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SEMILOGARITHMIC AMPLIFIER SYSTEM

W. W. Goldworthy

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Summary

This report describes an amplifier system having the capability of compressing several decade pulse height ranges into several linear step gain ranges covering only a single decade. Rather than employing the usual frequency, voltage, and temperature dependent low voltage semiconductor switching and conduction characteristics, this amplifying system utilizes the reproducible, easily calibrated, abrupt, high level limiting characteristics of a transistor operational feedback amplifier.

The several cascaded stages of this amplifier, whose individual outputs drive a pulse adder, use these abrupt, high level limiting characteristics to determine the boundaries between several well defined linear step gain ranges as a function of pulse amplitude. The gain in each of these ranges is established by that gain available prior to the point where limiting occurs in the amplifier system.

Such an amplifying system has great application in the measurement of wide range nuclear or other pulse height spectra, where the relatively poor resolution uniformity of multichannel analyzers either dictates an unreasonable number of available channels or the use of some sort of compression amplifier. The plot of Fig. 1 clearly displays this resolution nonuniformity.

Introduction

The block diagram of a three-step compression amplifier using this principle is shown in Fig. 2. Typical gain and output characteristics of the three-step ranges are expressed by the following relationships.

1. Range 1
   \[ E_o = K_t E_1 [G_2 G_3 + G_2 + 1] \approx 0 - 3.3 \text{ v} \]
   \( \text{(no limiting)} \quad A_1 = G_2 G_3 + G_2 + 1 \)

2. Range 2
   \[ E_o = K_t [E_1 G_2 + E_1 + K_3] \approx 3.3 - 6.6 \text{ v} \]
   \( \text{(Stage 3 limiting)} \quad A_2 = G_2 + 1 \)

3. Range 3
   \[ E_o = K_t [E_1 + 2K_3] \approx 6.6 - 10 \text{ v} \]
   \( \text{(Stages 2 and 3 limiting)} \quad A_3 = 1 \)

Application

The use of this amplifying system appears quite attractive for many types of wide range, high resolution analysis, since it possesses qualities of good resolution uniformity, ease of calibration, stability, and reproducibility.

Many of the problems specific to the design of this type of amplifying system and some of those which apply generally to the design of high performance transistor feedback amplifiers in general will be discussed.

Design Goals

In order to produce a useful step-linear amplifying system, many details with regard to overload performance, limiting characteristics, stability, control of step ranges, calibration, pulse addition, and noise must be considered.

Since this amplifier must operate over extremely wide amplitude ranges, its ability to handle high overloads with a minimum of error and blocking is imperative. Characteristics and control of gain and limiting must also be such as to produce an accurate, reproducible, easily calibrated, step linear amplitude gain function having a minimum percentage of interstep transition.
Minimum amplifier noise contribution so as to insure excellent resolution and the use of pulse adding techniques from the various level stages which will not produce undesirable feedback are also necessary. Means for controlling total gain, gain per step, transition points, and bandwidth are also highly desirable for a practical, flexible, amplifier design.

Operation

To more easily understand the operation of this amplifier, let us apply various input levels to the amplifier of Fig. 2 with the assumption that all stages limit abruptly at 10 volts and have gains of 10 in section 2 and 9 in section 1. Using these assumed gains and limiting levels, the output voltages at points A, B, and C as shown in Fig. 3 will result from various amplifier inputs.

It will be noted in Fig. 3 that the sum of the three voltages: A, B, and C, follows an amplitude-step linear relationship over a range of three decades. Those input voltages between .01 and .1 volts produce a linear sum range of totalized output voltages of from 1.0 to 10.0 volts, those between 0.1 and 1.0 volts produce a linear sum range from 10.00 to 100.00 and those between 1.0 and 10.0 volts produce a linear sum range from 10.00 to 30.00 v. This is however only approximate since the actual range breaks occur at voltages slightly higher than 10 and 20 volts but for the simplicity of illustration are shown as 10 and 20. In order to utilize and sum these individual voltages of A, B, and C, these individual outputs are fed into an operation adder amplifier having appropriate gain to transfer the output voltages at points A, B, and C as shown in Fig. 3 will result from various amplifier inputs.

<table>
<thead>
<tr>
<th>INPUT VOLTS</th>
<th>OUTPUT A VOLTS</th>
<th>OUTPUT B VOLTS</th>
<th>OUTPUT C VOLTS</th>
<th>SUM OUTPUT VOLTS</th>
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<tr>
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<td>.09</td>
<td>.09</td>
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<tr>
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<td>10.0</td>
<td>30.0</td>
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</tr>
</tbody>
</table>

Note - Since abrupt limiting occurs at 10 v. in all stages, all products > 10 = 10.

Design

Although only stages having gains compatible with linear decade steps are shown in the example, stages having other gains and amplifiers having any reasonable number of steps can be employed to select the amplitude step gain characteristics desired. Probably the most important consideration to the successful operation of such a step linear amplifying system is the attainment of very abrupt stable, quick recovery, limiting characteristics.

To fulfill this requirement, high feedback ratio, internally D.C. coupled operational feedback amplifiers are employed. These units provide exceptionally abrupt limiting characteristics, recover rapidly from overload, and have an easily adjustable limiting level.

So as to prevent objectional feedback propagation from amplifier output to input through the adder matrix, operational feedback amplifiers having input and output impedances low with respect to matrix impedances were chosen to provide sufficient attenuation. Where bandwidth might be great, it will probably be necessary to time phase the various adder amplifier inputs with appropriate time delays.

Due to the extremely large dynamic range handled by the amplifier and since many of the measured pulses drive portions of the amplifier into limiting conditions, much consideration must be given to providing excellent overload properties.

The techniques well known for reducing the effects of high overload on amplifier resolution are applied. These consist of using double delay line clipping techniques in order to develop pulse symmetry and reduce base line shift to a minimum, using internally D.C. coupled feedback amplifiers so as to prevent the blocking or operating point shift of the individual cascaded feedback amplifiers, and using interstage coupling circuits possessing close to unity charge-discharge characteristics.

Associated with the problem of double delay line clipping in transistor circuits are also:

1. Those problems of maintaining proper line termination over wide dynamic ranges, 2. Excessive gain loss when line matching, 3. Associated noise problems, 4. Maintaining proper base line recovery over wide dynamic ranges, and 5. Nonsymmetry development resulting from poor overloading symmetry.
The standard operational amplifier of Fig. 4 was chosen as the basic feedback amplifier loop building block because of its simplicity, stability, and input and output impedance characteristics. This type of amplifier is ideally suited to use with either shunt or open-type clipping lines, since the impedances at points B and C are very close to zero and almost ideal resistive line termination can result. Blocking problems are also minimized by the direct coupled arrangement used, and overload symmetry characteristics are easily controllable by the proper choice of parameters.

Line Clippers

Of the three basic types of line clippers available, the open-end reflective types appeared to be best suited for use with transistor circuitry. Shunt-type line clippers suffer from severe signal loss and/or high output impedance when properly terminated inviting in general poorer noise performance. Line clippers employing active elements suffer from poor base line recovery characteristics over extreme dynamic ranges, as a result of small active circuit nonlinearities. They also require additional circuitry and utilize greater clipping line length.

Open-type line clippers, on the other hand, have the advantage of sustaining no greater signal loss than two and provide lower impedance output drive than equivalent impedance shunt clippers. Both of these advantages aid in maintaining minimum electronic noise.

Probably the most serious problems encountered with clipping line use are the maintenance of proper termination and base line recovery over an extreme dynamic range. Termination problems when using transistors can be resolved by insuring the constancy of driver stage output impedance and driven stage input impedance under all conditions. Base line recovery problems can be well handled by employing completely linear circuitry, proper circuit parameters, and appropriate input pulse shape.

Constant low operational amplifier output impedance can be easily maintained by the use of the output cascaded emitter follower circuit shown in Fig. 5 which exhibits by itself the desired low output impedance in the presence of stage saturation or cut off.

Constant input impedance of the driven stage without discontinuity is possible by providing circuitry that prevents the cut off of the input transistor of this stage. One such circuit is shown in Fig. 6 where the diode CR-1 prevents stage cut off by reduction of its gain characteristics below unity on overdrive. The use of a less than unity gain driven stage can also eliminate input stage cut off. The use of a reverse diode connected between the input transistor base and emitter would be quite ineffective in preventing wide input impedance variations as a result of the discontinuity existing between conduction of the transistor base and that of the diode.

Prompt base line recovery will result if:
1. Input pulse shapes compatible with line loss characteristics are maintained, 2. The circuitry associated with the clipping lines is exceptionally linear, 3. Good pulse symmetry is maintained, and 4. Amplifiers possess close to unity interstage charge discharge characteristics.

Packaged lines rather than cable lines were chosen for clipping use. The slight roughness exhibited by pulses as a result of the use of these lines appeared to create no problem and the cost, space, and fabrication problems created by the use of cable lines were avoided.

Since the operational amplifiers are internally D.C. coupled, there is no problem with internal blocking for periods following high overload; however, since A.C. interstage coupling does exist, some thought was given to the effect of overload upon the recovery of these circuits. The maintenance of close to unity charge-discharge characteristics is desirable on both non-symmetrically and symmetrically operated stages to insure a minimum base line shift in the presence of high rate or high overload pulses. Reverse conducting diode CR-2 of Fig. 6 provides an approximation of this desired characteristic by providing a relatively similar reverse polarity impedance characteristic as the input transistor is driven beyond cut off.

Symmetry Problems

A further symmetry unbalancing characteristic inherent with transistor circuitry is that of base region charge storage generated by base saturation conditions. In a double delay line amplifier where symmetry balance is important, the storage time characteristic associated with saturation conditions developed on overload may cause a considerable dissymmetry. This dissymmetry may be kept to a minimum by choosing transistors exhibiting low storage time characteristics or by employing non-saturating circuitry.

Output Stage

Figure 7 shows the output stage of the amplifier which is similar to the other stages, with the exception of the complimentary emitter follower arrangement capable of low impedance dual polarity output drive. Resistor R6 in the output circuit of this stage is for the purpose of preventing ringing which might otherwise occur at a frequency determined by the length of cabling connected to the output. Although it is easily possible to obtain less than a one ohm output impedance from the amplifier if R6 is deleted, the nuisance of having to properly terminate the output cabling greatly outweighs the inconvenience of having a fixed output impedance of 10 ohms.
The resulting summed output pulses from the compression amplifier, which are derived partially by limiting action, very accurately follow the pattern of Fig. 3; but as individual stages are driven beyond the point of initial limiting, pulse broadening will occur. Because of this pulse shape change, many pulse analyzers which are pulse shape sensitive will incorrectly measure pulse height thereby displaying some nonlinearity. Some modification of the output pulse shape is desirable to eliminate this measuring error.

Pulse Shape Change

It was found empirically that a short R.C. differentiation (Approx. 10 μs T.C.) placed at the input of the last stage eliminated this type of linearity difficulty by providing a pulse shape that falls fairly rapidly from its peak value.

It should be pointed out that the gain steps will not be exactly equal in voltage range. This inequality will be greater at low step gain ratios as a result of the larger voltage contribution of the earlier stages in the adder arrangement. Balancing of these ranges can be accomplished, however, by employing unequal limiting levels at the various stages. Where several step gain ranges are to be made available considerably more complex range switching arrangements are needed because of the necessity of changing the limiting levels of the various stages. The unit described displays some variation in the width of the three step ranges since this switching of limiting levels is not provided.

Typical characteristics of the compression amplifier are displayed in the analyzer plots of Fig. 8. They show a compressed three decade pulse height range of pulse generator signals. Linear ranges and boundaries are quite evident.

Figure 9 shows the display of step linear gamma spectra.

Figure 10 shows the schematic of the three level step compression amplifier. As will be noted from observing this schematic, none of the means for pulse shaping control of gain and polarity selection are present. These are provided for in a presently existing linear amplifier (TranLamp LRL Dwg. 15X1695) which precedes the step compression amplifier. The schematic of this amplifier is shown in Fig. 11. This amplifier possesses properties previously described which are desirable for driving the wide range compression amplifier. These properties of low noise, pulse symmetry, low base line shift, and excellent stability insure the high resolution performance desired of the step compression amplifier system. The pictures of Fig. 12, taken from an analyzer display of a linear sweep generator spectrum, show the linearity as well as the abrupt range switching that may be achieved by using this amplifier technique. Only a very few analyzer channels representing less than 5% of the total range are consumed in the gain transition areas. Stability in each range is essentially independent of limiting characteristics since instabilities in the limiting characteristics will only produce effects in boundary positions and not in the gain available between these boundaries.

In the frequency region above that compatible with feedback amplifier techniques, other limiting and mixing techniques will have to be employed to provide the step compression described.

There seems no logical reason why this described technique cannot be applied to distributed and nonfeedback amplifiers which are commonly used in the region of frequencies approaching 200 mc.

Use of this described system can be employed wherever wide range display of information is desired. Its independence from the usual poor stability semiconductor characteristics makes it quite attractive for many types of wide range information display or analysis where greater stability and ease of employment will inevitably result.

ACKNOWLEDGMENTS

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REFERENCES

Fig. 1. Plot of pulse height resolution as a function of analyzer channel.
Fig. 2. Step compression amplifier block \( R_1 = R_2 = R_3 \). Limiting voltage at A, B, and C = 10 v.
Fig. 3. Voltage summing from points A, B, and C.
Fig. 4. Basic operational amplifier
Fig. 5. Low-Z output modification
Fig. 6. Amplifier with overdrive protection
Fig. 7. Complementary follower output stage
Fig. 8. Three decade pulse amplitude spectrum
Fig. 9. Log plot of combined spectra of Am\(^{241}\), Be\(^{203}\), and Na\(^{22}\) with a step gain ratio of 5.
Fig. 10. Schematic of Semilogarithmic Amplifier
Fig. 11. Schematic of Driver Amplifier LRL 15X1695
Fig. 12. Plot of amplifier output with a three decade sweep generator range input. Slight roughness results from pulser and sweep synchronization.
Fig. 13. Linear Driver Amplifier and output Compression Amplifier.
Fig. 2

\[ \text{Adder} \]

\[ E_{in} \]

\[ G_2 \]

\[ G_3 \]

\[ R_1 \]

\[ R_2 \]

\[ R_3 \]

\[ E_0 \]
Fig. 5
Fig. 7
Fig. 12
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