ELECTRONOTES 113

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GROUP ANNOUNCEMENTS

CONTENTS OF EN#113

At times in the past, we have had a bit of information come in one day and we were then able to write it into a newsletter going out the next day. In preparing this issue, we had Ian's material prepared and the first four pages were even printed when, while considering what to add after it, a letter came from Serge Tcherepnin. As the reader can see, this is the ideal companion piece for Ian's report. These two articles add greatly to what we need to know to apply these two new chips. As it turns out, there are subtle points to consider, and we can all be greatful for the insight and practical suggestions Ian and Serge have provided. It is probably the case that there is more dust to be stirred up with regard to these new chips, and there will be more time before it all settles.

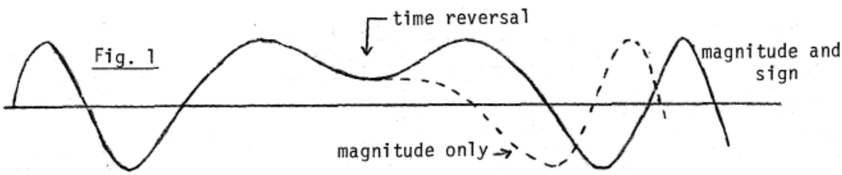
READER'S QUESTIONS:

▶ Q: I think I still do not understand through-zero frequency modulation? If you look at a negative frequency as a phase inversion, then why not just use one of your switched inverters when the control goes negative?

A: You are far from being alone here. Actually, it seems that just when you think you understand the problem, another "but what if" comes up. Let's go back a bit. Chowning in 1973 handled the problem by saying that "$-\sin(\theta)= \sin(-\theta)$", something that is surely true. Thus the idea that a negative frequency is a phase inversion. The first complication comes up when we realize that phase is relative, and must be relative to something. This fact you can appreciate if you suppose you enter a room where a sine wave appears on a scope face, and are asked if it is $\sin(\omega t)$ or $-\sin(\omega t)$. Since you have no reference phase, you have no idea [or whether it might be $\sin(\omega t + \phi)$ where ϕ is an arbitrary phase angle].

In the case of through-zero modulation, what kind of reference can there possibly be? Well - only the relative phase of the positive frequency to the negative one. Thus consider the following experiment. You are given a VCO which may or may not be capable of through-zero modulation. You give it a control voltage, positive let's say. Now you lower this control, slowly, and the pitch falls. About 15 Hz, you can't hear anything coming out. You keep going, slowly, through zero, and on to negative voltages. At about 15 Hz, you hear the tone, and the pitch starts to rise as you go to still more negative voltages. Did it go through zero or not? You can't tell, because you went slowly, and the dead zone between +15 Hz and the supposed -15 Hz was long enough that the ear's time constant expired, and the details of the original frequency were "forgotten." It is as though you left the room while the oscillator went through this dead zone. Thus we can see that an actual audible negative frequency effect will depend on the transition through zero, both in its details, and in its speed.

We are thus led to consider the "interface" of the negative frequency portion of the output waveform with the positive frequency portion. One means of achieving this is to use the following interpretation of a negative frequency: $\sin[(-\omega)t] = \sin[\omega(-t)]$. This is obviously true. What we have done is associate the negative sign of a combined phase factor ωt with t rather than with ω. Thus we are led to the idea that the transition to negative frequency can be viewed as a reversal of time. Thus we are led to the idea of using a reversing oscillator, one which in effect mirrors the waveform about the point in time at which the frequency passes through zero. A VCO using this interpretation uses the absolute value of the control voltage to determine the rate at which the oscillator would "normally" oscillate, while the sign of the control voltage determines whether the oscillator is running backward or forward. Keep in mind that "backward" or "forward" only have meaning where the transition occurs and only where the transition is rapid relative to the time constant of the ear. An FM synthesized waveform in this view thus looks like a sine wave of decreasing frequency. At zero control voltage, the oscillator stops, and as the control voltage goes negative, the waveform continues, but in a direction reverse to that at which it was going. If only the magnitude of the control is monitored, the waveform would continue as it was (see Fig. 1 below):



Next comes another "but what if" the modulating waveform is rectangular so that the change from positive frequency to negative is instantaneous. What happens, or what should happen. In this case, the instantaneous inversion may make the most sense. Clearly, a time reversal is not the same thing at all, or at least a jump in time is required.

The fact that there is an apparent difficulty here perhaps indicates that one of the two interpretations is incorrect, or that neither is correct totally, perhaps being special cases of a more general interpretation. That the latter may be most likely is perhaps indicated by the study of a waveform such as $\sin[\omega_o(\sin\omega_m t)\,t]$, which many would consider the proper FM equation (for zero carrier). You might suppose that the changes in direction such as that in Fig. 1 would occur when $\sin\omega_m t$ changes sign. [This is what is assumed in the reversing oscillator]. They don't. They occur when the total phase $\omega_o(\sin\omega_m t)t$ changes direction. This depends on how the ever advancing t is modified by the $\sin\omega_m t$ term. A general interpretation based on this consideration might lead to the proper model.

There are additional considerations. Computer FM methods and reversing VCO's both show the apparent time reversals of Fig. 1. Also, note that regardless of whether we are doing FM or something else, it is the result that counts. Thus the reader should understand that his question is one which all of use should be asking. Perhaps some other readers can offer suggestions.

SWITCHABLE VCA/BALANCED MODULATOR CIRCUIT BASED ON THE LM/XR 13600 TRANSCONDUCTANCE AMPLIFIER:

-by Ian Fritz

I. INTRODUCTION

The module described in this article serves as both a VCA (voltage controlled amplifier, or two quadrant multiplier, or 2QM) and as a balanced modulator (four quadrant multiplier or 4QM), with the mode of operation selected by a switch. The design has evolved from previous designs using transconductance amplifiers, so much of the circuitry will be familiar to Electronotes readers. New features of the design include the following: 1) The new 13600 dual transconductance amplifier [1] available from National Semiconductor and Exar is used for improved signal-to-noise performance (∿10 db improvement) 2) A modification of the simple 4QM design with a single transconductance amplifier - which is given in the manufacturers' application notes [2] and which was discussed by Hutchins [3] - is employed. This modified 4QM design has better accuracy than the original circuit. 3) The design is efficient in that it only used 1-1/2 chips: a quad op-amp plus half of a 13600 4) Finally, the voltage-to-current converter for the control input is somewhat different from previous designs. A "bias" control that selects a weighted average of the Y input and a fixed bias voltage replaces the usual "initial gain" control of previous VCA designs, and also serves tc unbalance the input in the 4QM mode to mix in the carrier signal.

II. DESIGN OF THE TRANSCONDUCTANCE SECTION (X INPUT)

To begin with, it will be assumed that the amplifier bias current I_b driving the 13600 is given by the following expressions, where V_y is the control voltage, or the Y input voltage:

$$I_b = \frac{V_y + 5}{20} \qquad \text{(4QM)} \qquad \text{(1a)}$$

$$I_b = \frac{V_y}{10} \qquad \text{(2QM)} \qquad \text{(1b)}$$

For the 4QM case V_y is between -5 and +5 volts, and in the 2QM case V_y is between 0 and +5 volts. Thus I_b covers the range 0 to 0.5 ma in both cases. Design of the circuitry to produce I_b will be discussed in Section III.

A fairly extensive study, both theoretical and experimental, of the operation of the 13600 has been undertaken. The details of this investigation will not be given here, but the major results will be briefly summarized. First it must be pointed out that the analysis used by Hutchins [1] is not quite correct. This is because the current through the input resistor of Figure 6b of Reference 1 is (approximately) equal to <u>twice</u> the current I_s of Figure 3 (same Reference). The assumption that these currents are equal is what leads to the measured output voltage appearing to be off by a factor of two.

It must be emphasized that the diode predistortion available with the 13600 does not provide exact compensation of the input nonlinearities, as does the input structure of the SSM chips [4]. The manufacturers only claim about a 10 db improvement in signal-to-noise ratio (referred to 0.5% harmonic distortion), and it does not appear possible to do much better. This means that the inputs can be driven to about ± 30 mv as opposed to the ± 10 mv used without predistortion. A design utilizing ±50mv of drive was developed, but it requirec an active current source for the diode bias, and had only about 10% headroom before the onset of hard limiting.

It is also found that as large a diode bias I_d as possible should always be used. The suggestion [1] of increasing the input drive by decreasing I_d is not recommended, as this increases the nonlinearities of the input structure. The present design uses I_d = 1.5 ma. The absolute maximum allowed value is 2 ma.

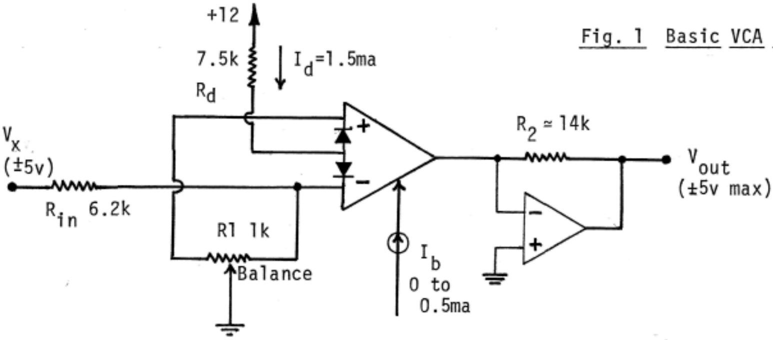


Fig. 1 Basic VCA Structure

The basic circuit for the 2QM is shown in Figure 1. The value of R_{in} was determined by solving the equations for the input network, and R4 was determined empirically to give unity gain. It was found that R4 was smaller than predicted, possibly due to non-ideal transconductance of the chip (a range of about ± 30% is specified by the manufacturers).

As mentioned, the 4QM is a variation of the circuit described previously by Hutchins [4]. That circuit is shown in Figure 2. Analysis of Figure 2 shows that the overall gain is determined by the ratio of R_M to R_y, which in turn is determined by the balance condition for the multiplier. The situation is different if the setup of Figure 3a is used. Here the Y voltage input is converted to a current by active circuitry. The VCA design corresponding to Figure 3a is given in Figure 3b. A fairly simple analysis yields the following neat result: For the 2QM, assume that R_{out} is chosen for unity gain at maximum bias current, so that $V_{out} = V_x V_y/5$. Then the 4QM will be properly balanced and have $V_{out} = V_x V_y/5$ when $R_M = 2 R_{out}$. Now it is easy to see that a suitable 2QM/4QM could be made based on Figure 3: the bottom end of R_{out} could be switched either to ground (2QM) or to a resistor whose other end is connected to V_x (4QM). Readers who have been following this rather sketchy argument carefully will have noticed that the two V-I converters in Figure 3 are different, so some switching is required there also.

A problem, whose origin is not understood, arises with the circuits of Figure 3. Both circuits are quite nonlinear as a function of I_b. This nonlinearity (~20%) does not appear to depend on the value of V_x (V_x is fixed and I_b is varied), nor on the value of I_d. The nonlinearity does not occur, however for the VCA of Figure 1, and therefore appears to be associated with the output structure. Fortunately, there is a 4QM structure analogous to Figure 1, and this is shown in Figure 4. The second op-amp in this circuit is interesting in that it simultaneously works as a current to voltage converter for the output of the 13600

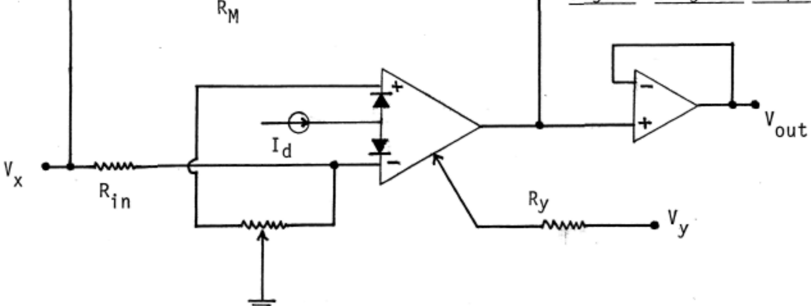


Fig. 2 Original Simple 4QM

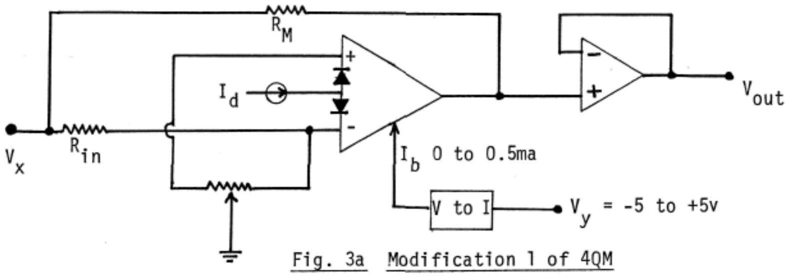


Fig. 3a Modification 1 of 4QM

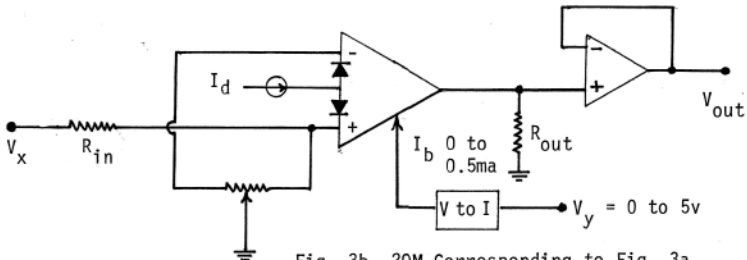


Fig. 3b 2QM Corresponding to Fig. 3a

and as an inverting amplifier for the input voltage. Analysis indicates that the 4QM is balanced and has $V_{out} = V_x V_y/5$ when $R_a = R_b = 2R_2$, where R_2 is shown in Figure 1. Thus we can change the 4QM of Figure 4 to the 2QM of Figure 1 simply by moving the left-hand end of R_a from the input of the circuit to the output! (Again, of course, we have to change the V-I converter at the same time).

III. CONTROL SECTION (Y INPUT)

For the Y input circuitry it is necessary both to implement Eqns. 1 and to provide some sort of manual gain control. A means to unbalance the 4QM is also a useful feature. The design adopted here is shown in Figure 5. The two resistors R_A and R_B are equal. It is convenient to analyze the structure from the end backwards. The V-I converter produces an output current I_b ranging from 0 to 0.5 ma in response to an input V_a ranging from 0 to -5 volts. If the control R_B is in the CW position, then V_a is -5 volts, independent of V_y.

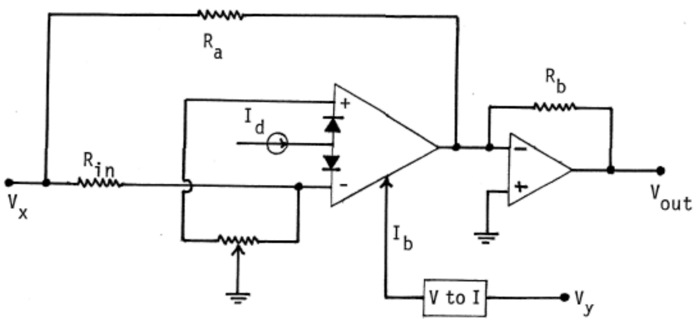


Fig. 4 Modification 2 of 4QM for Improved Accuracy

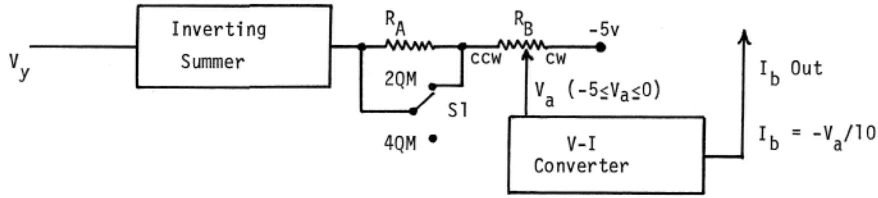


Fig. 5 Schematic of Control (Y) Section

Thus the transconductor is turned on to full gain and the output of the module is V_x. (This statement is true independent of whether the X section is in the 2QM or 4QM mode). Now suppose R_B is in the CCW position. There are two cases to examine. First suppose that the switch S1 is closed (2QM position) and the transconductor is in the 2QM mode (Figure 1), so that the unit is being used as a VCA. Values of V_y from 0 to 5 volts are of interest, so V_a clearly ranges from 0 to -5 volts. Thus the transfer function of Figure 5 is

$$I_b = \frac{V_y}{10} \tag{2}$$

as required by Equation 1a. The second case is for 4QM operation with S1 open. Here V_y ranging from -5 to +5 is of interest. Since R_A and R_B are equal, V_a is the average of $-V_y$ and -5 v. Thus $V_a = (1/2)(-V_y-5)$, and

$$I_b = -\frac{V_a}{10} = \frac{V_y + 5}{20}, \tag{3}$$

as required by Equation 1b.

In general R_B provides a weighted average of V_y and -5 volts that allows the modulation or control input voltage to be "traded off" against a fixed bias. The advantage of this kind of control is that it is normalized in the sense that I_b is always restricted to its proper operating range.

IV. FINAL CIRCUIT

The final circuit is given in Figure 6. A couple of points need to be made here. A non-inverting buffer is used on the X input (OA1a) to preserve signal polarity through the unit. An inverting summer could be used if the overall inversion does not matter. The purpose of R9 is to offset the Y balance slightly when PC3 is in the full CCW position. This allows the control envelope to be burried slightly in the 2QM mode, and gives some range for the Y balance point in the 4QM mode. The correct position for PC3 can generally be set by ear. The circuit as drawn is set up for ± 12 v supplies. For ± 15 v operation R4 should be changed to 10 kΩ and R13 to 2 kΩ.

The circuit as indicted has a bit of distortion which can be seen on a scope, but which doesn't seem to have any significant audible effect. The circuit has smooth limiting and saturates at about ± 8 volts. Predicted values for R_a and R_b (Figure 4) are 38 k Ω, whereas the measured values are about 33 k Ω and 26 k Ω, respectively. The gains of the 2QM and the 4QM are slightly different (within about 15%), and TP2 may be set to give exact unity gain for one or the other, but not both. The setup procedure for the unit is as follows:

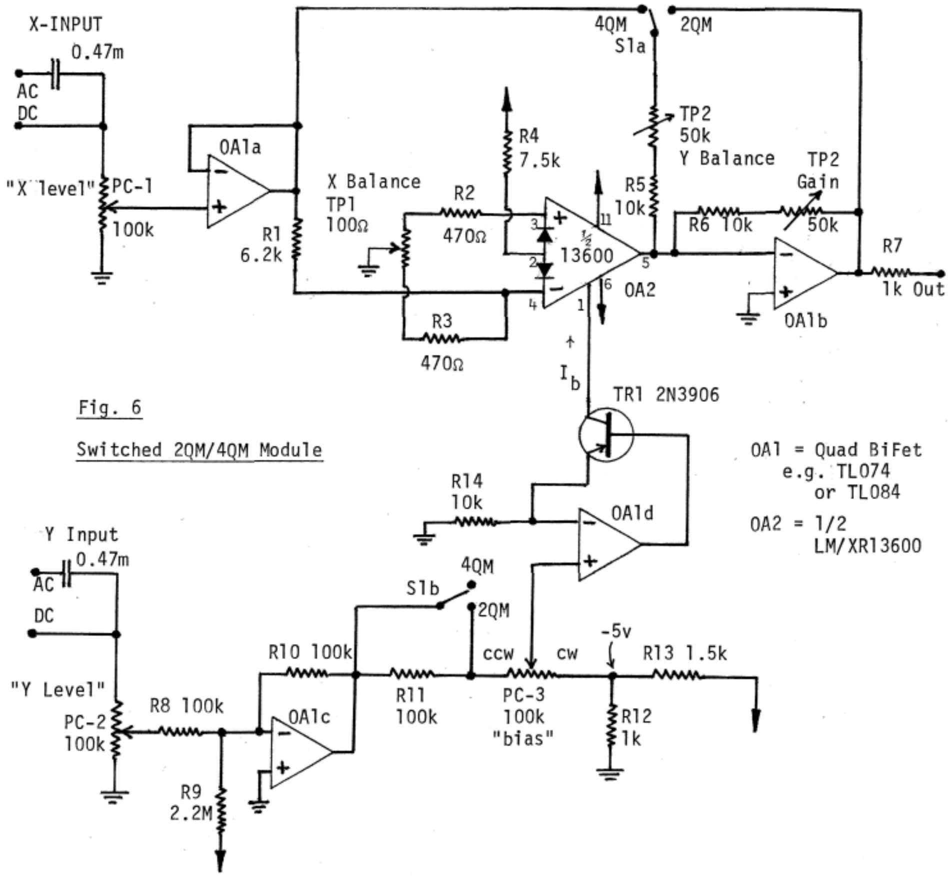


Fig. 6

Switched 2QM/4QM Module

Step 1.
S1 position ("mode"): 2QM
PC3 ("bias" control): full CCW
X input: 0 v
Y input: ± 5 v audio, or + 5 v pulse
Adjust TP1 (X balance) for minimum feedthrough

Step 2.
S1: 2QM
X input: ± 5 v audio
Y input: 0 v
Adjust "bias" control (PC3) to just below point where VCA starts to turn on. This is the "threshold" position and should be marked on the panel. This position represents the nominally correct Y balance position. Leave PC3 in this position for the remaining adjustments.

Step 3.
S1: 4QM
PC3: "threshold"
X input: ± 5 v audio
Y input: 0
Adjust TP2 "Y balance" for minimum feedthrough

Step 4. S1: 4QM or 2QM
PC3: "threshold"
X input: ± 5 v audio
Y input: + 5 v dc
Adjust TP2 "gain" for unity gain

Step 5. S1: opposite from Step 4
X input: ± 5 v audio
Y input: + 5 v dc
Check for approximately unity gain.

REFERENCES

1. B.A. Hutchins, Electronotes 107, 3 (1979).

2. Exar Integrated Systems, 750 Palomar Ave., P.O. Box 62229, Sunnyvale, CA 94088, Application notes for the XR-13600.

3. B.A. Hutchins, Electronotes 107, 13 (1979).

4. D. Rossum, Electronotes 67, 3 (1976).

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COMMENTS ON THE NEW 13600 AND 3280 OTA DEVICES:

-by Serge Tcherepnin

EDITOR's NOTES: This material is based on a letter from Serge, and since it was both informative and readable in its original form, no attempt was made to make it into something more like an article. ---Bernie

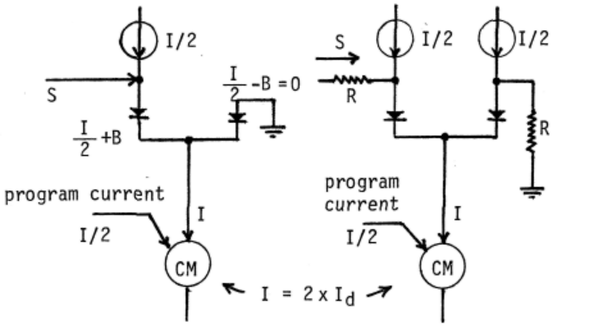
I read EN#107 ("The New OTA's; The CA3280 and the LM13600") with interest since so little is written about that particularly slippery device, the 3280. Also, your discussion of the 13600 was especially interesting to me, since I didn't even have a data sheet concerning this device.

When I came to the part in the notes in which you reveal the discrepency between the experimental vs. theoretical results, I thought to myself, aha, here's another victim of the 3280's poor data sheet (as I was one myself, and had exactly the same discrepency in my tests). I thought I'd write you a short note to explain the matter. However, I decided to take a look at the bad results regarding the 13600 also. Amazingly, the causes of the discrepency are entirely different for each of the devices, though the net result was the same.

In the case of the 3280, the cause is the fact that twice the I_d is reflected, and thus each of the diodes sees ½ of 2 times I_d. Where is this found in the data sheet? The RCA engineers should be strangled (I guess what will save them from my wrath is the fact that they invented and marketed such a marvelous device as the 3080!). By the way, the current mirrors built into the 3280 are not the garden variety found in the 3080; rather they have been optimized to work at much higher currents than the old 3080. Thus there is some fancy (unspecified in the data sheet) processing and circuitry which makes the current mirrors provide about 90% ratioing as concerns the diode programming, and 80% for I_{ABC} vs. output current. More on this later.

Your simplified equations are perfectly accurate for the current sourced diode network; they are not so for the 13600 diode network as hooked up in your experimental circuit. This can be quickly verified by asking the following question with reference to the various circuits shown (top of next page): what values S and B are needed for the I-B current through one of the arms of the diode network to approach zero? (See figures for explanation).

An interesting perception results when the complete equations governing the RCA vs. the National biasing schemes are developed (which includes the logging terms).

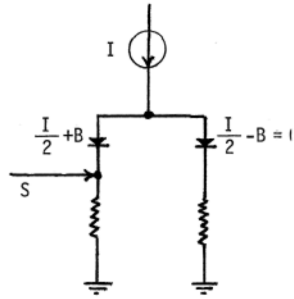


3280 Single Ended

$S = B = I/2$

3280 Differential Input

$S = B = I/2$



13600

$S = 2B = I$

Thus equation (4) of EN#107 for this case should read

$$\frac{\frac{I_d}{2} + \frac{I_s}{2}}{\frac{I_d}{2} - \frac{I_s}{2}} = \frac{\frac{I_{ABC}}{2} + \frac{I_{out}}{2}}{\frac{I_{ABC}}{2} - \frac{I_{out}}{2}}$$

It is this: that the distortion rises quicker with increasing inputs for the National scheme versus the 3280. With 9/10ths of the maximum input signal current (equalling approximately I programming), distortion for the 3280 current mode biasing scheme is about 0.3%. For 500 ohm terminated grounded scheme such as the 13600, distortion at that level signal is more than 8%. A little further analysis showed that as the terminating resistances were increased, distortion decreased proportionally. This fact makes sense, of course, seeing the direction of the change toward pure current sourcing. The least distortion would be had with a 3280 current biased network fed from a pure current source signal: the 0.3% distortion is mainly due to the fact that a resistor was assumed sourcing the input current. Thus it is with the National part that using large resistors to a negative voltage would lead to better distortion figures.

I could write a volume concerning the 3280 which I consider one of the slipperiest devices (full of non-obvious quirks...) that I have seen. In the balance, I consider it a better work of the IC designer's art than the 13600, mainly because of the interdigital input transistor structure which ensures very low noise (about as good as the 381) and excellent offset voltage matching over changes of I_{ABC}. How often have we had to de-select a 3080 for extremely bad offset degradation?

Typically, a 76db S/N ratio can be achieved (ref 0 db) with the 3280. 3dbs can be gained if the two amps are used in parallel for a single VCA; and another 3 dbs for a smashing 83 db S/N can be had by paralleling two entire packages! (RMS logic). The CV rejection is also similarly improved. However, to achieve the S/N, a careful trim of the I_d max should be made to obtain the necessary I_{ABC} minimum.

For maximum CV rejection, differential resistors <u>and</u> DC coupling must be used, since the change of input bias current with changing I_{ABC} will otherwise wreak havoc. And of course, the larger I_d, the better these biasing currents are swamped. We've realized better than 60 dbs of CV rejection; on the other hand, when CV rejection is not primordial, a single ended input resistor, with the second amplifier input tied to ground, is very convenient because it allows capacitive coupling.

Due to parallel biasing of the internal PNP mirrors directly by I_{ABC}, it is not possible to inject a current into the available emitter pins, (unless an I_{ABC} of similar magnitude is also input. This is one of the lousy features of the 3280 which makes direct drive by an external exponential source impossible, alas....

My tests revealed that the inputs to the mirrors have very tight matching; thus the terminals can be tied together quite simply (as for use with VCF's).

As you may be aware, the equation for the 3280 is 16 times I_{ABC} : G_m. It turns out however, that the old equation for the 3080 applies in a modified form; 19.2 times $I_{out\ max}$: G_m. The consequences of this fact are beneficial to whomever wishes to design VCA's and to know what peak output voltage one might expect for a given G_m. No doubt this is why you discovered in EN#107 that I_{out} was 20% low, i.e., 16:19.2 low.

The 3280 is not as well speced with regard to input and output isolation and leakage as the 3080; thus is not advisable for sample-and-hold circuits.

My tests also reveal that the reflected I_d is 1.8 the programming current. Another good feature is that the PNP currents in each of the two branches match extremely closely (within 2%). On the other hand, the sum of the PNP currents is not precisely equal to the NPN source, and thus a common mode difference exists. The worst of this is that it can be positive or negative.

Other facts:

V_{ni} is 3.4db better with the 3280 than the 13600 at I_{ABC} = 150 na

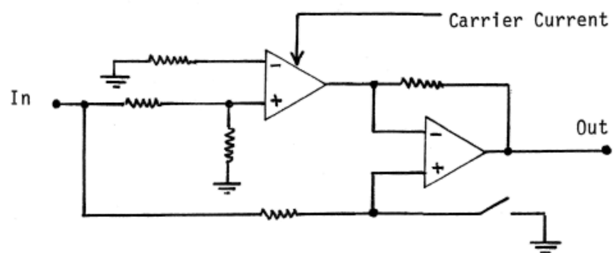
V_{ni} is 4.6db better with the 3280 than the 13600 at I_{ABC} = 500 na

V_{ni} is 4 db better with the 3280 than the 13600 at I_{ABC} = 1 ma

For the 3280, V_{ni} is -117.4 db/0 ref at 150 na (BW = 16 - 16kHz)
-120.4 db/0 ref at 500 na
-122 db/0 ref at 1.5 ma
-127 db/0 ref at 5 ma

Flicker noise however increases in the milliamp region

One last word: I appreciate reading about your simple ring modulator. In fact, I designed a related circuit many moons ago (1973)



This had the advantage that I so configured the circuit to allow using the PC board for either a ring modulator (with VC changeover from ring through amplitude to no modulation) or as a VCA (switch above closed) with exponential and linear CV inputs. The trick for achieving both circuits was using a PNP long tail multiplier to supply the I_{ABC}. Somewhat sly was the fact that in my realization the trimming of signal cancellation was left to the user (in form of the knob which controlled whether the circuit behaved as a ring, amplitude, or non-modulator). Sly, because without diode predistortion as in the 3280, G_m shifts with temperature making signal cancellation a changing thing. The 3280's advent was the cause for my redesigning the module. In fact, I find that 3080-like transconductors are far superior to true ring modulators (possibly because there are fewer internal V_{BE} cancellations needed). Superior because carrier suppression can be far superior. My new ring modulator, however, is another breed altogether. Carrier rejection is better than 78 dbs down / 0db ref. This uses two 3280's. Two of the amps are used for compression of the input signal, and expansion of the ring modulated signal. Very economical since the same precision rectifier controls both equally! (With a small averaging cap since distortion cancels).

The other two amps are used 1) for the modulator and 2) to provide the I_{ABC} driving multiplier as needed to smoothly change the circuit from full ring through amplitude and non-modulation modes (manually or via VC). In this circuit, nulling is <u>not</u> left to the user! It's really a pleasure to hear very distortion free ring modulation at all signal amplitude levels: very, very clean, and with no squelching glitch as in the Moog-Bode module. (Compression and expansion are other great uses for the 3280's good linearity).

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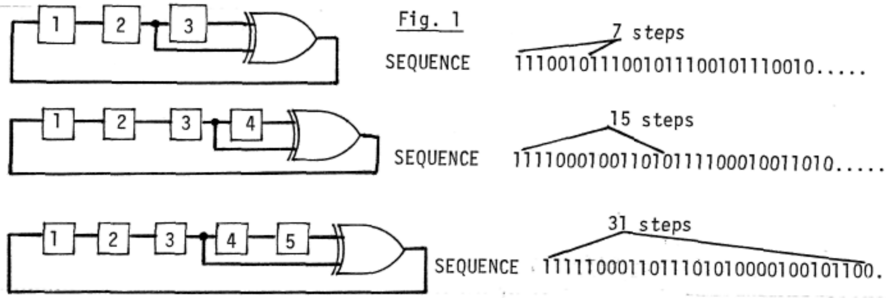
<u>SPECTRA OF SOME SIMPLE SHIFT-REGISTER SEQUENCES</u>:

-by Bernie Hutchins

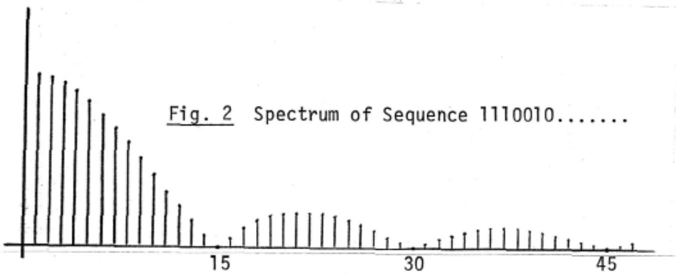
In connection with some work I am doing that has nothing to do with electronic music, I needed to find out how strong the fundamental component was in simple Pseudo-Random Binary Sequences (PRBS). This is simple, and the result is also simple, but I know that we have never looked at it before in these pages. The result while simple, is quite interesting.

The PRBS generators we will look at are the three stage, four stage, and five stage ones shown in Fig. 1 below:



<u>Fig. 1</u>

Basic information on PRBS generators can be found in the <u>MEH</u> Chapter 5h, and in EN#64 and EN#76. The basic idea is that you can choose proper taps from an n-stage shift register, go through an exclusive-OR process, feed back, and obtain a maximal-length sequence of 2^n-1. Thus we see a seven-step, a 15-step, and a 31-step sequence. Very long PRBS generators make excellent noise sources. Short sequences such as the ones above are so short that if repeated rapidly, an audible pitch at the repeat rate of the sequence is achieved. We can easily obtain the spectra of the sequences from their Fourier Series (by a simple method and program outlined in AN-160. Put this all together, and we get a simple result, of which Fig. 2 for the 15-step sequence is an example:



<u>Fig. 2</u> Spectrum of Sequence 1110010.......

The most striking thing about Fig. 2 is probably that you think you have seen it before. In fact, it is exactly the same in relative harmonics (except DC) as a pulse waveform with duty cycle 1/15. Note the Sin x/ x shape of the spectral envelope, and missing harmonics at multiples of 15. By extension, we can imagine that very long sequences have very close harmonics, none missing until some very large number, and a fairly flat top coming out from zero. Thus we can understand basically why the PRBS gives a white noise output under appropriate conditions. Based on the results of a study of the sequences in Fig. 1, we can suggest the following general findings:

→ The spectrum of a PRBS from an n-stage shift register (2^n-1 sequence steps) is the same as that of a pulse with duty cycle 1/(2^n-1). The corresponding pulse has amplitude:

$$2^{(n/2 - 1)}$$

and phase is not taken into account here.

Obviously what this means (at least) is that an actual 2^n-1 length PRBS is not of much use in timberal synthesis, since a pulse is just as good. It does, at the same time, add some insight into what a proper noise producing PRBS is like.

This is not to say that all shift register sequences are useless for timberal synthesis - quite the contrary is true. [See for example R. W. Burhans "Pseudo-Noise Timbre Generators", J. Aud. Eng. Soc., Vol. 20, No. 3, April 1972, pp 174-184]. For example, if we use taps on stages 4 and 5 of the five-stage shift register of Fig. 1, we get a 21 step sequence (111110000100011001010....) instead of the 31 step PRBS. This 21 step sequence has the spectrum of Fig. 3, which is not that of a 1/21 duty cycle pulse, but does still have some obvious pulse-like envelopes in its structure. Thus we can expect to find a variety of spectral shapes from non-PRBS shift-register sequences.

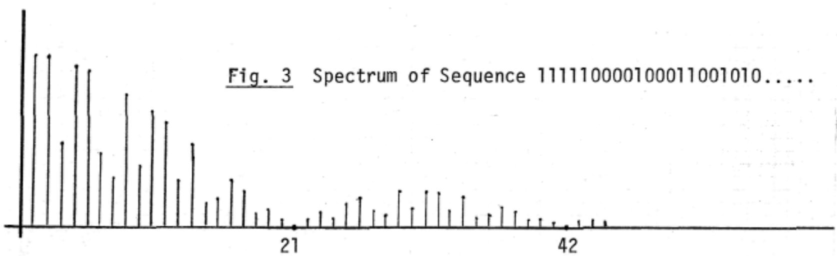


Fig. 3 Spectrum of Sequence 111110000100011001010.....

It is difficult not to suppose that we know the reason why the PRBS has a pulse-like spectrum. We just consider the PRBS to be a series of random pulses, take the spectrum of each one (which is a pulse spectrum), and add them up. Sorry. You are not allowed to do this in general. You can't add in the time domain (sum of random pulses) and also in the frequency domain (sum of spectra). Why does it work here - at least in some way? There must be a simple explanation.

Well, I see I am running out of room, and could give that as an excuse for not giving the explanation. However, in truth, I'm not at all sure, although it would seem to have something to do with the correlation properties of the PRBS (the auto-correlation is a pulse). Who can explain it?

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