consideration such as flaring of the winding[15]. In addition, insertion of impedance transformers on both sides of the delay line is preferable, since it normally has considerably higher impedance. However, only the alteration of terminating resistors VR and r, without the use of one or two of the impedance transformers, is often tolerable for the processing of slow pulses, which must be experimentally found. Impedance transformers using a thin cable, twisted pair or multifilar-winding and a ferrite toroid have been described elsewhere[15,20-22,24,26,28,29,31].

c. Ferrite Toroids

The larger the core permeability, the fewer turns are required for a given low frequency response and the larger the over all band width[21]. The equivalent shunting resistance and inductance values in energy travelling in the coil mode are approximately proportional to the square of the turns and the core cross-sectional area. The inductance is inversely proportional to the mean core circumference. For this reason, a toroidal core of small over-all size should be used when possible, limited by either window size or saturation of the core[20]. The usual maximum μQ_0 product criterion is not valid in this case. Instead a high value of μ_0 and a large value μ/ε (the ratio of high frequency pearmeability to high frequency dieletric constant) are desirable for low frequency response and keeping the equivalent shunt resistance high[20].

The above details have no direct effect on drag of the toe or undershoot, nor on the pulse height of clipped pulses but a slight effect on variation of deformation of the input waveform, hence, waveform of the clipped pulse in vicinity of its top. However, as to the low frequency response, it has been reported that the decay time constant is not single but implies an increase of effective permeability with time[20]. The effect may cause a toe by a reason similar to the time-variable load impedance mentioned in the above section (ii).

d. Decay Enhancement by the System

This delay line clipper transforms the input step-function form to a waveform with a tilt. The effect slightly deteriorates the flatness of the pulse top. On the other hand the effect due to the break of similarity between the normal and delayed pulses often arising in low-frequency regions is diminished and the possible slight inequality of the potentiometer arm of the resistors R_1 and R_2 or DC resistance of the inductances L_1 and L_2, results in no effect against settling of the baseline, since the frequency components in connection with the above are scarcely included.

By the way, when we adjust the clipper system best, using a pulse generator with decay time constant of 50 μsec, the toe or undershoot appearing due to the deviation of a given intrinsic decay time constant T_1 (μsec) from 50 μsec is estimated to be within Δ%;

$$\Delta \leq 100 \left| \exp \left(\frac{1}{50} - \frac{1}{T_1}\right) T_D - 1 \right| \tag{8}$$

or,

$$\Delta \leq 2T_D(\mu sec) \qquad \text{for } T_1 = 25 \ \mu sec \sim \infty \tag{9}$$

Then, for example, in the case of 150 nsec clipping, it is unnecessary to readjust the system over a range from T_1=25 usec to infinity when 0.3% of toe or undershoot is tolerable.

5. CONSTRUCTION OF A DOUBLE DELAY-LINE CLIPPER

An ordinary combination[13)] of two simple delay line clippers and an interstage buffer amplifier between them cannot make a perfectly symmetric bipolar pulse, nor acheive complete removal the cancellation remnant because of the cancellation remnant of the first clipping, and/or the amplitude damping along the second clipping cable. On the other hand, the replacement of any one of the two simple delay-line clippers by the new clipper, brings about nearly perfect restoration of the baseline, and the replacement of the two makes a perfect bipolar pulse in the sense of its pair symmetry and baseline restoration.

A different type of delay-line clipper supplying bipolar pulses using only one delay line[32] can also be transformed into passive element construction. For that purpose it is sufficient to raise up the resistance of the terminators VR and r to very high values in the new pulse adder, then the new clipping network becomes equivalent to the whole circuitry already proposed.

The cable effects (See 4b (i)(ii)) cannot be compensated by the bipolar pulse method.

6. EXPERIMENTAL ARRANGEMENT

Although many kinds of precise measurements are to be exhausted for completion, the construction of networks with such totally passive elements as the mentioned above guarantees beforehand the good stability, noise figure, rise time, linearity and dynamic range of the delay line clipper; the fast part of the pulse or rise time is not affected by the long delay line (See 3a, 4b(i)) or the characteristics of the ferrite core (See 4c); the linearity and dynamic range are disturbed only indirectly through the decrease of inductance of the transmission-line transformers for the slow (flat) part or low frequencies, if saturation of the ferrite core occurs due to abnormally high and long relevant signals. If not, or in usual nuclear instrumentation, they are expected to be extremely good (See 3a, 4c and refs. 20, 21 and 28). As to the trail of the clipped pulse, the utility of this technique and the propriety of the discussion of the complete cancellation are demonstrated here with photographs.

A circuit diagram of the single delay-line clipper is shown in Figs. 5-7. The first circuit was designed specially for nuclear reaction experiments with a short burst beam of about 2 μsec in durations and for this reason a 100-nsec clipping time was chosen and a fast amplifier of high input impedance with small capacitance, consisting of a field effect transistor (Toshiba, 2SK19-BL) and an integrated circuit (NEC, μPC 103A) was equipped behind the direct output (f) as the cable driver stage to send the pulse a distance of 30 m. An inductance-resistence shunt was also attached across the input in order to reform the distortion of the input waveform mentioned in sec. 4b(ii). In addition, a variable resistor of about 22 k Ω was inserted in parallel with R_2 to help, in practice, the fine adjustment of VR. Other parameters of the components used are; R_1 and R_2: 102.5 Ω, r: 370 Ω(at maximum), VR: 72 Ω and DL: RG-58A/U of 20 m.

The whole circuit built in a NIM standard bin is shown in Fig. 8, where SW, Adj and VR_2 are the polarity switch, variable resistor of maximum 500 Ω forming a part of the inductance-resistance shunt and variable resistor in parallel with R_2, respectively. The inductance was of about 40 μH. The transmission-line transformer used as the phase inverter and pulse adder are also indicated in the photograph.

Fig. 9 shows typical input and output waveforms observed with an oscilloscope (Tektronix 7503; the input impedance of its probe is 1 MΩ with capacitance of less than 2 pF in parallel). The vertical sensitivity (mV/div) and horizontal sweep rate (nsec or μsec/div) are displayed in each photograph. A pulse generator with output impedance of 50 Ω was also used.

Photograph 1 demonstrates that the folded deformation of input waveform due to the frequency-dependence of cable impedance was well reduced by the inductance-resistance shunt. The first bump around 100 nsec in time lapse is due to the cross talk of the fast part of the signal between channels 1 and 2 through the effect of the stray capacitance at the direct output (See 4a(iv); the discussion is applicable to the capacitive load). The second bump around 200 nsec is due to both the reflection of the fast part of the signal by the stray capacitance at the direct output (See 3a (i), (ii)) and the trace still remaining even after the reformation of the input waveform (See 4b(ii)). It is necessary that the load capacitance at the direct output be kept to an absolute minimum (usually, $<$ 5 pF), if such a reflection is to be ineffective in the pulse height analysis. The variable resistor, r, is to be adjusted on spot because of the deviation from the nominal value of the so-called characteristic impedance of delay cable and its slight interference with the slider position of VR (See 3b(ii)).

Photographs 2 and 3 show examples of under-cancellation resulting in toe and over-cancellation resulting in undershoot according to the excess and shortage of attenuation of the delayed signal, respectively. Photographs 4 and 5 show the best balance that exists halfway between them.

Photograph 6 shows the other example obtained with use of thinner delay cable 1.5D-9EN of about 90 m, when the shunt inductance was changed to 82μH. The leading spike is due to a characteristic of the amplifier.

To obtain a precise balance, a set of, e.g., double-step pulse generator, fast linear gate, pulse stretcher, multichannel pulse height analyzer, etc. should be used.

7. CONCLUSIONS

It is usually required at the direct output of the single delay-line clipper that $R_L \gtrsim 2\ k\ \Omega$, stray capacitance $\lesssim 5$ pF and R_L' is to be inserted for $R_L \lesssim 20$ k . The selection and adjustment of the circuit parameters are not critical.

The utility of the new pulse adder is that it easily performs the perfect single or double delay line clipping.

However, the finite differential nonlinearity or limited dynamic range of the preceeding stage prior to the delay-line clipper, i.e. usually a preamplifier, may hurt the ability to some extent, due to the pile-up possibly existing there. Those clippers are especially expected to give full play in such a field as nuclear reaction experiments using a pulsed beam accelerator. The development is to form a part the preparation program of photonuclear reaction experiments using a solid state detector system and an ordinary electron linear accelerator.

ACKNOWLEDGEMENTS

The authors would like to gratefully acknowledge the continuing encouragement and advice of Prof. Alfredo Marques de Oliveira. One of the authors (K.A.) wishes to thank the great hospitality during his stay on leave from Tohoku University, Japan, he also wishes to thank Comissão Nacional de Energia Nuclear for their financial support. Thanks are due R. Poledna for his technical help.

REFERENCES

1) J.L. Blankenship and C.H. Nowlin , IEEE Trans. Nucl. Sci. NS-13 nº 3 (1966) 495.

2) C.H. Nowlin , IEEE Trans. Nucl. Sci. NS-17 (1970) 226.

3) M. Tsukuda, Nucl. Instr. and Meth. 14 (1962) 241.

4) C. H. Nowlin, J.L. Blankenship and T.V. Blalock, Rev. Sci. Instr. 36 (1965) 1063.

5) R. Wilson, Phil. Mag. 41 (1950) 66.

6) G. Amsel, R. Bosshard and C. Zajde, IEEE Trans. Nucl. Sci.NS-14 nº 1 (1967) 1.

7) R. Littauer, Pulse Electronics,(1965)67, 80-83 (McGraw-Hill Book Co., Inc. New York).

8) L.O. Johnson and R.L. Heath, IEEE Trans. Nucl. Sci. NS-17 nº 1 (1970) 276.

9) R.L. Chase and L.R. Poulo, IEEE Trans. Nucl. Sci. NS-14 nº 1 (1967) 83.

10) R.L. Chase, Nuclear Pulse Spectrometry (1961) p. 30 (McGraw-Hill Book Co., Inc. New York).

11) T.V. Blalock, Rev. Sci. Instr. 36 (1965) 1448.

12) H. Koch and R. Johnston (eds.), Multichannel Pulse Height Analysers (1957) p. 21 (Natl. Acad. Sci. - Natl. Research Council Publ. 467, cited in ref. 10).

13) E. Fairstein, Rev. Sci. Instr. 27 (1956) 475.

14) M. Tsukuda, Hōshasen-keisoku-no Erekutoronikusu (1969)sec. 2-4-3 (Iwanami-Shoten, Tokyo).

15) I.A.D. Levis and F.H. Wells, Millimicrosecond Pulse Techniques (1954 and 1959) sec. 2,3 (Pergamon Press Ltd. N.Y.).

16) R.L. Wigington and N.S. Nahman, Proc. IRE. 45 (1957) 166.

17) G.L. Turin, Proc. IRE. 45 (1957) 878.

18) C.H. Nowlin and J.L. Blankenship, Rev.Sci.Instr. 36(1965)1830.

19) A.I. Talkin and J.V. Cuneo, Rev.Sci. Instr. 28 (1957) 808.

20) C.N. Winningstad, IRE Trans. Nucl. Sci. NS-6 nº 1 (1959) 26.

21) C.L. Ruthroff, Proc. IRE. 47 (1959) 1337.

22) S. Gorodetzky, A. Muser, J. Zen and R. Armbruster, Nucl. Instr. and Meth. 17 (1962) 353.

23) R.W. Rochelle, Rev. Sci. Instr. 23 (1952) 298.

24) I.A.D. Lewis, Rev. Sci. Instr. 23 (1952) 769.

25) H.G. Rudenberg, The Distributed Transformer (April, 1952).(Research Div. Raytheon Mfg. Co., Walthan, Mass. USA).

26) I.A.D. Lewis, Eletronic Engineering 27 (1955) 448.

27) J.-C. Balland, R. Gaillard, J. Pignert and J.-J. Samueli, Nucl. Instr. and Meth. 45 (1966) 275.

28) I. Hayashi (ed.) Genshikaku-Jikken-Erekutoronikusu (1963) p.129--155 (Maki-Shoten, Tokyo). Genshikaku-Kenkyujo-Hōkoku (1961). JNS-TL-52 (Institute for Nuclear Study, University of Tokyo, Tokyo).

29) I.Hayashi and N. Yoshimura, Nuclear Electronics (1962) p. 157 (Proc.Conf.Nucl.Electronics.IAEA, Vienna).

30) J.R. Whinnery, H.W. Jamieson,T.E. Robbins,Proc.IRE.32(1944)695.

31) S. Tansal and H. Sobol, Rev. Sci. Instr. 34 (1963) 1075.

32) S. Gorni, Nucl. Instr. and Meth. 47 (1967) 74.

33) E. Fairstein and J. Hahn,Nucleonics 23 Nov (1965) 50.

34) F.E.Terman, Radio Engineers'Handbook (1943) sec.10 (McGraw-Hill Book Co., Inc. New York).

35) Radio Engineers of the Westinghouse Electric Corporation, Industrial Electronics Reference Book (1948) p. 292-293 (John Wiley

and Sons, Inc. New York).

36) R.W.P. King, *Transmission-line Theory* (1955) p. 93-101 (McGraw--Hill Book Co., Inc. New York).

FIG. 1 - Relation between an input signal decay with time lapse and an amplitude attenuation when the pulse propagates along the delay line, at a given time represented by the T_D unit scale.

A_w or A_s: weak or strong peak amplitude attenuation of delayed signal, respectively. D: decay lapse of input amplitude. a: instantaneous amplitude. d_w or d_s: delay time corresponding to the peak amplitude attenuation by a factor e^{-1} for A_w or A_s, respectively. τ: decay time constant of input signal. T_D: clipped pulse width.

FIG. 2 - Simple delay-line clipper (short circuit type[7]).

FIG. 3 - Delay-line clipping bridge[11].

FIG. 4 - Phase inverter.

a and a': inner conductor. b and b': outer conductor. 1.5D-9EN: (Fujikura-Densen Co., Ltd. 1-5-1, Kiba, Tokyo 135); which is replaceable by 1.5D-2N (Dainichi-Nippon--Densen Co., Ltd. 7-3, Umeda, Osaka 530). H_{5A}-T_{16}:Tokyo--Denki-Kagaku-Kogyo Co., Ltd. 2-14-16, Uchikanda, Tokyo 101).

FIG. 5 - Transmission-line adder.

FIG. 6 - Schemes of single delay-line clippers.
(a): normal. (b): with polarity inversion.

FIG. 7 - Single-delay line clipper with a polarity switch.

FIG. 8 - Photographs of a single-delay line clipper with an amplifier.

FIG. 9 - Pulse photographs (1-5: using 20 m of RG-58A/U cable, pulse width ∿ 100 nsec. 6: using about 90 m of 1.5D-9EN cable, pulse width ∿ 450 nsec).

1: input waveform. 2 and 3: 'direct output' waveforms obtained by wrong adjustment of VR resulting in toe or undershoot. 4 and 5: 'direct output' waveforms at the best adjustment of VR. 6: 'amp. output' waveform with the best adjustment.

TABLE 1 PROPERTIES OF SOME COMMERCIAL CABLES

Helical Lines										
Type***	External diameter (mm)	Characteristic impedance (Ω)	Delay time (nsec/m)	Risetime (10-90 %) for a step-function pulse entering a 1 μsec delay line	Attenuation 1 MHz	4 MHz	7 MHz	10 MHz	30 MHz	Propagation time to damp the amplitude by factor e^{-1} (μs)
RG-65/U	10.3	950	140	~ 60 ns 7)	180 db/km			706 db/km	1310 db/km	6.6 (1 MHz)
HH-1500	9.8	1500	260	30 ns 7)			2.2 db/μs			3.9 (7 MHz)
HH-1600	7.0	1800	3500	100 ns 7)		1.9 db/μs				4.6 (4 MHz)
HH-2000	9.9	2200	360	35 ns 7)			2.2 db/μs			3.9 (7 MHz)
HH-2500	7.6	2900	2000	105 ns *		1.6 db/μs				5.4 (4 MHz)
HH-4000	9.0	3800	3300	130 ns 7)		2.5 db/μs				3.5 (4 MHz)
Rectilinear Lines				Risetime (0-50 %) for a step-function pulse entering a 150 nsec delay line	(db/km) 1 MHz	10 MHz	100 MHz	200 MHz	1000 MHz	(1 MHz)
AF(ZE) 50-7	25.8	50	4.1	~ 30 ps **	1.5	4.8	17	(400 MHz) 38		24
RG-19/U	28.5	52	5.0	53 ps 7)	1.7	6	22		115	25
RG- 8/U	10.3	52	5.0	0.35 ns 7)	5.7	19	69		295	7.6
RG- 9/U	10.7	51	5.0	0.35 ns *	5.7	19	66		279	7.6
RG-68A/U	5.0	50	5.0	1.8 ns 7)	14	43	174		722	3.0
RG-62/U	6.2	93	4.1	0.71 ns 7)*	11	33	102		328	3.2
RG-63/U	10.3	125	4.1	0.32 ns 7)*	8	20	66		230	4.4
3C-2V	5.8	75	5.0	1 ns **	13	42		194		3.4
3D-2V	5.7	50	5.0	~ 1.5 ns **	14	46		215		3.1
1.5C-2V	2.4	75	5.0	~ 4 ns **	27	83		389		1.6
1.5D-2V	2.4	50	5.0	~ 4 ns **	25	85	285		1000	1.7
MX50-2.2	2.2	50	4.8	1.7 ns **		(63) *	190		610	0.66 (10 MHz)
MX50-3.6	3.6	50	4.8	0.6 ns **		(24) *	110		360	1.7 (10 MHz)

* Estimated by authors from pulse photographs or data in the manufacturer's catalog, or from slight revision based on ref. 7.

** Estimated roughly by authors based on refs. 15-17.

*** Manufacturers supplying data are Showa-Densen-Denran (Kawasaki 210, Japan) for RG-65/U and RG-19/U ∿ RG-63/U , Hitachi-Densen(2-1-2, Uchinomaru, Tokyo 100) for HH-1500 ~ HH-4000 and Dainichi-Nippon-Densen (see Fig.4) for AF(ZE)50-7 and 3C-2V ~ MX 50-3.6.

N.B. As to the rise time for a unit-step-function pulse, two kinds of definition are employed. The (10-90 %) rise time for a helical delay line increases empirically roughly with the square root of the length7). The (0-50 %) rise time for a rectilinear uniform line increases with the square of the lenght7) and must be multiplied by a factor ∿ 607) (∿ 30 according to refs. 15-16) so as to be transformed to the (10-90%) rise time.

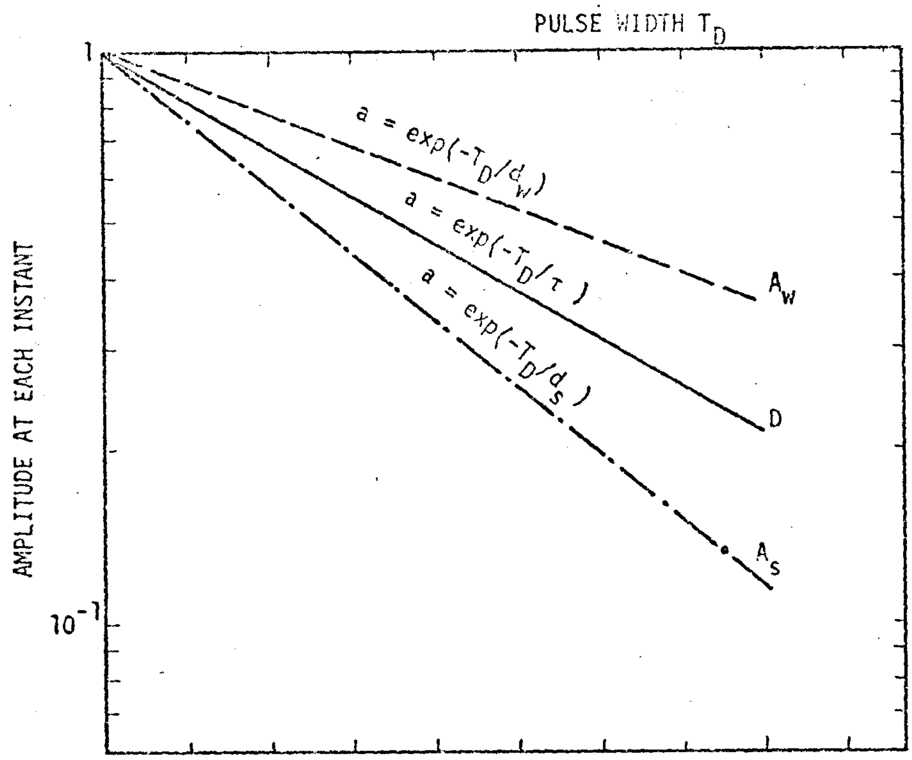

PULSE WIDTH T_D
1
10^{-1}
AMPLITUDE AT EACH INSTANT
$a = exp(-T_D/d_w)$
$a = exp(-T_D/\tau)$
$a = exp(-T_D/d_s)$
A_w
D
A_s

Figura 1

FIG.2

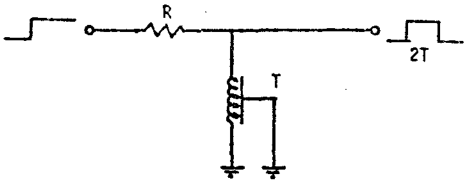


FIG.3

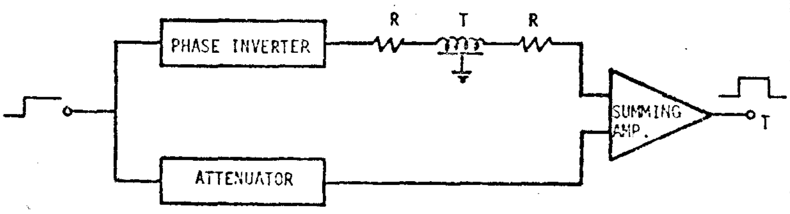
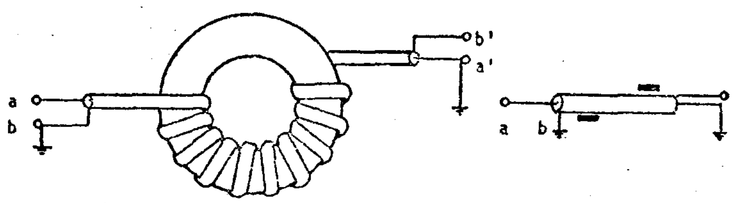
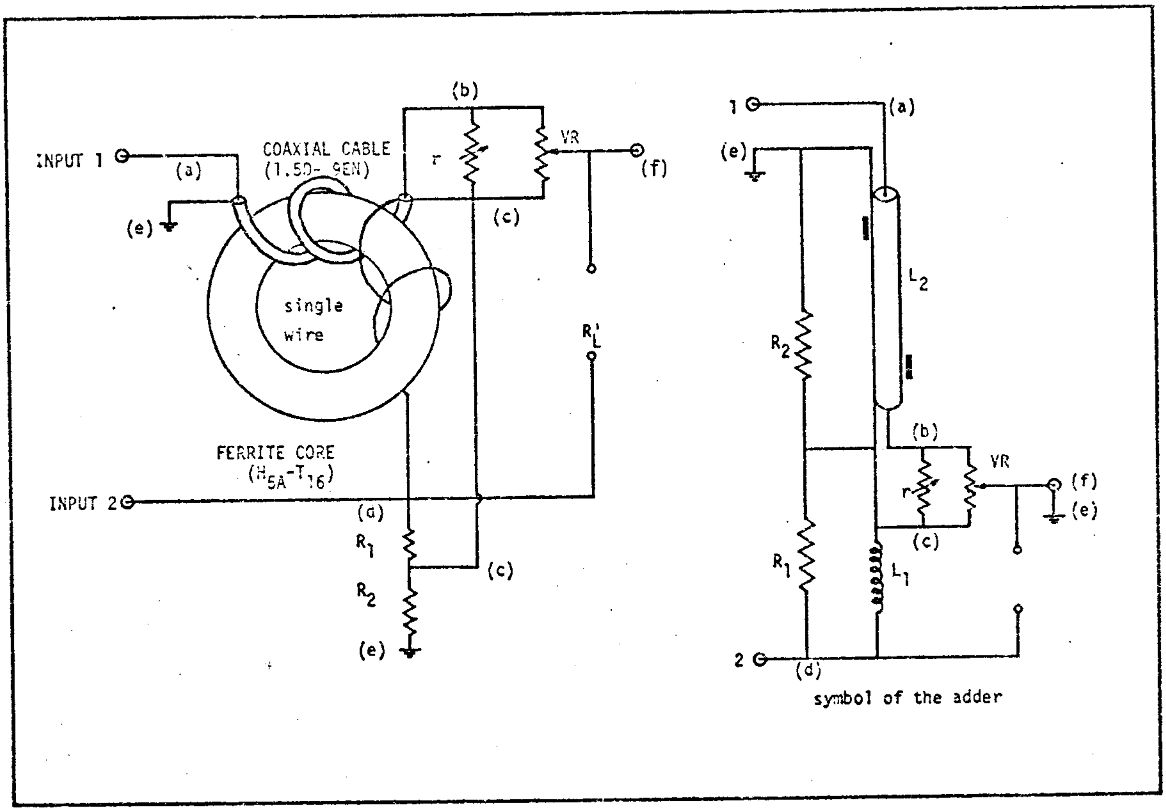


FIG.4

FERRITE CORE(T.D.K., H_{5A}-T_{16})



COAXIAL CABLE (FUJIKURA, 1.5 D-9 EH), 15 TURNS

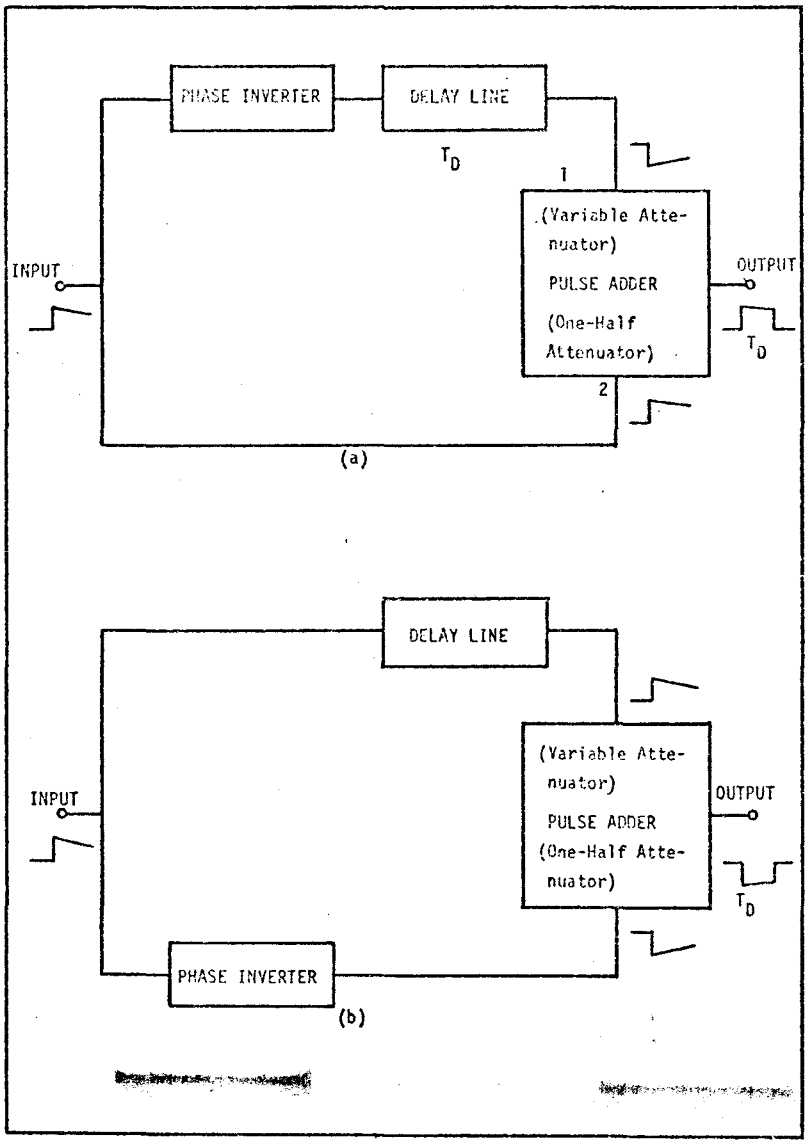


INPUT 1
(a)
(e)
COAXIAL CABLE
(1.5D- 9EN)
single
wire
FERRITE CORE
(H_{5A}-T_{16})
INPUT 2
(b)
r
VR
(f)
(c)
R'_L
(d)
R_1
R_2
(c)
(e)
1
(a)
(e)
L_2
R_2
(b)
VR
(f)
(e)
r
(c)
R_1
L_1
2
(d)
symbol of the adder

FIG. 5

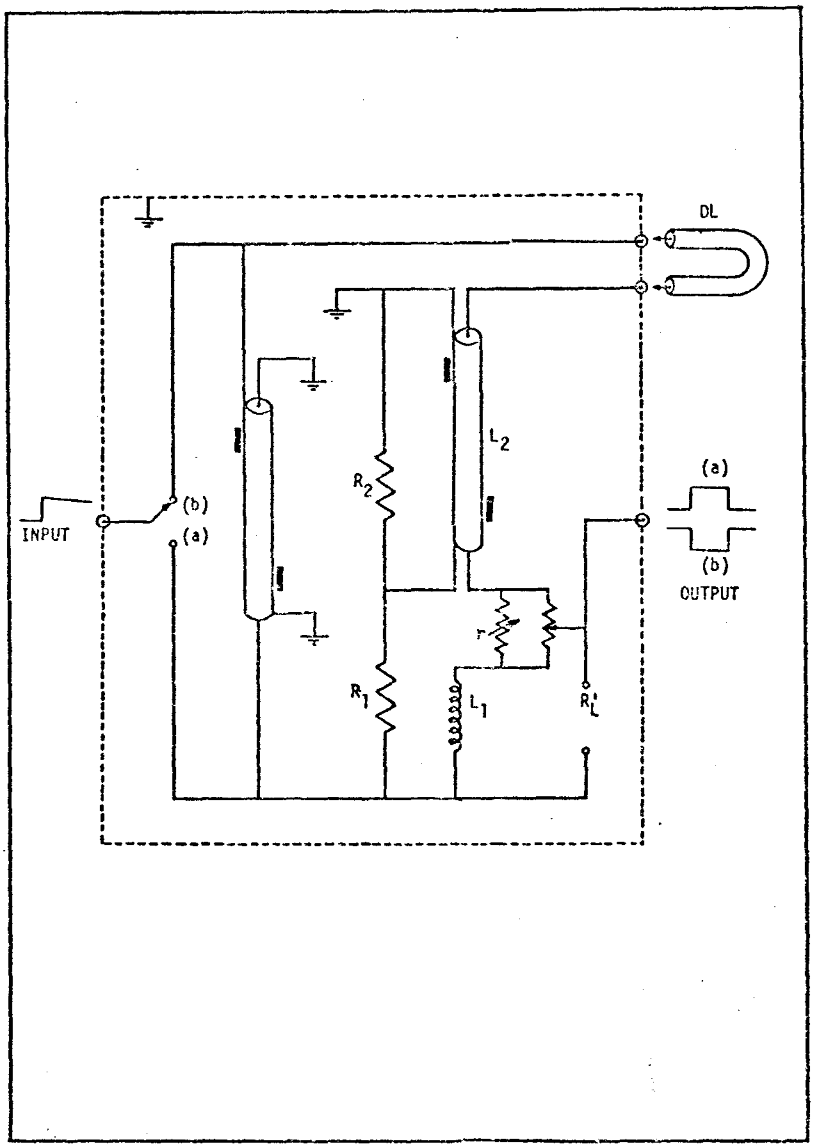
(b)

FIG. 6



PHASE INVERTER
DELAY LINE
T_D
1
(Variable Atte-
nuator)
PULSE ADDER
(One-Half
Attenuator)
2
INPUT
OUTPUT
T_D
(a)
DELAY LINE
(Variable Atte-
nuator)
PULSE ADDER
(One-Half Atte-
nuator)
INPUT
OUTPUT
T_D
PHASE INVERTER
(b)

FIG. 7



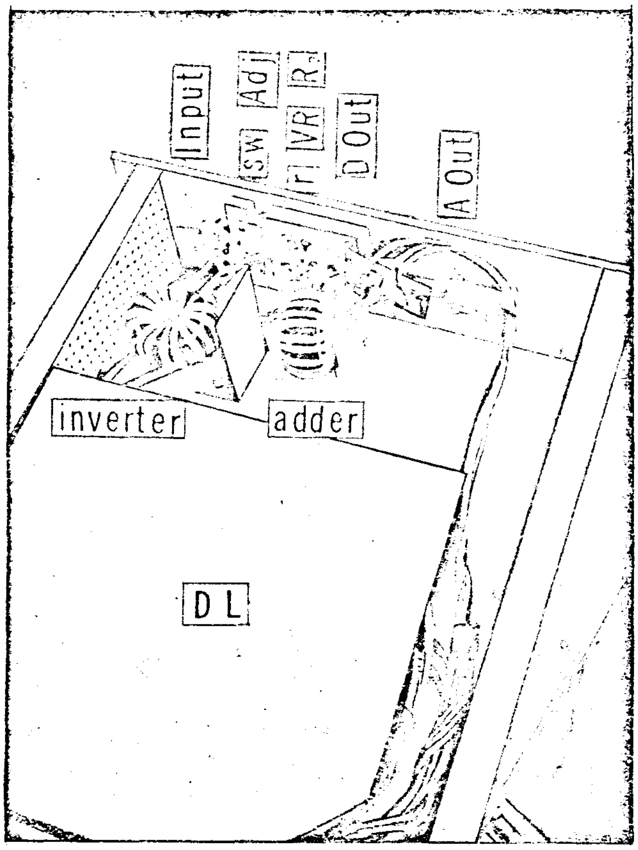


Figura 8