ELECTRONOTES 72

Newsletter of the Musical Engineering Group
203 Snyder Hill Road
Ithaca, N. Y. 14850

VOLUME 8, NUMBER 72 DECEMBER 1976

GROUP ANNOUNCEMENTS:

CONTENTS OF EN#72

We are pleased that there are a lot of new types of circuits in this newsletter. In the ENS-76 series, we have three new devices which have not been included in any previous ENS synthesizer series. Also, these three modules are not generally found on conventional commercial synthesizers, so these may be of interest to builders who are expanding a commercial system. There are also new devices to be found in the Reader's Circuits section.

As our readership expands, we are finding it more and more difficult to answer individual letters. To do this properly would probably take one full work day each week. We can not do this at the present time. To help things out, we are taking two steps. First, we will be answering more and more questions in the newsletter by making Reader's Questions a more regular feature. Secondly, we will try to reactivate the "consultant group" by updating the list. Persons who would like to answer questions by mail are encouraged to participate. If interested, write us telling us in which areas you would be willing to answer questions. We would then print your name and address along with information on the type of questions you would handle. Readers could then write directly to you with their questions and will supply you with self-addressed stamped envelopes for your reply.

We remind you that this is the last issue of the current volume and that all subscriptions expire with this issue. If you have not sent in your renewal, you should do so by Jan. 20, 1977 in order to get the renewal discount which we offer.

The second group of application notes (#11 - #20) are nearly finished and will be mailed to those requesting mid-month service for 1977 as a second bonus. The individual titles will be listed next issue.

News Items:

John Snell has asked us to announce that a new publication, Computer Music Journal, is about to begin publication. The journal will appear every other month beginning in January 1977 and will be published by PCC, Box E, Menlo Park, CA 94025. The subscription fee is $14 for one year. The first issue is expected to be about 50 pages and length will grow later on.

Synapse has changed its format slightly and is now available from Schill and Schill Publishing, 1151 Macneil St., San Fernando, CA 91340. The price is $6 for six issues in the US and Canada ($7.50 foreign). Volume 1, Issue 3 contains many items of interest including an interview with Tom Oberheim, comments from Larry Fast about Synergy, Kraftwerk Interview, and several circuits and electronic device descriptions. We feel that the new format and publishing philosophy as evidenced by Issue 3 make this well worth looking at.

"Beyond the Sun" is an electronic realization of Gustav Holst's "The Planets" done by Pat Gleeson on the Eμ Polyphonic Synthesizer. The recording is on the Mercury lebel SRI-80000 and is probably in your local record store. "Brainwave Music" is a recording of the music of David Rosenboom issued by A.R.C. Records, PO Box 3044, Vancouver, BC, Canada, V6B 3X5. The cost of the record is $5.49 postpaid (add $1 for air). For a complete list of other records and publications of A.R.C. (Aesthetic Research Centre of Canada), write to A.R.C., PO Box 541, Maple, Ontario, Canada L0J 1E0. The listing includes work in biofeedback and sound sculpture.

ARP Instruments, 320 Needham St., Newton, MA 02164 is looking for persons to fill several positions in the senior technician - junior engineer area for R&D work. Resumes should be sent to David Friend at ARP. Copies of resumes filed with Electronotes have been forwarded to ARP.

Reader's Questions:

Q: Where can I get diagrams of the waveforms and spectra of traditional instruments?

A: We often get questions of this general form. We feel it is implied by this sort of question that the questioner wants to use this information to synthesize the sounds of these instruments. Therefore, it is important that it be understood that this sort of information is not generally available, and that the reason it is not is because a waveform or spectral description (that is, a stationary description) is not very useful.

If you look at an actual waveform of a musical tone from a traditional (acoustic) instrument such as a trumpet or a violin, etc., you will find that no two cycles of the waveform are exactly alike. In general, the cycles will be most alike in the steady state portion of the waveform, and least alike during the transient portions (e.g., during the attack). But there's more! Even during steady state, no waveform at one pitch will be exactly like the corresponding waveform for a different pitch. In fact, there may be great differences in some instruments for changes of pitch that are less than one octave. But there's still more! Even for two instruments of the same type (e.g., two violins), the waveform will be different even for the same pitch. Furthermore, the waveform will be different for the exact same instrument played by different players or by the same player at different times.

Now, a fixed spectral description is slightly more successful (probably because the ear is relatively insensitive to phase while waveform depends on phase). But even with a spectral distribution, all the variations that were mentioned for waveform apply as well. Thus, neither the waveform or a fixed spectrum is useful.

What is useful? Probably it is most fruitful to try to get the attack transient of the instrument, and the time varying spectrum of the steady state and the decay. The answer is not completely in at the present time.

Some Theory On The Design And Application Of Variable-Slope Filters:

-by Bernie Hutchins, ELECTRONOTES

Since variable-slope filters are not devices commonly found in voltage-controlled synthesizers, we feel it is appropriate to say a few words about them prior to giving a working circuit in the ENS-76 series. Naturally, we have to say what they are, how one might build one, and what they might be good for. We should also point out right away that we are just beginning to think about these, so there may well be other and better ways of doing these things.

WHAT IS A VARIABLE-SLOPE FILTER?

To know what a variable-slope filter is, we must first know what slope is. We must also know what we mean by variable - that is, we should know that we want it voltage-variable, just like any other filter parameter. You probably know that a filter (low-pass let's say) has a roll-off rate that is described as decibels per octave, and that commonly we see roll-off slopes of 6db/octave, 12db/octave, 18db/octave, 24db/octave, and so on. Slope seems to come in units of 6db/octave. Why? Well, for example, consider the first-order low pass filter shown in Fig. 1. The transfer function is:

$$T(s) = \frac{-1}{1 + sCR}$$

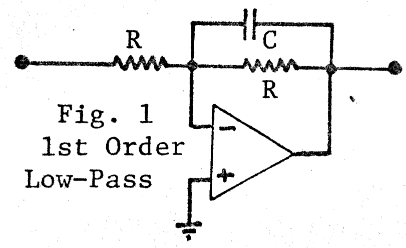


Fig. 1
1st Order
Low-Pass

For high frequency (large values of s), the transfer function has a magnitude $|T(j\omega)| \approx 1/2\pi fRC$. This is the frequency response function. Clearly, if the frequency doubles (one octave), the response will drop by a factor of two (6 db). Thus, at high frequencies, the response drops by 6db/octave. Likewise, if we had a second-order filter, the response would drop off as $1/f^2$, an octave change in frequency would drop the response by 1/4 (12 db), and the response would be 12 db/octave. This response is shown in Fig. 2, and is properly termed "asymptotic" slope, but is often just called slope. Another feature of the response is the cutoff frequency, usually taken to be the point at which the response is down by 3db from the maximum. This is also called the "corner" frequency. Clearly, in the vicinity of the corner and below, the rate of change of response is less than the asymptotic slope in the example above. In general, we can expect to find a variety of values of slope in the pass-band region and in the vicinity of the corner frequency. Chebyshev filters for example will have positive, negative, and zero slope in the passband, and the initial roll-off slope (i.e., at the corner) will be greater than the final (asymptotic) slope. All filters of the same order will have the same asymptotic roll-off however. It is convenient to observe these slopes by plotting the response on log-log graph paper. A 6db/octave slope is a 45° angle on such paper while a 12db/octave slope is sharper (≈63°), 18db/octave is still sharper (≈72°) and so on. Thus, asymptotic slope is just a matter of the filter order. However, as a practical matter, we have a good number of different slope values that appear for a given response. Thus, in general we want to look at a sort of average value of slope that occurs in the region of most interest, which will generally be the region in which the response drops from its maximum value down to 1/100 or 1/1000 of this initial value.

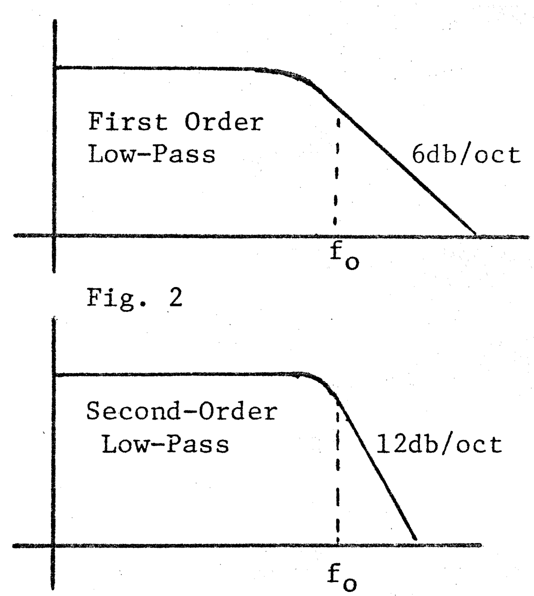
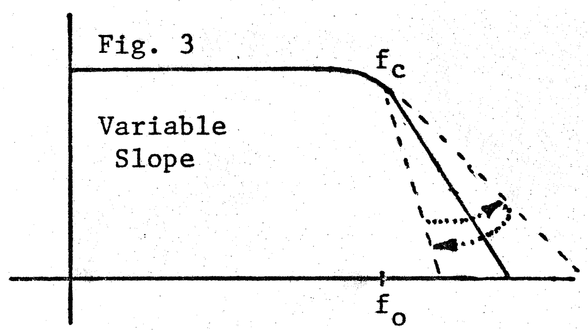
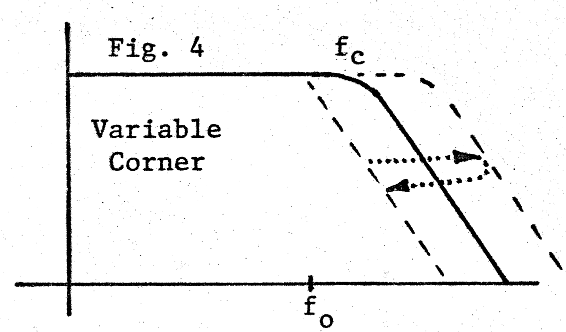


Fig. 2

Even given this understanding of slope, we still do not know how to control it well, let alone voltage-control it. While there are variations of slope in any response curve, and we can change this by changing the order of the filter or its characteristic (as from Butterworth to Chebyshev), we are still greatly confined by certain general trends and do not have a way of varying the slope in a continuous manner. Ideally, we might like to have something like the variability suggested by Fig. 3 where the corner frequency is fixed while the roll-off slope varies up and down (in response to a control voltage) attached to this corner. Contrast this with the standard voltage-controlled low-pass technique where the response curve is set and the corner frequency goes up and down (Fig. 4). We have in mind here a variation in filter parameters corresponding to what we would want to occur during a single note. In both cases (Fig. 3 and Fig. 4), the initial frequency f_o is set by the pitch level of the system (as from a keyboard) while the variation indicated by the arrows is caused by an envelope control voltage.



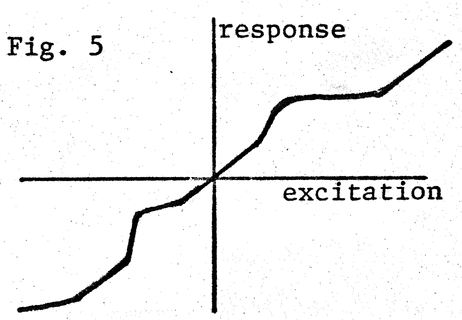


It is clear that the variations in harmonic content caused by the filter responses of Fig. 3 and Fig. 4 are similar in that they both increase the total high-frequency content as the envelope rises, and decrease the high-frequency content as the envelope falls back. There are also clear differences which we also have to consider for their musical implications.

WHY WOULD WE WANT TO USE A VARIABLE-SLOPE FILTER?

We mentioned above that both the variable slope technique and the variable corner frequency technique lead to an increase in upper harmonics as the envelope rises. In many ways, this is the general trend that we find with traditional (acoustic) musical instruments, and with pots and pans, and just about anything we might want to bang on in the real world. We are quite familiar with this type of sound, and perhaps should note that the origin is quite basic. If we look at the response function of some mechanical object to some exciting force, it is likely to be non-linear, except near the rest point (See fig. 5). Thus we expect to have low harmonic content for small excitations (low distortion) and increased harmonics for large excitations. Naturally, we expect also to find a louder sound as a result of larger excitation. Thus, we come to associate higher harmonic content with louder sounds. Of course, there is no reason why music can't be made with other relationships. In fact, this is one of the great new undiscovered areas where electronic music makes possible new music composed of what might otherwise be considered unfamiliar (hence to some, unmusical) sounds. However, higher harmonic content with higher intensity of sound is extremely popular, as evidenced by the large number of portable synthesizers using mainly low-pass filters moving upward with envelopes and being fed waveforms of high harmonic content. Here, we want to look at this more carefully to see if we can determine an optimum relationship between loudness and harmonic content.

Fig. 5



David Luce has shown ("Dynamic Spectrum Changes of Orchestral Instruments," J. Aud. Eng. Soc., Vol. 23, No. 7, Sept. 1975, pg. 567) that to a rough approximation the spectral content of traditional acoustic instruments can be described by a parametric model of the form shown in Fig. 6. There are two (or more) curves for each instrument with the lower curve corresponding to the softest sound (pp) while the upper curve corresponds to the loudest sounds (ff) available from the given instrument. While there are exceptions, and while the straight lines shown are only to be considered as rough approximations, there is a general trend as shown in Fig. 6: the high-frequency roll-off tends to become less steep as the intensity increases. This is the trend we would be immitating by using a variable slope technique as shown in Fig. 3, but not by using a variable corner frequency as in Fig. 4 since the corner frequency does not (in general) change in Luce's parametric model.

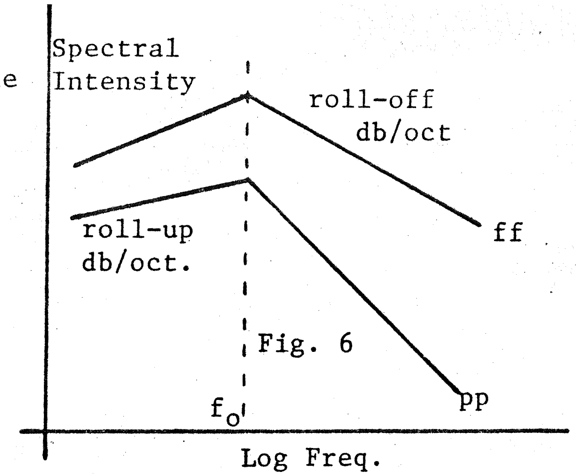


Fig. 6

We can thus try some sort of patch like the one shown in Fig. 7. Here we show the variable slope filter as being controlled by the same envelope that controls intensity (the VCA envelope). This is not absolutely necessary, and in general the user would probably want to experiment with envelopes that are different, but the patch as shown serves to demonstrate the premise that intensity (amplitude) and final roll-of slope are directly related. Here we make no provision for the change of initial roll-u slope since this is less dramatic an effect. We can then ask about the necessary rang of filter parameters. To get some idea about this, we look at the data in Luce's paper We see that roll-off slopes corresponding to the lowest amplitude range from about 10db/oct. for string instruments, through intermediate values for the woodwinds, up to 20-24db/oct. for middle and low brass. The reduction in slope from low amplitudes to high amplitude is from about 1-3db/octave in the strings to values of about 10db/oct for trumpet, trombone, and bassoon. We are thus led to the conclusion that we need slopes in the range of 24db/oct. to 6db/oct. for our experiments. Furthermore, it seem that we can expect the technique to work best for the more penetrating type of sounds and to be less effective for string type sounds. By this time, we have pushed the data of Luce's paper much further than we really should, but have shown how we can speculate on the way we may want to use the patch of Fig. 7.

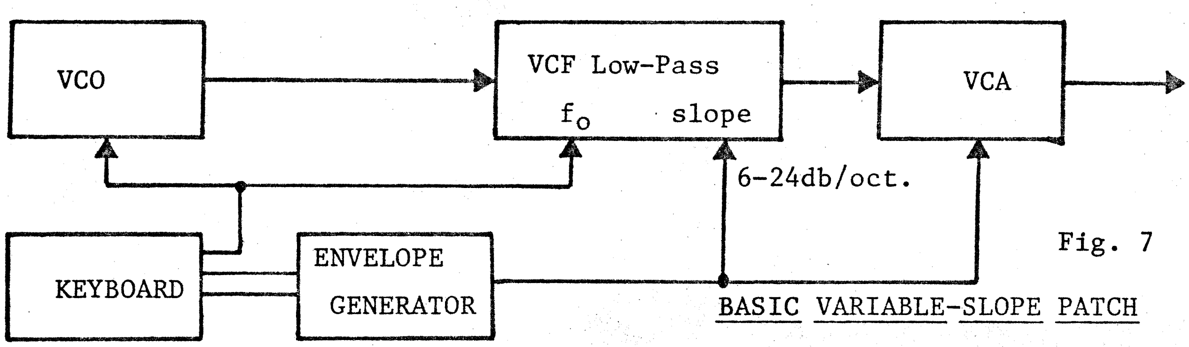


Fig. 7

BASIC VARIABLE-SLOPE PATCH

HOW DO YOU BUILD A VARIABLE-SLOPE FILTER?

To the best of our knowledge, there is no well known or standard way of making a variable-slope filter. An earlier discussion of this was given in EN#59 (2) and a method was proposed there. We have checked this out and found it to work within certai limits. A slightly more restricted method, but one which gives a smoother variation will also be discussed below, and is the basis of the ENS-76 circuit to follow.

The first thing that comes to mind when one tries a variable slope filter design is to start with a very sharp slope and a very shallow one and try to do some sort of averaging. For example, a 12db/oct. and a 0db/oct. filter might be averaged. We could start with something like a 2nd order Butterworth for the 12db/oct. and a piece of wire for the 0db/octave "filter" and try an average as indicated in Fig. 8. There

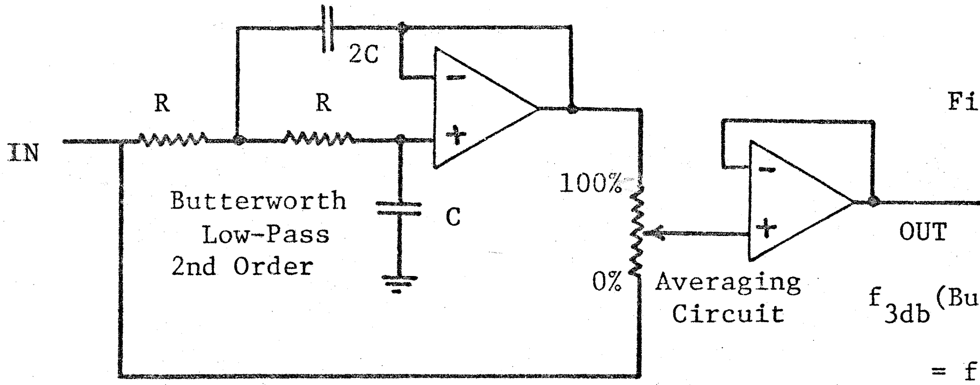


Fig. 8

$$f_{3db}(\text{Butterworth}) = \frac{1}{2\pi\sqrt{2}RC}$$

$$= f_{90°}(\text{Butterworth})$$

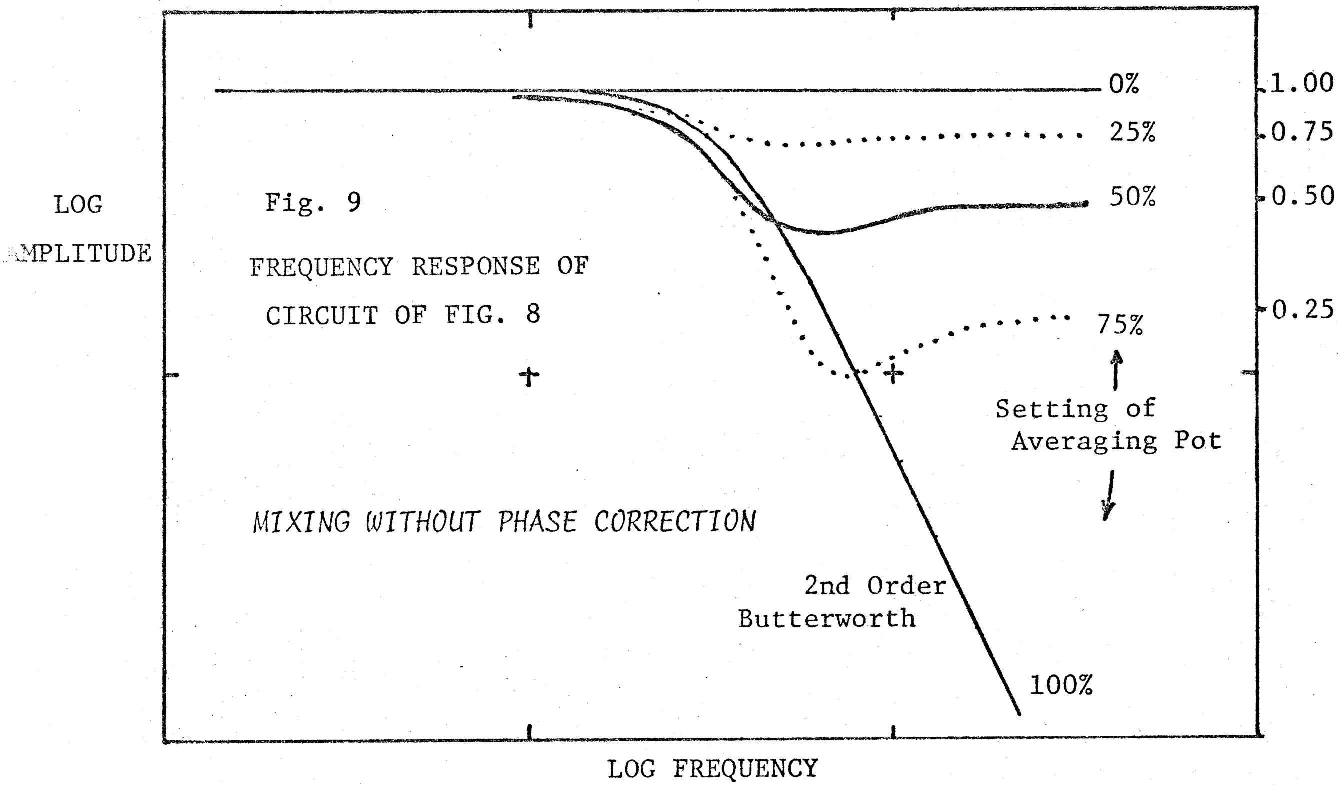


Fig. 9

FREQUENCY RESPONSE OF CIRCUIT OF FIG. 8

MIXING WITHOUT PHASE CORRECTION

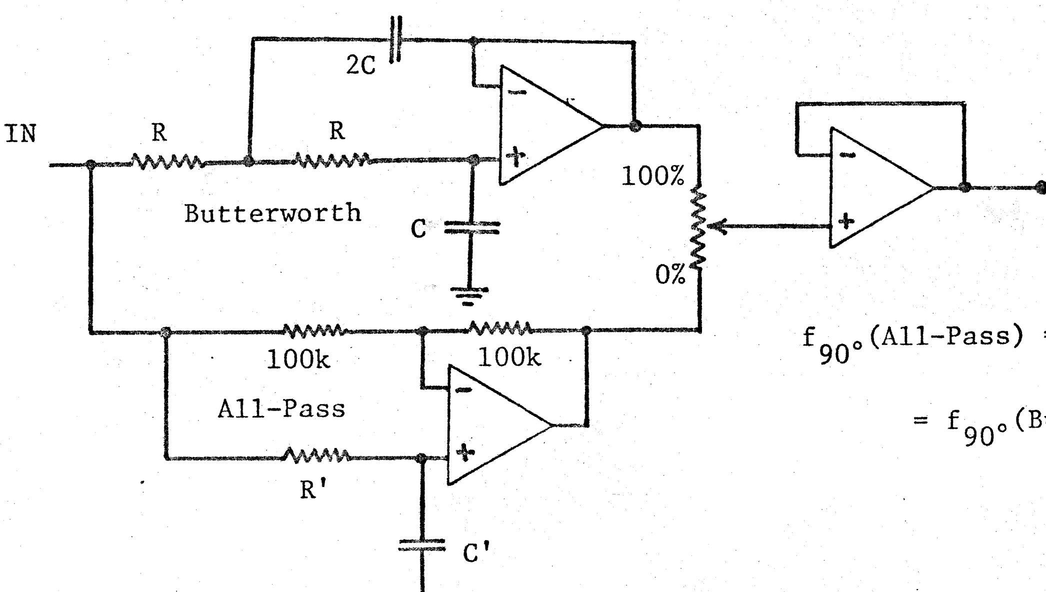


Fig. 10

$$f_{90°}(\text{All-Pass}) = \frac{1}{2\pi R'C'}$$

$$= f_{90°}(\text{Butterworth})$$

are at least two things wrong with this. First, if we do average, trying an equal mix for 6db/oct. for example, we will end up at a constant level of 1/2 for high frequencies and not a continuing 6db/octave roll-off. Secondly, since there is a phase shift across the filter, but not across the piece of wire, the averaging process will be greatly complicated. The response of the circuit of Fig. 8 is shown in Fig. 9. So we next try some sort of all-pass filter in place of the wire to give some phase correction (we try to match the phase of the Butterworth for example). This is shown in Fig. 10 where the first order all-pass network varies from 0° to 180° with the 90° frequency matched to the 90° frequency of the Butterworth. This is a fairly good phase correction as can be seen by the results in Fig. 11, but we still have the problem that the slope does not continue down but instead levels off for a constant value of response. We could then try to add something to keep the response rolling down. In the EN#59 discussion this was done by adding another all-pass which would roll off above the first. The output of the second all-pass filter is then added to the first and they cancel as frequency continues upward. It is easy to show that the two summed all-pass filters

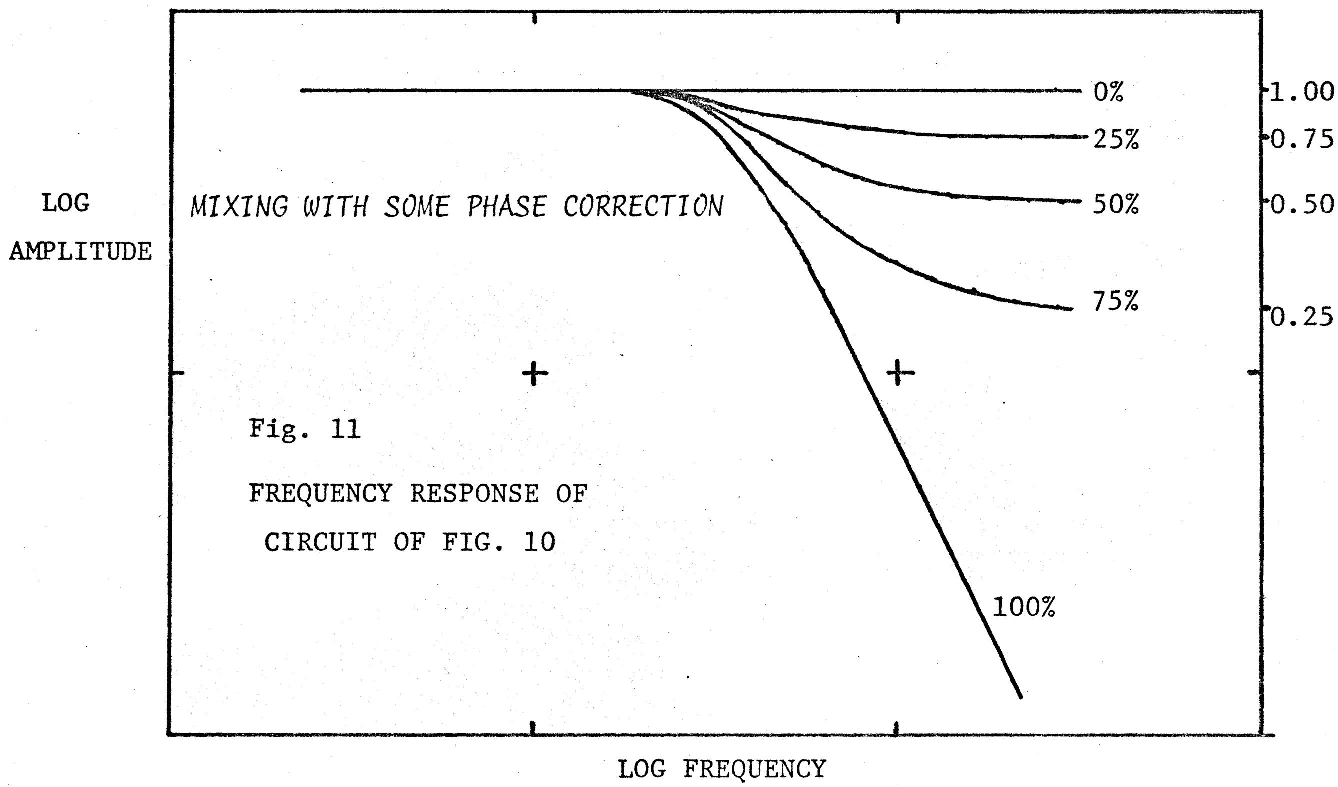
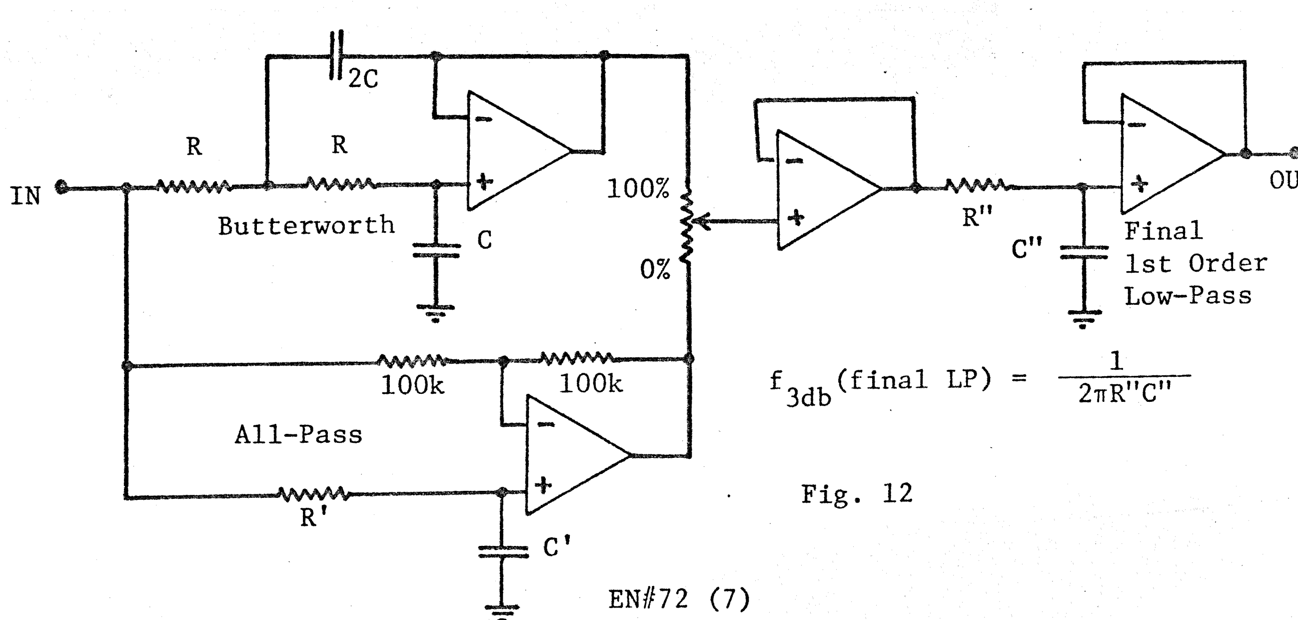


Fig. 11

FREQUENCY RESPONSE OF CIRCUIT OF FIG. 10

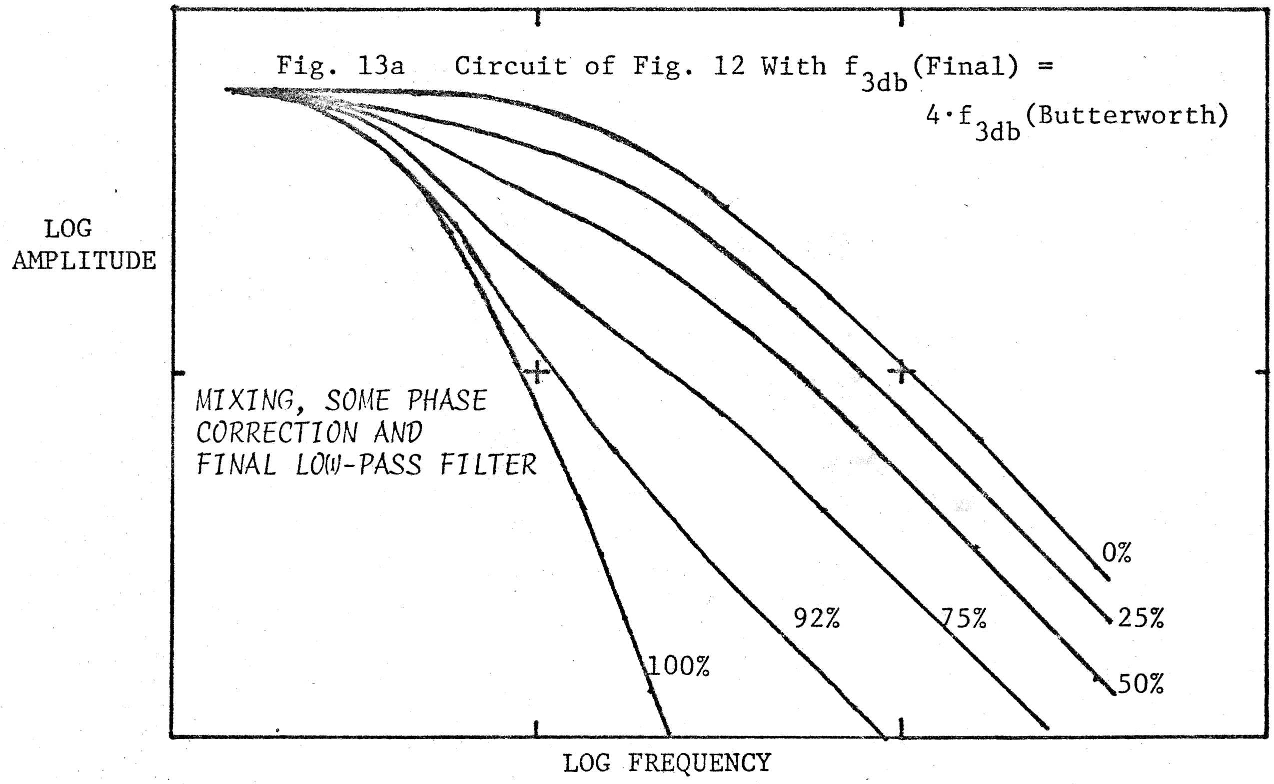


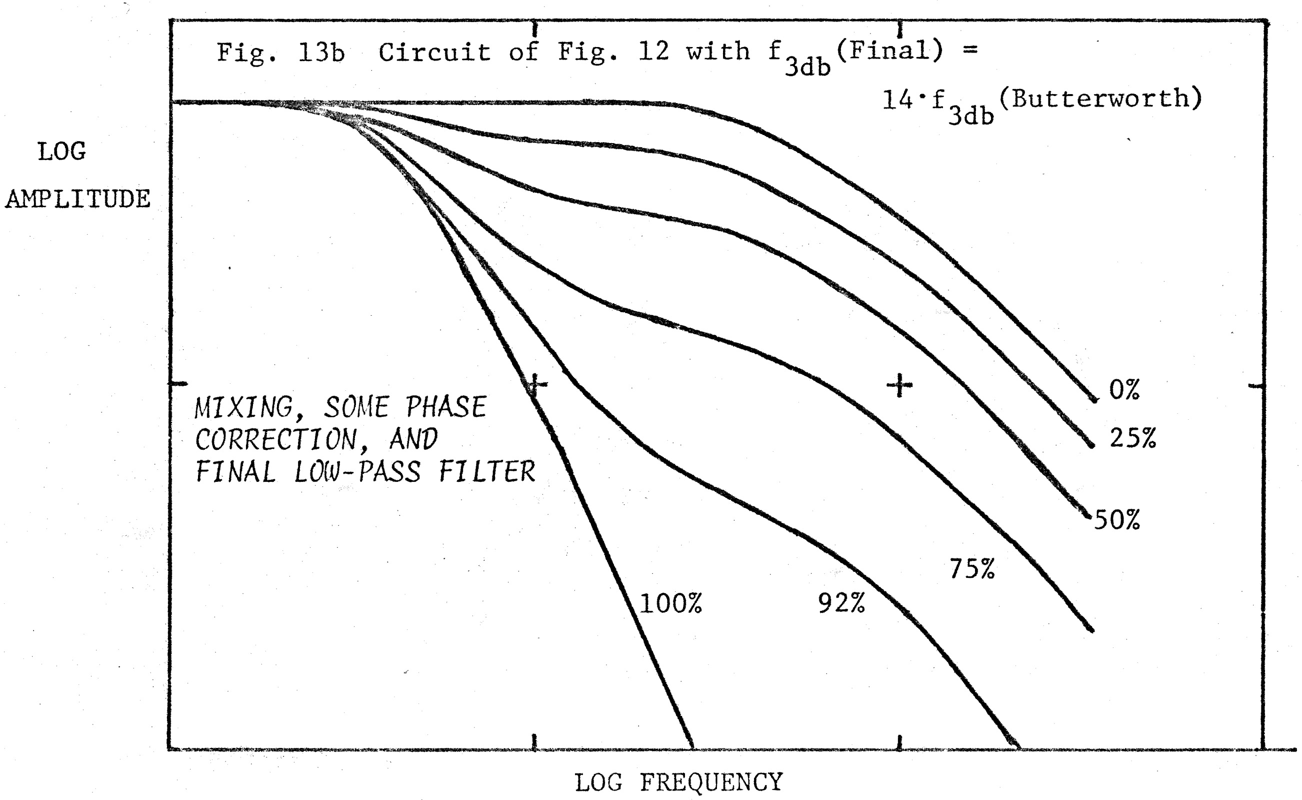
$$f_{3db}(\text{final LP}) = \frac{1}{2\pi R''C''}$$

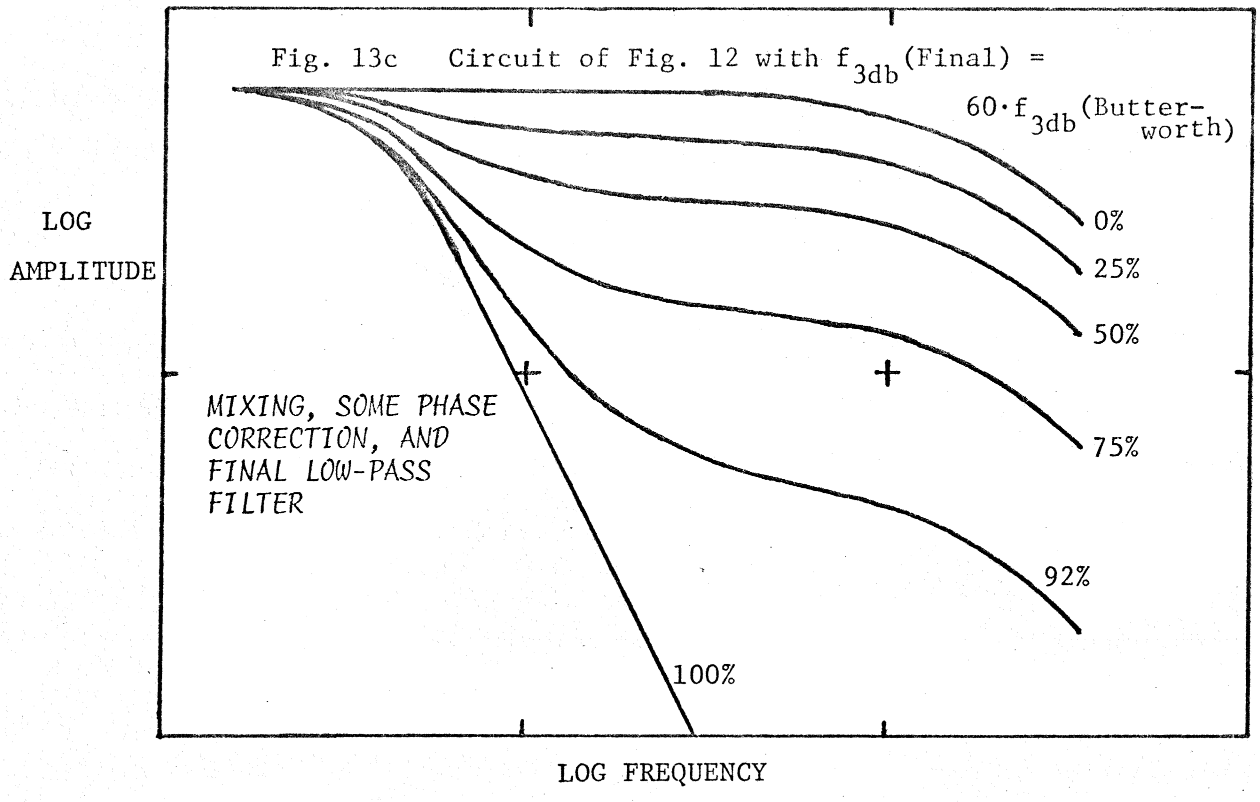
Fig. 12

are equivalent to a single-pole low-pass (6db/octave) in this case. Thus, we chose the average slope between 12db/oct. and 0db/oct. in this case. This all works, but it is probably more rational to add a single pole low-pass to the output rather than have the two all-pass filters summed for approximately the same result. This setup is shown in Fig. 12.

The results of the setup of Fig. 12 are shown in Figures 13a, 13b, and 13c, each of which represents a different upward dispalcement of the final low-pass filter above the 3db frequency of the initial Butterworth filter. Note in all three cases that the final roll-off rate of 6db/octave (45° angle) is approached for the higher frequencies. It is







also clear that there is a wide range of roll-off slopes distributed among the three graphs. In Fig. 13a, we see reasonably straight line roll-offs for slopes from 12db/octave (the 100% curve) to about 6db/octave (the 75% curve). Reasonably good approximations to slopes less than 6db/octave can be found in Figures 13b and 13c. As can be seen, these curves wobble around a bit, but it is necessary to view this wobble in the proper perspective.

First, there is no musical a priori reason to prefer absolutely uniform roll-off curves to a set of curves that are just an approximation. In fact, there is no musical a priori reason to prefer the variable slope (as opposed to fixed slope), so we can not say ahead of time that we must have straight line roll-offs. Secondly, it is probably the case that it is the wide range of variation rather than the exact details of the individual steps of the variation that is important. Thus, we can consider the method we have outlined as a reasonable approach to be tested. However, since we really would like to have a 24db/octave slope to begin with, we would probably start with a 4th Order Butterworth and use a second order all-pass and second order final low-pass. The general idea is shown below in Fig. 14. Note that this diagram shows a generalized version of Fig. 12 with voltage-control added. Note that we show a path of the slope control voltage to the final low-pass, in the likely event that we will want to adjust this somewhat as the percentage mix is changed. The need for this is inferred from a study of Figures 13a, 13b, and 13c. All the necessary voltage-control mechanisms for this setup are fairly standard designs.

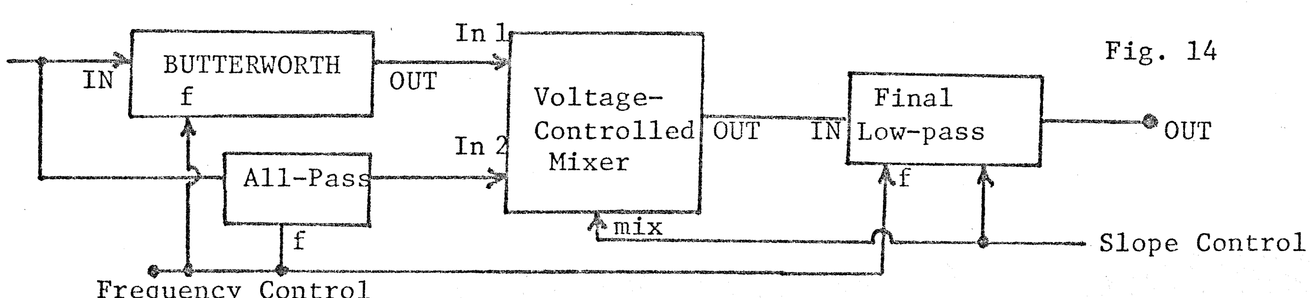


Fig. 14

A SECOND METHOD OF SLOPE CONTROL

The second method of variable slope control is one which is slightly more restricted in range of slope, but which is probably easier to implement. It is the method we will be using for the ENS-76 variable-slope filter, and is one which can be attempted by using synthesizer modules of a conventional synthesizer, as long as three or more VCF's are available. The general method should be clear from the specific example we give, which is a brief description of the ENS-76 circuit to follow.

Fig. 15 shows six individual first-order low-pass VCF's all connected in series. All six sections are controlled in parallel at one volt per octave by the corner freq. control voltage. In addition, a second control input is used in the second through sixth sections. If the slope control voltage is zero, all six stages are lined up at the same frequency, and the asymptotic slope is 6·6db/octave = 36db/octave. However, since all the sections are first order, the actual slope over the region of interest is more like 24db/octave max. Now, suppose we "remove" one of the sections by raising its cutoff frequency to some upper value above the region of interest. The asymptotic slope remains the same (36db/octave), but the slope in the region of interest is reduced from 24db/octave to about 20db/octave. Similar removal of another section will change the slope in the region of interest to about 16db/octave, and so on down to one remaining section for about 4db/octave in the region of interest. Now, if we remove these sections by sliding them up gradually, the change becomes continuous. We could slide one up, then go back for the next, and so on, but it is probably simplest to just slide them all at one time, but each one at a different rate. This is what is accomplished by the circuit diagrammed in Fig. 15. The roll-off curves for the setup in Fig. 15 are shown in Fig. 16. The full range of slope is from slightly in excess of 24db/octave up to slightly less than 6 db/octave. This range was one of our original goals if we wanted to obtain the range suggested by Luce's paper. However, we will not get a slope of less than 6db/octave with this method. Note that if you have three standard state-variable low-pass VCF's or similar, you can approximate this response by using the 1 volt/octave inputs and the variable inputs, doing this with second order sections. In the ENS-76 circuit to follow, we actually use six simple first order VCF sections.

It is instructive to look at what happens with this scheme if we view it in terms of the movement of poles in the s-plane. This is indicated by Fig. 17. When the slope control voltage is zero, all the poles are piled up as shown. As the slope-control voltage rises, the poles spread out in the $-\sigma$ direction and as they move away from the $j\omega$ axis, they have less and less effect on the frequency response relative to the one pole (section 1) that remains in its original position.

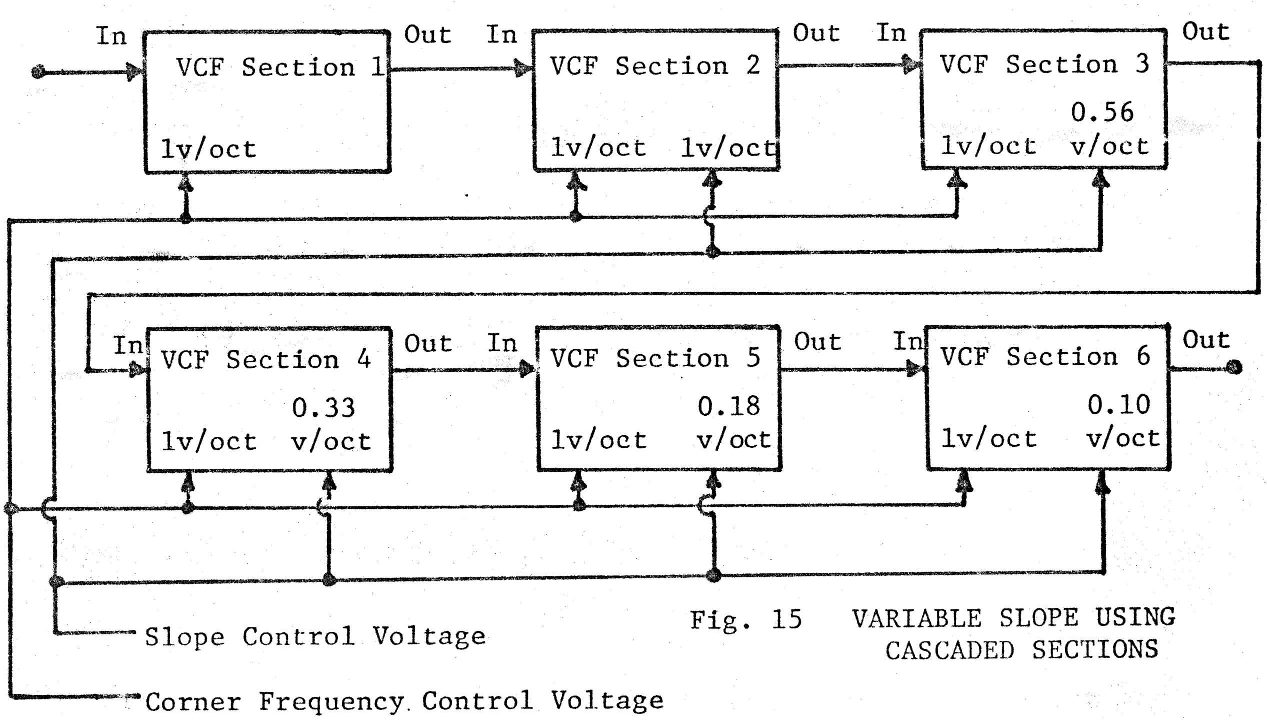


Fig. 15 VARIABLE SLOPE USING CASCADED SECTIONS

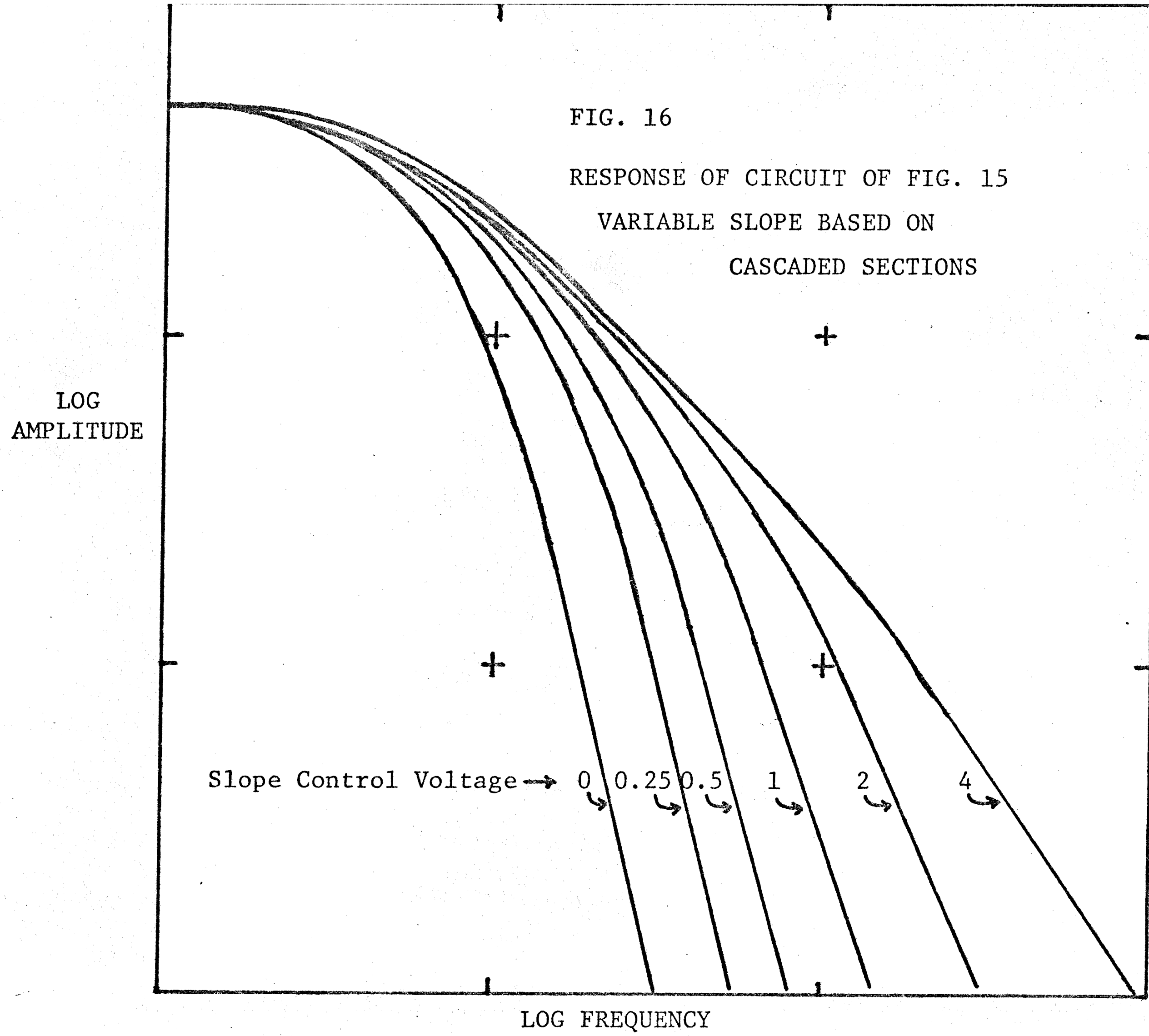


FIG. 16

RESPONSE OF CIRCUIT OF FIG. 15
VARIABLE SLOPE BASED ON
CASCADED SECTIONS

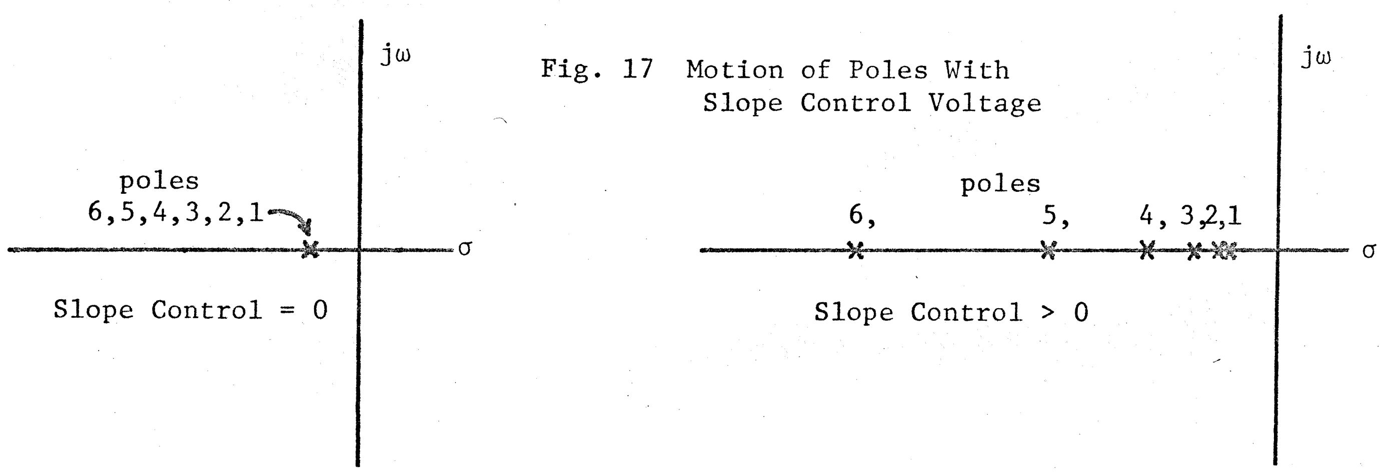


Fig. 17 Motion of Poles With Slope Control Voltage

Fig. 17 further suggests that it might be possible to obtain sharper corners and straighter roll-off slopes if one manipulated complex poles rather than just the real poles of the first order filters. We know that we can manipulate such poles in second order filters by changing the frequency and Q of the sections. Thus, it may well be possible to use three second order sections and vary the slope by changing the filter frequency and the Q of the section in the proper manner. We have not as yet done any research in this direction, but it seems evident that it can be done. Further, users of synthesizers with three or more filters may have used variable slope without knowing it.

The ENS-76 Home Built Synthesizer System - Part 6:

- by Bernie Hutchins

ENS-76 VCF Option 3, Variable-Slope Filter:

This third option for a VCF is a special type of VCF which we have not presented in the past. This filter works on a variable slope principle (See report starting on page 3 of this issue) and a discussion of the specific variable slope scheme begins on page 10. The reader should review the theory at some time, but for now he need just realize that basically we have to construct a series cascade of exponentially controlled first order low-pass sections. The first order section we will use has been discussed in various places (see EN#58 for example) and we need say no more about it here. It consists of a CA3080 and an op-amp integrator of which A7 and A13 are typical in Fig. 2 on the next page. Since we have to build six exponential converters, we can't afford to make them too fancy. Here we use fixed attenuators to set one volt/octave, and use unmatched 2N3904-2N3906 pairs for the exponential converting transistors. A typical exponential converter stage can be seen as A1, T1, and T7 in Fig. 2. We just construct six of these virtually identical stages, cascade them, and add appropriate driving voltages.

The driving voltages are supplied by the summing circuits of Fig. 1 below. The first summer (OA-1 and OA-2) receives an external control voltage (e.g., from the keyboard) and adds this to initial coarse and fine control voltages from panel pots PC-1 and PC-2. This becomes the frequency control section and drives points f1, f2,...f6 of Fig. 2. A single volts/octave trimmer is available for making small adjustments to the response to tune it to 1 volt/octave. Of course, each of the individual sections may be in error within the tolerances of the resistors, but as a practical matter, it does not seem to be a problem. The second summer (OA-3 and OA-4) provides the slope control voltage and drives points s2, s3, s4, s5, and s6 of Fig. 2. Since variable slope is the main feature of this filter, we have provided a means of either increasing the slope or decreasing it (or both) in response to control envelopes. The initial slope control varies the slope from about 24db/octave to 6 db/octave. The full control range for slope is swept by a value of the voltage driving s2....s6 from zero to +5. Fig 15 on page 11 was made using this filter, so you can get an idea of the control function from this graph.

It is possible to plot the slope control function for this filter if we can agree on a way to measure the slope. Bear in mind that the asymptotic slope of this filter is always 36db/octave no matter what the spacing of the various center frequencies of the separate sections is. However, the slope in the region of interest is always less than 36db/octave. We can measure the slope for example if we agree to take it for each of the curves between the same amplitude limits. We might want to compare slope control

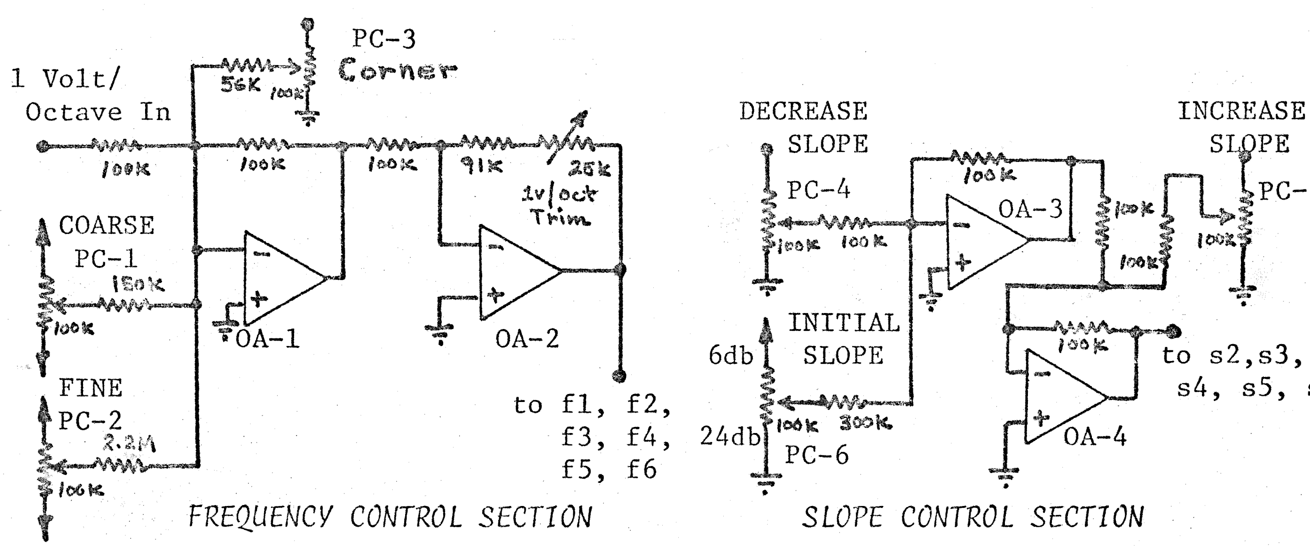
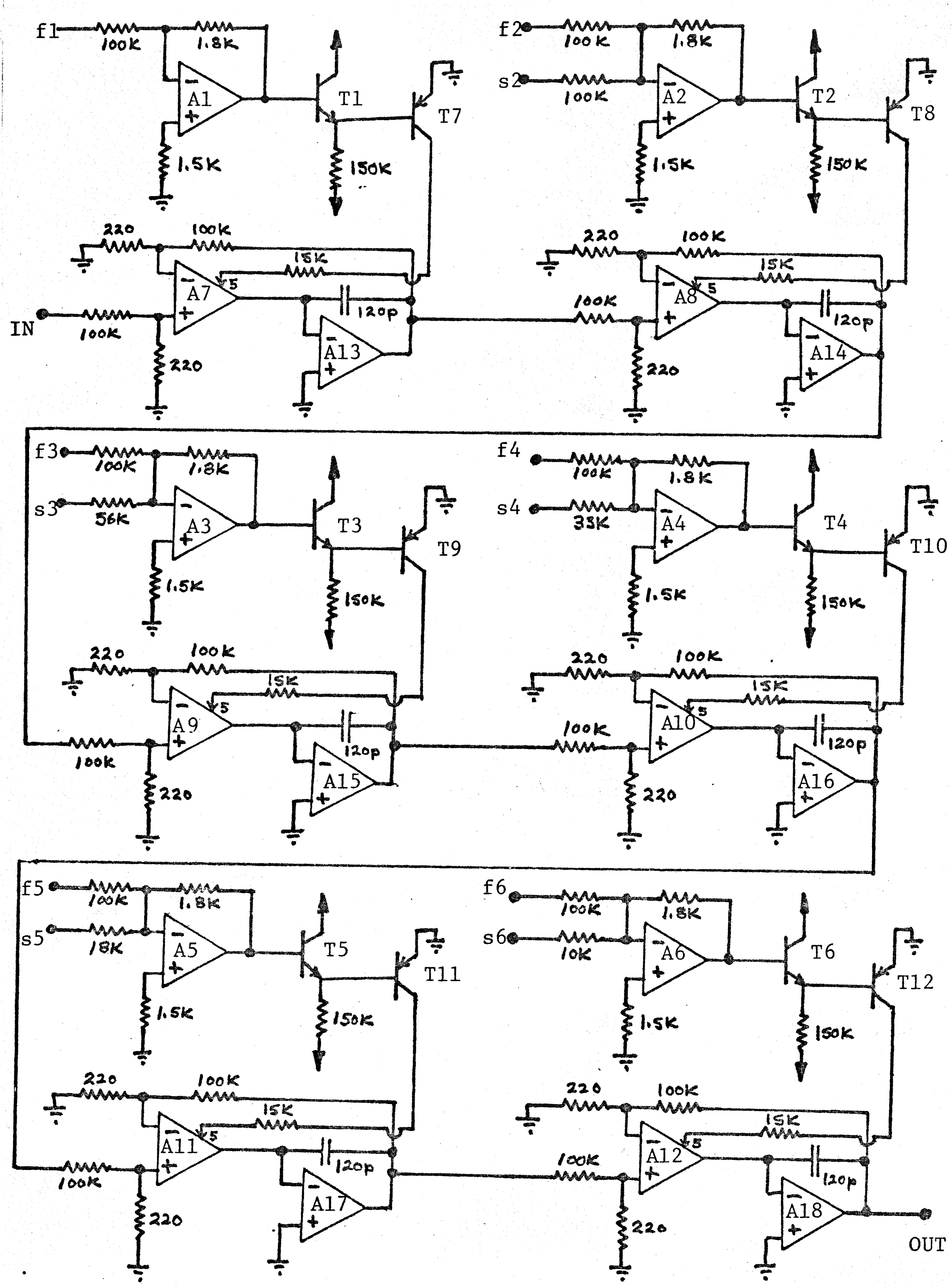


Fig. 1 CONTROL SUMMERS FOR ENS-76 VCF, OPTION 3, VARIABLE SLOPE FILTER

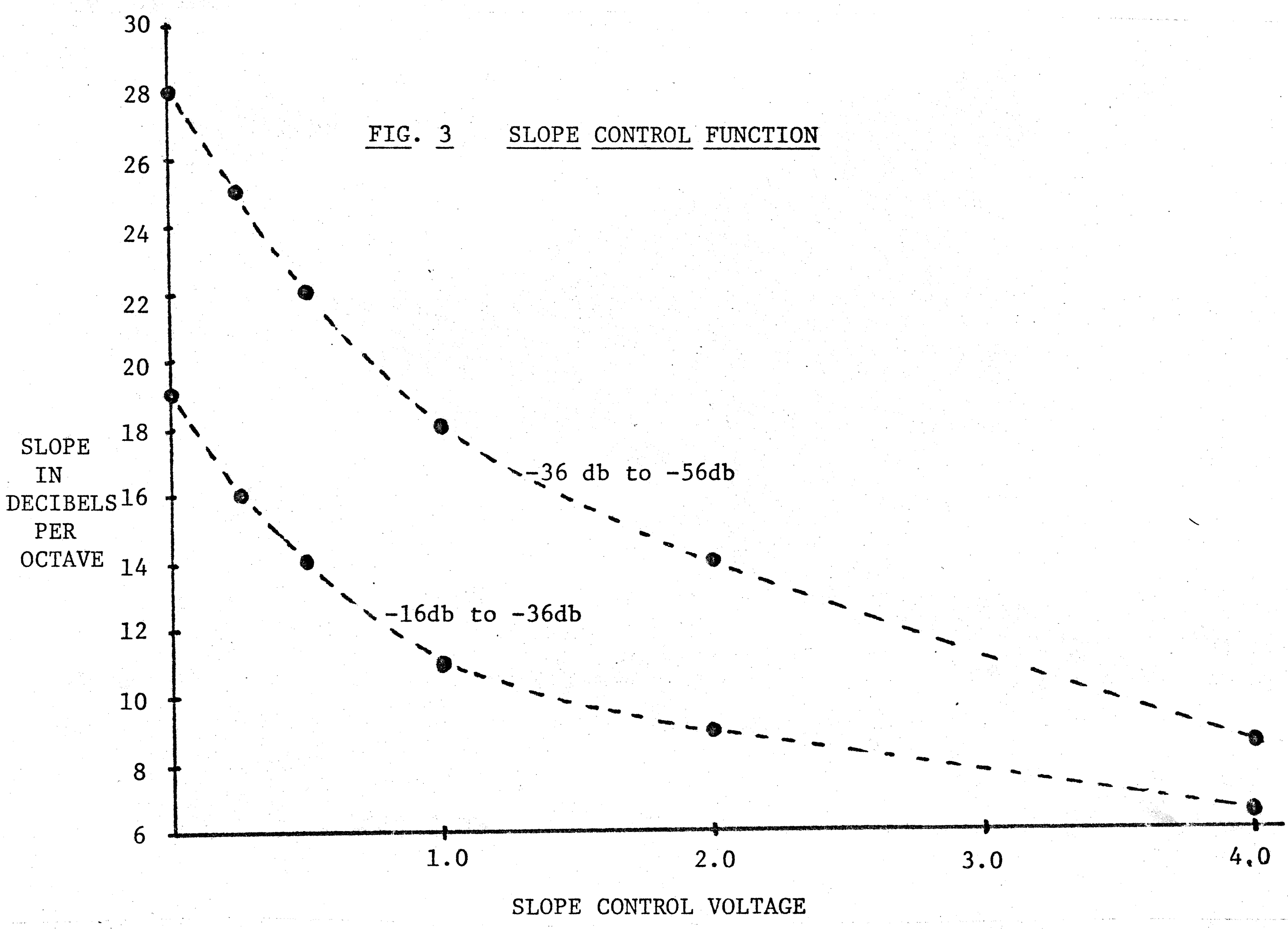
Fig. 2 MAIN FILTER SECTION OF ENS-76 VCF OPTION 3, VARIABLE SLOPE FILTER



OP-AMPS in Option 3: OA-1 through OA-4, 307 or 1/2 558;
A1, A2,...A6 = 307; A7, A8,...A12 = CA3080;
A13, A14,...A18 = 556 (or can use 307 for smaller signal levels)

TRANSISTORS: T1, T2,...T6 = 2N3904; T7, T8,...T12 = 2N3906
Pairs, T1-T7, T2-T8, etc. should be glued together with epoxy

with the slope between points on the curve where the amplitude changes from 1/10 the peak value to 1/100 of the peak value. We could also make comparisons for different limits. For the data of Fig. 16 on page 11, it was convenient to measure the slope between -16db and -36db, and then between -36db and -56db. This was done for the curves shown in Fig. 16 (0 volts, 0.25, 0.5, 1.0, 2.0, and 4.0). The results are plotted in Fig. 3 below.

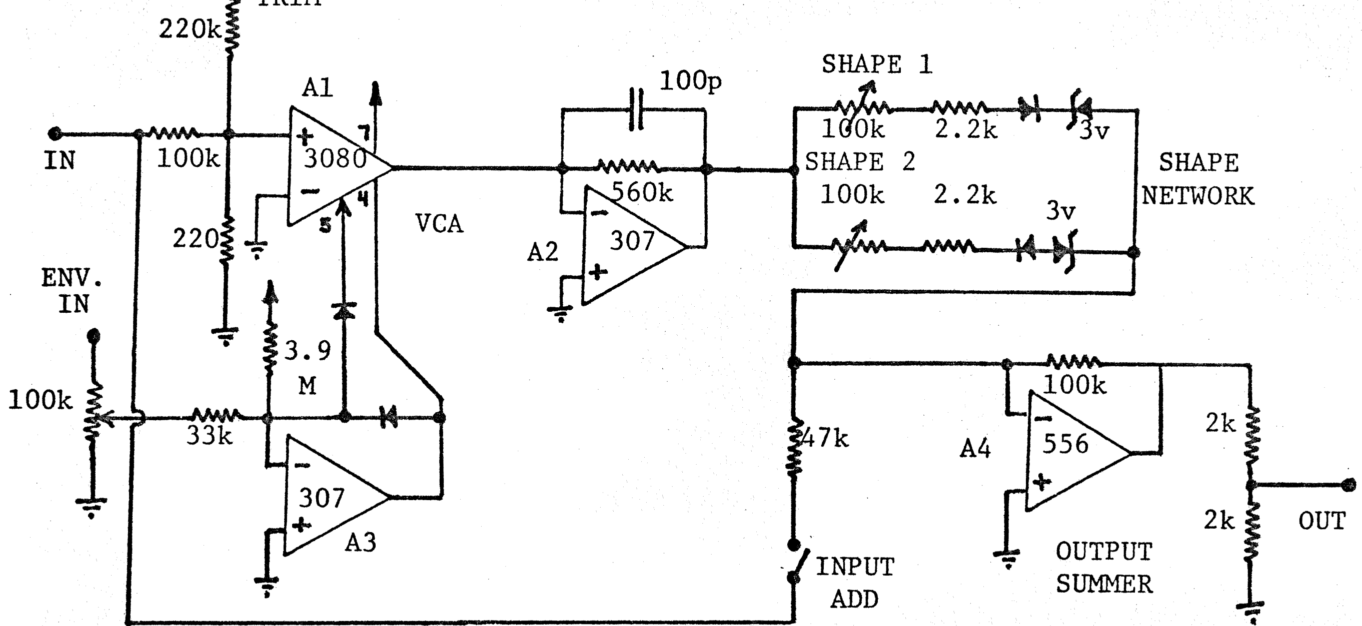


TIMBRE MODULATOR - OPTION 1

Like a variable-slope filter, the "Timbre Modulator" is not your everyday type of synthesizer module. Also like a VCF, the timbre modulator is concerned with controlling the tone color of a signal. In fact, a VCF is a timbre modulator. The present device is a special type of timbre modulator - one which adds harmonics to a waveform of low harmonic content (such as a sine or triangle). If truth be known, we developed this circuit while attempting to do something else (which we will decline to mention to avoid looking unduly foolish). When we found that it imparted variable harmonic content to sines and triangles, it was obvious that we had arrived at a timbre modulator that was in many ways the counterpart of the VCF (which works best for sawtooth waves and pulses). The circuit is quite simple as can be seen by studying Fig. 4. It is basically a VCA which drives a zener diode input stage to an op-amp. The zener diodes prevent signals of less than about 3.5 volts from reaching the output of the module. Alternatively, the signal from the VCA can be added to the input for more complex waveforms. With the input added in, the most dramatic timber modulations are achieved, but the VCA action is blocked, so another VCA is needed somewhere in the circuit (and is of course, usually available).

TIMBRE MODULATOR - OPTION 1

Fig. 4



The VCA circuit of the timbre modulator is formed by A1, A2, and A3 in a manner similar to that described in EN#63. A couple of things have been added. The 3.9M resistor holds the VCA on slightly in the absense of an envelope. This assures that the CA3080 is not cut off when the envelope is low, as this condition would otherwise pin the output of A2. The capacitor in the feedback loop of A2 is just to prevent a slight oscillation that appeared on occasion, but otherwise has no function. The output of A2 drives the input signal (as controlled by the VCA) through the shaping network into the output summer A4. In order for a signal to get through this network, the amplitude must be in excess of about 3.5 volts. Thus we are center clipping the input waveform as can be seen from Fig.5a which shows a sinewave input. Here we have assumed that the two shape controls are set to approximately the same value, otherwise the output would not be symmetrical. A case where the two controls are not set the same is shown by Fig. 5b.

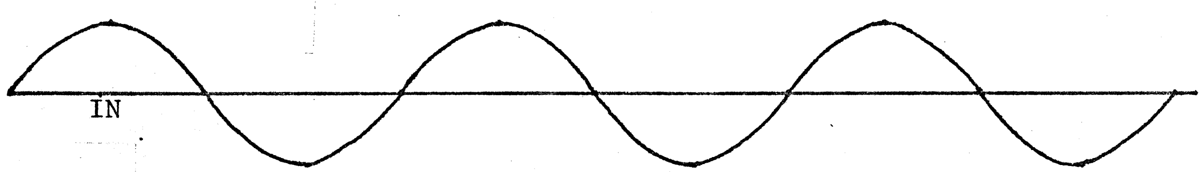
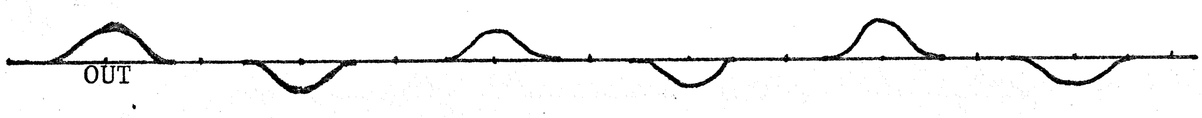


FIG. 5a SINE WAVE INPUT AND OUTPUT OF TIMBRE MODULATOR, ADD INPUT SWITCH OPEN



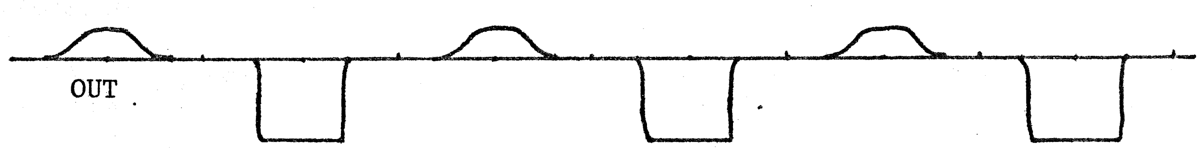


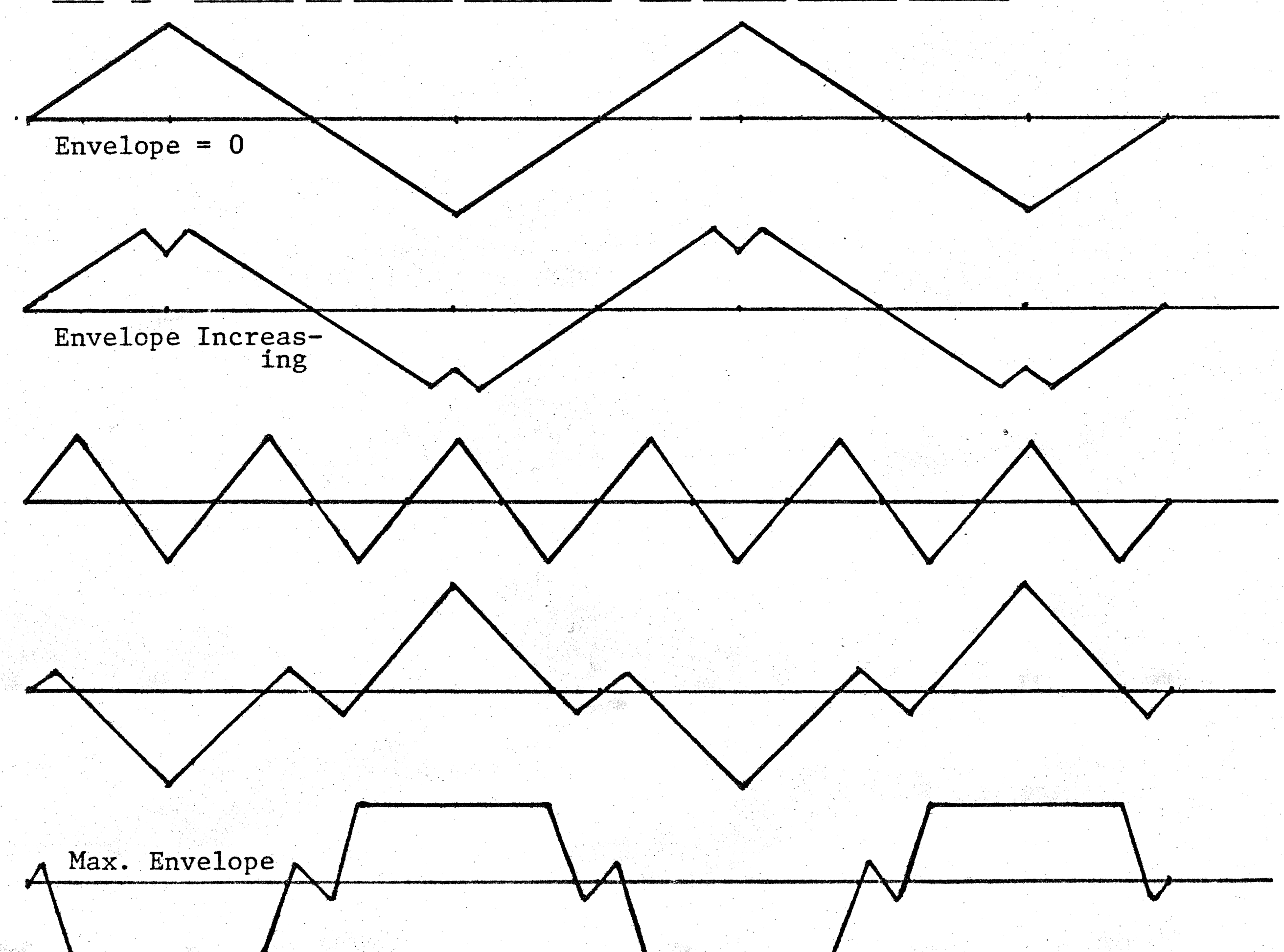
FIG. 5b SINE WAVE INPUT AND OUTPUT, ADD INPUT SWITCH OPEN, SHAPE CONTROLS UNEQUAL

By far, the most interesting effects with the timbre modulator are achieved with the INPUT ADD switch closed. With the switch closed, and the envelope at zero, the input waveform is passed to the output. Now, as the envelope rises, what is basically an inverted version of the input begins to appear at the output of A2. This would normally cancel the input if it were added to the output summer, but here it must first pass through the shape network. This leads to some very interesting results. An example of what one can expect is shown in Fig. 6 below.

When the envelope is low, the input triangle passes through unaltered. When the envelope rises, the first thing that will get through the shape network is the peak of the triangle. With the shape control pots set to about 10k each, this means that the peak of the triangle will start to cancel (and indeed cut deeper into) the peak of the input triangle, resulting in the waveform seen in the second line of Fig. 6. With a further rise of the envelope, we eventually arrive at the middle waveform of Fig. 6. While this is not a perfect waveform, it is very close to being a triple frequency version of the original triangle input. [This of course suggests a method of frequency tripling a triangle with a non-voltage-controlled version of this circuit.] As the envelope continues to rise, the output takes on the form shown by the fourth and fifth lines of Fig. 6. The fifth line is not unlike a square wave, and can be made to look more and more like a square wave by decreasing the resistance of the two SHAPE pots.

Thus, we have a device which starts with a waveform of low harmonic content, and greatly increases and alters this content as a control envelope rises. The effect is much like that achieved with a low-pass VCF and a waveform of high harmonic content. The harmonic evolutions are not the same as one gets with a filter however, so the timbre modulator should be a useful addition. It can also be used in parallel with a VCF (using the neglected sinewave output for example) and the two results can be mixed.

FIG. 6 OUTPUT OF TIMBRE MODULATOR, ADD INPUT SWITCH CLOSED



Delay Module/Sub-Module - Option 1

We have in the past seen several delay line circuits and the use of analog delay lines in electronic music has greatly increased in the last year or so. We have spent some time considering how best to incorporate delay lines into the ENS-76 system. We have arrived at what we call a "Delay Module/Submodule." It is a module in the sense that it can be easily included on a synthesizer panel, for example, and will provide a voltage-variable delay which is available for whatever use the user may decide to put it to. It is a sub-module in the sense that it is the basic framework on which any specifi module employing a delay line may be constructed.

We look at a delay module/sub-module as a device which has a voltage-controlled delay capability, a synthesizer compatible input, a delayed output with unity gain, standard power supply voltages, and a well defined bandwidth. Thus, for standard bucket brigade delay lines, the VCO which drives the delay chip should be inside the module, as should any special power supply devices required. We also prefer that the delay line be DC coupled. These requirements are made for convenience - so that the user may consider it as close to an ideal delay as is possible. Also, with the great advances in charge-transfer devices that are being made at the present time, the delay module/submodule approach allows us to make changes in systems as new devices become available.

The delay module/sub-module option 1 is shown in Fig. 7 on the next page. It consists of seven principal sections (10 sections counting duplications for the second line). We show in Fig. 7 a dual line, but it is possible to simplify this into a single line with twice the delay as will be indicated. The principal sections are [1] a multiple exponential current source (IC-1, T1, T2, T3, and T4); [2] a VCO (IC-2, T5, and IC-3; [3] a clock driver (IC-4); [4] a signal input section (IC-5); [5] the actual delay line chip (IC-9); [6] an output amplifier and DC level shifter (IC-6); and finally [7] a smoothening VCF (IC-7 and IC-8). We will look at these one at a time.

[1] The exponential current source is very similar to what we have seen in the past. The main source (IC-1, T1 and T2) drives the VCO, but note that T3 and T4 are also driven by T1 and thus (assuming identical transistors) supply identical currents to that supplied by T2. It is possible to drive additional transistors in the same manner as T2, T3, and T4 if additional control currents are needed.

[2] The VCO (IC-2, T5, IC-3) is fairly standard. It is driven by an exponential current source and thus provides approximately 1 volt/octave response. This in turn means that the delay is cut in half for each increase of one volt in control voltage. Since we are not concerned with the shape of the triangle wave in this VCO, the high frequency response is compensated with resistor R13 in a variation of a method used by S. Franco ["Hardware Design of a Real-Time Musical System," U. if Ill. report UIUCDCS-R-74-677, Oct. 1974, pg. 32]. The square wave output (output of IC-3) is shifted from ±15 to -15↔Ground by resistors R16 and R17. This drives the CMOS inverter chip IC-4.

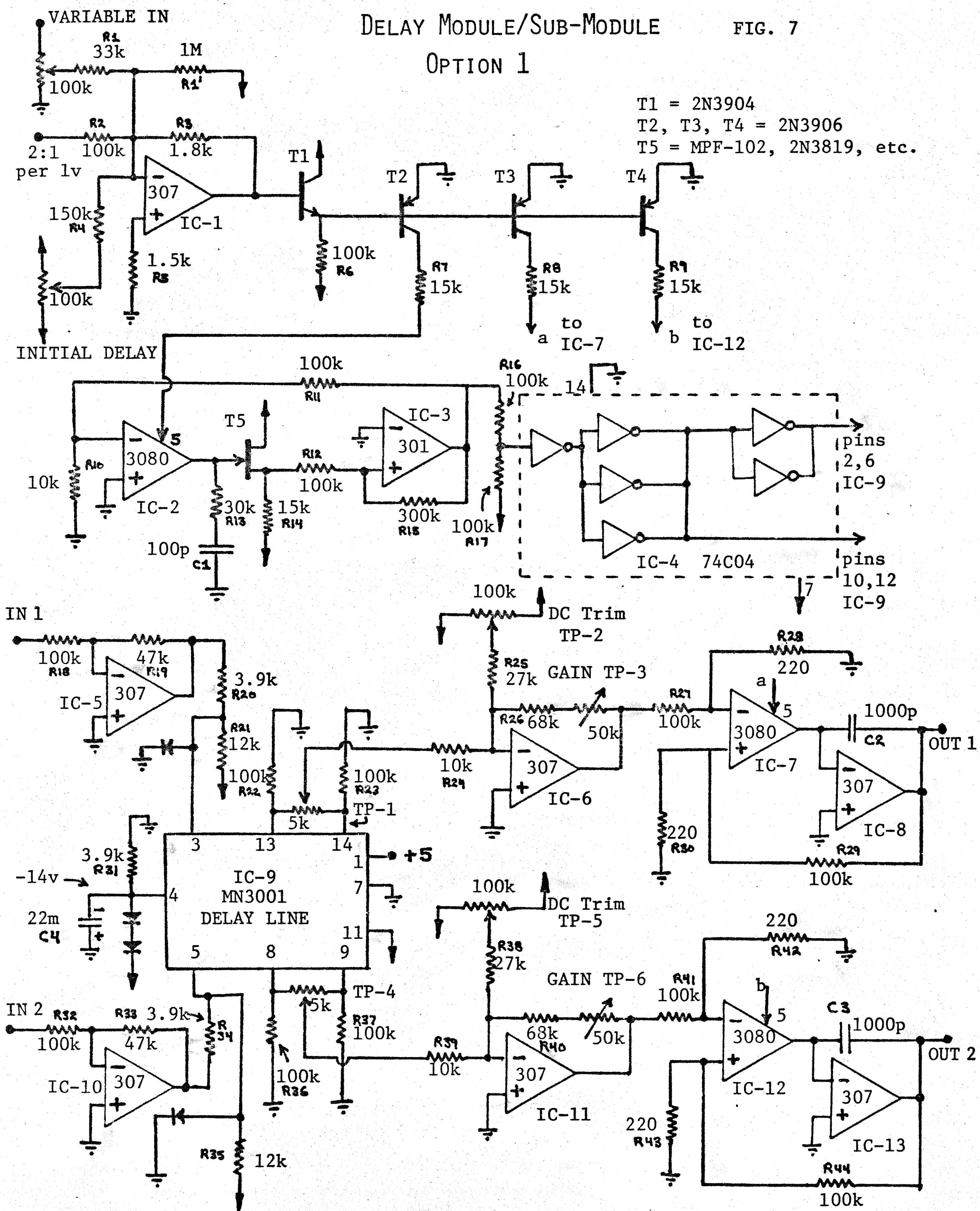
[3] In EN#68 (2) we suggested that the preferred way of driving the clock lines of the MN3001 chip was to use three sections of a 74C04 inverter in parallel. Also, we obtained the original two phase clock signal using a flip-flop in the EN#68 circuit. Here we just use the square wave directly. However, we use one section of the inverter chip to square up the output of IC-3 since the 301 is not as fast as the CMOS. This leaves two sections of inverter for one clock, and three for the other clock. This seems to work just fine. Note that the 74C04 (IC-4) is powered between ground and -15.

[4] Previous circuits required AC coupling capacitors and a bias supply of -3.7 volts for inputting signals to the MN3001. Here we use a voltage divider driven from the input op-amp to supply the proper DC level. This divider (R20 and R21) attenuates the signal by about 0.75, but since we have more than enough signal to begin with (the synthesizer signal levels are at least ±5 volts while the signal level on the MN3001

DELAY MODULE/SUB-MODULE

OPTION 1

FIG. 7



should only be about 1 volt amplitude), this provides no problem. A signal level of ±5 at the input (In 1) will drive the delay line chip at near its full dynamic range. The diode on pin 3 of IC-9 prevents this pin from going positive should the output of IC-5 go too high. Note that this input stage achieves DC coupling of the input.

[5] IC-9 is an MN3001 dual 512 stage bucket brigade delay line. It is a delicate and expensive MOS device and should be treated with care. It should be the last chip installed, and usual MOS handling precautions should be observed. The clock driver (IC-4) drives pins 2,6,10, and 12 of this chip. The clock lines should be short and of heavy gauge. We suggest mounting the 74C04 and MN3001 as close as possible. The chip requires four voltages (+5, ground, -14, and -15). The -14 volt supply is shown connected to pin 4. The diodes in this supply (and in the rest of the circuit) are just small signal silicon diodes such as 1N914 or 1N4148, etc. If you do not have +5 as a standard supply voltage, this can be obtained from the +15 line using a LM309H regulator as an "on card" supply. The +5 and -15 volt lines should be bypassed to ground with 22 mfd capacitors. The bypass on the +5 line should be made as close to pin 1 as is possible.

[6] The output stage begins with TP-1 on pins 13 and 14 of IC-9. This is a 5k trim pot which serves to null out as much clock noise as is possible. The signal is then amplified back to unity gain relative to the input (In 1) by IC-6. IC-6 also serves to DC shift the output of IC-9 which would otherwise be pinned by the -3.7 bias level. This DC shift is set by TP-2, and the gain is controlled by TP-3. Note that the gain control (TP-3) interacts with the DC offset (TP-2). Thus, one must first set TP-2 to an approximate value, then set the gain to unity using TP-3, and finally reset TP-2 for zero DC offset.

[7] IC-7 and IC-8 form a single-pole VCF which is the same one used in the variable slope filter above. This section is driven by T3, and thus tracks the VCO which clocks the line. This filter is responsible for removing clocking noise. The exact positioning of the filter relative to the clock is determined by C2 and was set by trial and error in this case. This can be adjusted to your own needs.

As with any discrete time system, delay time and usable bandwidth are related. For the MN3001, this is shown in Fig. 8 at the right. The delay time is related to the number of stages as:

$$T_d = \frac{\#\ \text{Stages}}{2\ f_{clock}}$$

The usable bandwidth is given as approximately 0.3 times the clock frequency. Also shown in Fig. 8 is the control voltage for the VCO that gives the time delay shown. As an example, suppose you want to know the max. delay for the dual chip (1024 stages) for a full audio bandwidth (15 KHz). Following the dotted line you see this requires a clock of 50KHz and gives a delay of 10 ms.

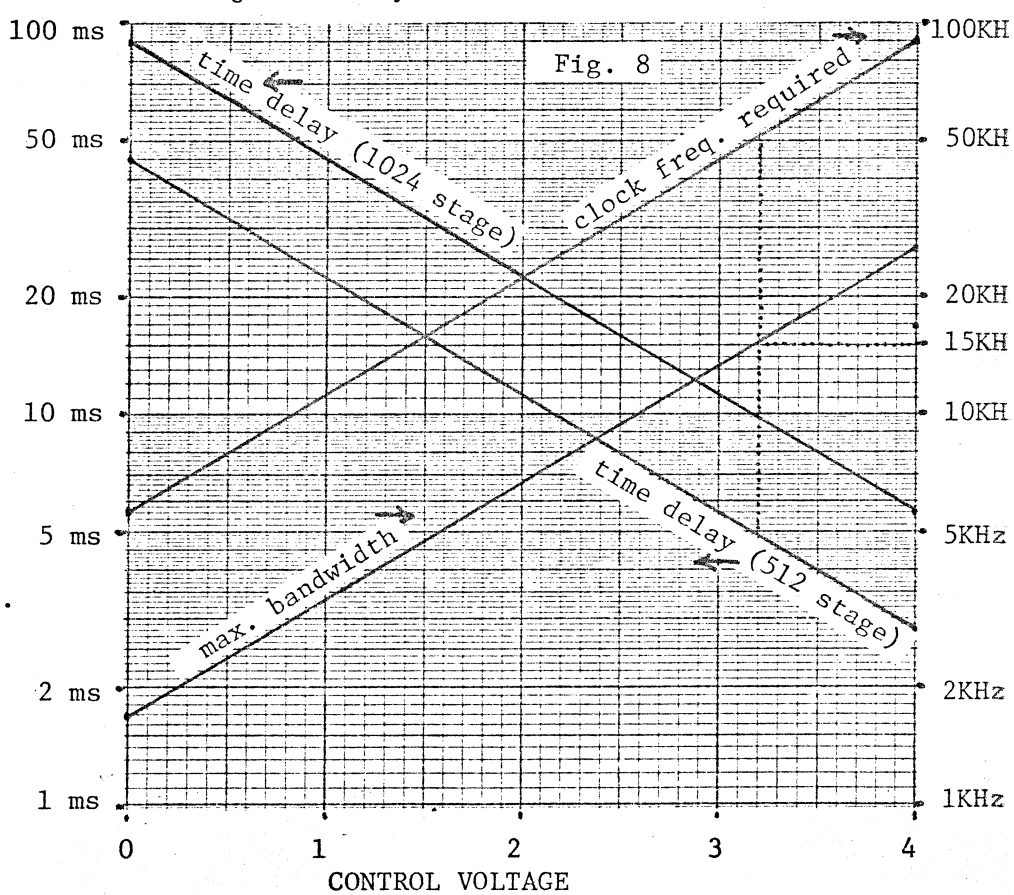


Fig. 8

Since the delay line is DC coupled, it might seem that there is no reason why we can't reduce the clock frequency, accept the reduction in bandwidth, and pass very low frequencies and envelopes through the delay. The VCO will certainly clock down to below 10 Hz (corresponding to negative control voltages in this case). Actually, there is quite a bit of leakage that will reduce the gain of the circuit for lower clock frequencies, but it is possible to clock the line down to 100 Hz. With this, DC signals and envelopes can be delayed up to about 5 seconds. However, this will be plagued by DC offsets and attenuation. It may be possible to use the control voltage to the VCO to correct for these, but we did not try this.

There are numerous alterations that can be made to the circuit. As shown, we have two lines which operate in parallel. These can of course be cascaded. It is also possible to cascade without decoupling the lines through an external link. To do this, everything is removed from pins 13 and 14 except for R22, and everything is removed from pin 5. Pins 5 and 13 are then connected. You now have 1024 stages of delay with a simpler circuit, but will not have a dual system. The output in this case comes out of IC-13, and T3 and R8 will not be needed.

Some users may want to add more delay line chips to the module. If more are added, each chip should have its own clock driver chip (74C04). Each half chip may have its own input and output circuitry (as with IC-5, 6, 7, and 8) or the direct cascade method may be used. Up to 5 chips may be cascaded. A tapped delay line with 11 outputs could be implemented with 5 cascaded chips if each had its own output circuitry. An output chip such as IC-8 could take the place of an input chip such as IC-10 if all lines are internally cascaded.

There are many alterations that would use additional VCF's. These VCF sections can be controlled using more transistors such as T2, T3, and T4 all driven off T1. For example, another stage of filtering could be added to the outputs. It might also be desirable in some cases to place tracking filters on the inputs as "guard filters" for the delay line. We left them off because the input signal for synthesizer modules is usually well known. Probably, guard filters would need to be 2nd order or higher. Also, aliasing might be a useful special effect with these lines.

For simplicity we let more positive control voltages correspond to shorter delay times. If you think in terms of clocking of the line and VCO's, this makes sense. However, if you prefer more positive voltages to correspond to larger things (longer delays), you can put an inverter in series with IC-1. To do this, R3 would be made 100k, R5, would be 22k, and the inverter stage would look like IC-1 does at present but with only R2, R3, and R5 attached with their values as shown in Fig. 7. To make a long story short, you just invert first, and then attenuate (see Fig. 9).

If you use the module/sub-module as a synthesizer module, you will be able to achieving phasing (flanging) effects by summing delayed and original signals in a mixer module. If you want to use it as a sub-module in a phasor, etc., you can get an idea of the structures from looking at EN#61 (18). If there is enough interest, we will write some notes on this.

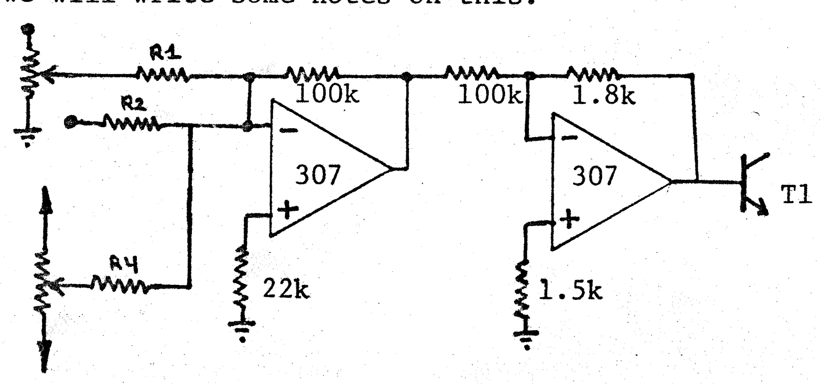
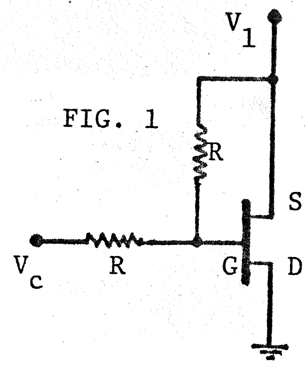


Fig. 9

Alterations to Fig. 7 so that more positive voltages correspond to longer dealys.

READER'S CIRCUITS:

FIG. 1

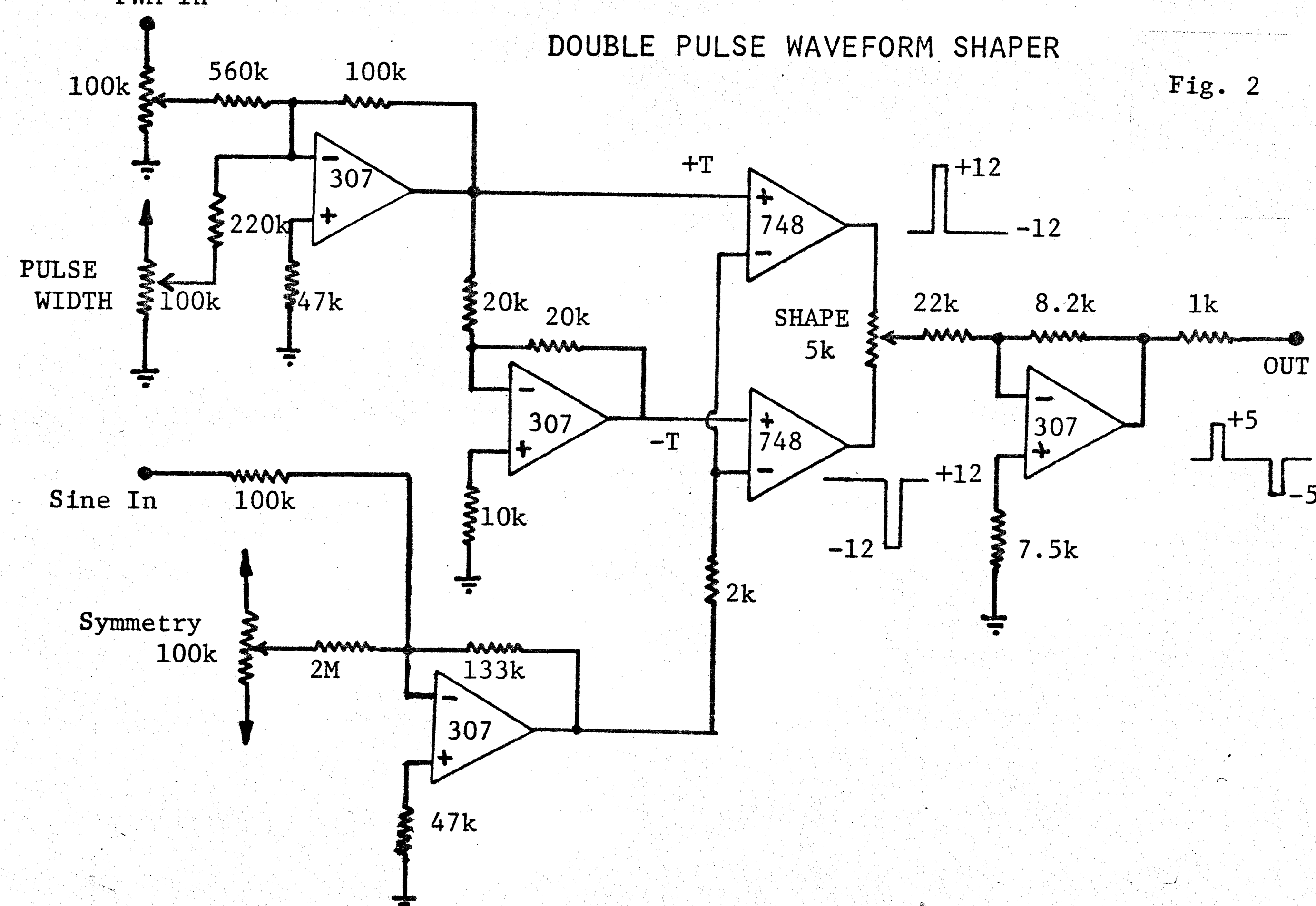


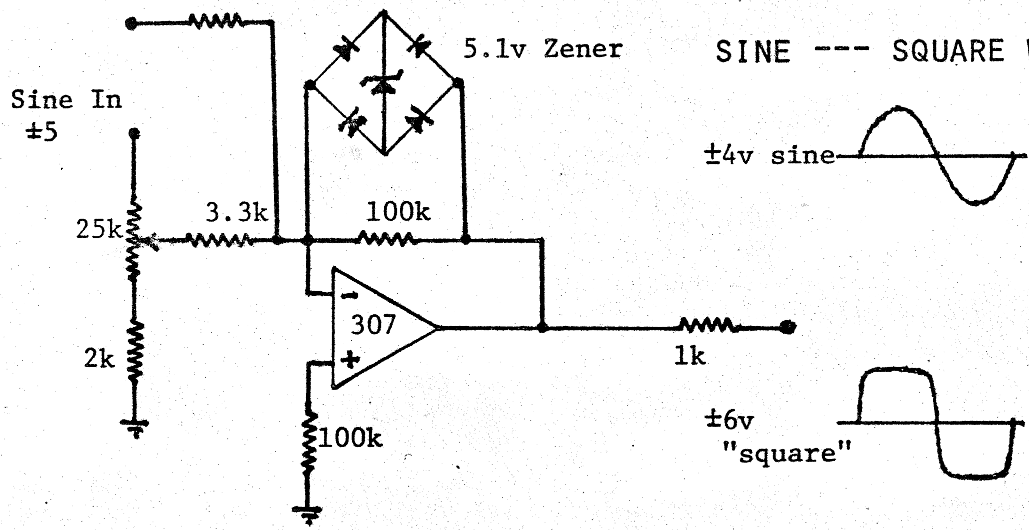
Jacob Moskowitz suggests that when FET's are used as voltage-controlled resistors that the feedback arrangement shown in Fig. 1 be used as this will improve the dynamic range by a factor of 5 to 10 over the case where the V_c terminal is applied directly to the gate. The arrangement permits signals of about 1 volt p-p and any distortion you get is "soft" and symmetrical. Jacob indicates that varying the amount of feedback between S and G does not make much difference, and that both resistors equal, and in the range of 680k to 4.7M, gives good results. If V_c is bypassed to ground and only AC control is important, 2R becomes the maximum resistance of the VCR, so really only one extra resistor is needed, since most VCR circuits using FET's require a parallel resistor to set the maximum resistance. FET's with high pinch-off voltage give the best dynamic range. For example, 2N4091's work well.

Ian Fritz has submitted three waveshaping circuits which are somewhat different from those generally used, and thus should be of interest to persons using waveshape as a principal method of timbre control. The first circuit is shown in Fig. 2 and is a pulse-width modulator for a "double pulse" rather than the usual single pulse. The second circuit (Fig. 3) is a shaper circuit which squares up a sine wave without actually turning it into a square wave (it rounds off the corners). The third circuit (Fig. 4) can be used to either chop off the bottom half of a sine wave, or to cut back on the amplitude of the lower half, according to the setting of the control pot. This results in the generation of even harmonics which would be absent from many shapers (such as the one in Fig. 3 for example).

DOUBLE PULSE WAVEFORM SHAPER

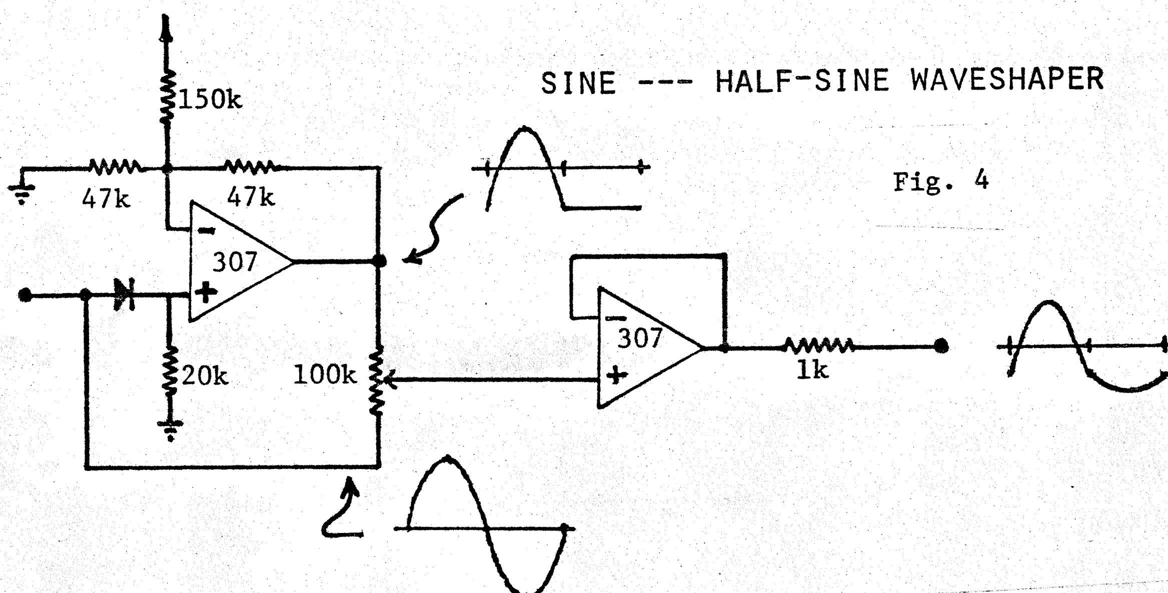
Fig. 2





SINE --- SQUARE WAVEFORM SHAPER

Fig. 3



SINE --- HALF-SINE WAVESHAPER

Fig. 4

CLASSIFIEDS:

FOR SALE: EMS Synthi AKS Synthesizer. Features matrix patching and built in digital sequencer. I'm interested in buying a Psychtone and JBL 075 tweeter. Also, am designing a polyphonic keyboard a la Polymoog, ARP Omni - but not exactly, and would like to correspond with others working along similar lines. Alan Rowoth, 179 E. Genesee St., Auburn, NY 13021.

FOR SALE: Complete set of Electronotes #1-72, including bibliographies and supplements. Need money for college. Best offer. D. Craig, 67 Hill St., Norwood, MA 02062

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