A PROGRAMMABLE CORRUPTED CHANNEL SIMULATOR

by

DAVID R. ESTEN

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Signature redacted

Signature of Author..
Department of Electrical Engineering, 10 May 1972

Signature redacted

Certified by..
Professor B. Blesser, Thesis Supervisor

Signature redacted

Accepted by..... ....
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ABSTRACT

This paper presents the results of a project to build a programmable corrupted channel simulator. The simulator is an analog device intended for use in a 500 KHz band at low frequency.

The types of degradation available are discussed at some length and the motivation for choosing them is explained. The five types of disturbance adopted (wideband noise, amplitude fading, swept tone bursts, impulse noise, and frequency shifting) are programmable in some twelve possible combinations.

A complete set of circuit diagrams is given and the individual circuits may be correlated by reference to a set of system block diagrams. Some discussion of failed experiments is provided with explanations of circuit operation. Results of tests on the complete simulator in operation are presented and a retrospective look at the project as an educational experience is included.

PREFACE

In an effort to ease the burden of the reader, an informal narrative style is employed throughout this paper. A bare minimum of mathematics is presented to help avoid dryness and to try to give some of the flavor of the author's experience with the project. Illustrations are liberally used and, at one time or another, each circuit of the simulator is diagrammed. Thus, it is hoped, that the simulator is explained in an easily digestible form.

The author would also like to express his gratitude to his advisor, Professor Blesser, who suggested, supported, and supervised this project. Thanks are also in order to the author's wife who was awarded the contract to type this paper.

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I. INTRODUCTION

The project here described grew out of twin desires. The author felt the need of experience in circuit design and his advisor, Professor Blesser, required the construction of various black box systems for use as teaching aids in a formal laboratory course dealing with audio frequency communications. One of the modules required was a corrupted channel simulator which the students could employ to compare the merits of various communications schemes (e.g., AM, FM, PCM) under reasonably controlled conditions while operating at frequencies in the neighborhood of 100 KHz. Thus, the author undertook the design and production of a programmable corrupted channel simulator.

It was decided that the simulator should provide a number of types of degradations for the channel which would be externally selected by the user, i.e., be programmable. Further, the alignment procedure, if any, must be quickly and easily performed by a user to whom the device is little more than a black box. Finally, and perhaps the most severe constraint, the assembled simulator must be of fairly low cost. (A figure of $30.00 above cost of parts readily available within the Electrical Engineering Department was finally established.)

The ultimate goal, then, was to provide interested students with a means to acquire by experiment an intuitive feeling for the effects of at least some common types of degradation on any type signalling system he might care to build. It is hoped that, by use of the simulator, the student might gain some insight into the issues of analog vs. discrete signalling, coding vs. direct transmission, and system complexity without recourse to methods or experiences outside the scope of the normal undergraduate education.

In a sense, this project was unique in that there is no general use or demand for the simulator which is intended solely as a teaching aid, both for the author and subsequent users. The author knows of no other similar device, past, present, or future, although he would have certainly been grateful to be able to consult other works for general concepts and implementations.

II. GENERAL CONSIDERATIONS

In the implementation of any type of information system, the engineer is always confronted with the problems of finite budget, finite technology, and potential loss of information between origin and destination. Of interest to many electrical engineers is the problem of reliable waveform communication over a corrupted channel. Very often, an intuitive grasp of the issues is of equal or greater importance to the designer than a rigorous analysis; the more so since many types of corrupted channels are not amenable to a tractable mathematical analysis. To help provide this intuitive feeling, the programmable corrupted channel simulator was conceived and constructed.

It is unfortunate, but true, that the general corrupted channel does not exist. The types of degradation that the designer will have to consider are subject to nearly as many variables as the degradations themselves introduce, e.g., portion of the frequency spectrum in use, length of path, time of day, local population, current technology, and so on. Accordingly, it was necessary to select some types of degradation which could reasonably be expected to be somewhat representative of the types actually encountered in practice.

As a first step, the limitations on the channel had to be set. This in turn required that some assumptions be made about the character of the signals with which the simulator would be used. It was deemed likely that most systems to be tested would be operated in the vicinity of 100 KHz, a frequency within the reach of most current operational amplifiers which are the devices of choice for most applications at present. Two other devices presently available to the students for whose use the simulator is intended, a voltage controlled oscillator and a four quadrant multiplier, are capable of operation to 500 KHz. With this in mind, the decision was made to design the simulator to have a nominal frequency response from 500 KHz to as low a frequency as would prove practical (which turned out to be about 20 Hz). Thus the simulator would impose its first limitations on the prospective user as dc blocking low pass channel.

In order to give some definition to the upper cutoff frequency, the simulator was planned to accommodate signals occupying bandwidths of about 10 KHz. This implied that the most commonly employed definition of cutoff frequency (-3 dbV from mid-band) would not necessarily be valid. Apparently, the concept of channel bandwidth would be more than ever an arbitrary opinion on the part of the user since it would depend strongly on the user's signal bandwidth and the slope of simulator transfer characteristic in frequency range occupied by the user's signal. Thus, at least on this count, the definition of simulator bandwidth was left an open question.

Another decision of a general nature concerned simulator input and output impedance. Insofar as possible, the simulator was planned to be compatible with user output impedances as high as and input impedances as low as 10 K (ohms).

By use of large value coupling capacitors at input and output, low frequency response would be preserved while the problem of matching dc levels could be avoided. And finally, a user input signal amplitude of no more than two volts peak-to-peak and an insertion loss of no more than 20 dbV were deemed adequate for most anticipated experimental uses of the simulator.

With the major general decisions either made or tabled pending limitations that might appear during the design stage of the project, some sort of selection would have to be made among the numerous types of degradation known. One basis for this selection would be the original cost constraint, another the desire to maintain reasonable design simplicity in order to provide a measure of reliability and to insure that a working system be produced in the one term allotted to the project.

A preliminary list of some representative types of degradation which might be realized by relatively straight forward design procedures was compiled. This list, the rationale behind the functions appearing on it, the selections made from it, and the possible realizations of the selected functions considered by the author form the subject of the next section.

III. SELECTION OF FUNCTIONS

The preliminary list of possible functions contained eleven items. With a short discussion of their relevance and realizations considered, they were:

1. Wideband Noise - the constant companion of the communications system. One might like to provide a facsimile of the mathematically tractable white Gaussian noise, the common example in introductory statistical communication theory. Commercial white noise generators employing temperature limited vacuum diodes in a magnetic field are available at a cost of several hundred dollars. A more reasonable approach would be to use the intrinsic noise of semiconductor amplifiers which commonly display uniform spectral density above 10 KHz.
2. Gated Wideband Noise - or static crashes. A common affliction of AM systems and easily implemented by using a commutator on the output of a wideband noise source.
3. Frequency Shifting - translation of the received signal to a different and probably varying frequency from that at which it was transmitted or supposed to have been transmitted.

This effect can be observed as Doppler shifting in air-to-ground and satellite-to-ground systems and as drift in poorly stabilized transmitters. The most direct realization is a single sideband modulation process with a randomly varying carrier frequency. A more exotic solution would be to determine the amplitude of the incoming signal and impress it upon the output of an oscillator of randomly varying frequency which is in some sense correlated with the input signal frequency. While perhaps an interesting and potentially elegant solution, the difficulties in realization are enormous. How does one track an input which might be in one of several octaves or maintain an estimate of the carrier frequency when confronted by CO or 100% modulated AM?

4. Whistlers - or swept frequency tone bursts. Best known as a natural phenomenon from studies of the ionosphere where it was observed to occur because of the variation of propagation time versus frequency for reflections of the energy released by lightning strokes which is guided by the earth's magnetic field. More commonly, it is used as

an ECM technique effective against both AM and FM systems. It is an attractive technique since it can be implemented with only a voltage controlled oscillator and a ramp generator.

5. Amplitude Fading - nearly as common as wideband noise. Fading affects even line of sight microwave links. Realized in a straightforward manner with a multiplier and a random voltage source.
6. Selective Fading - variable amplitude response versus frequency. Encountered in scatter radio systems where the received signal is composed of a few components arriving over paths in such a manner as to interfere destructively in a narrow frequency band. The most obvious solution is to sweep the stopband of a narrow filter through the signal. No matter what the range over which the filter sweeps, whether unrestricted or keyed to the input signal, the design of an electrically tunable filter which can span several octaves is a difficult problem.
7. Impulse Noise - a concomitant of heavily populated areas in one sense. Frequently caused by poorly shielded switching circuits and often by ignition systems of internal combustion

engines. Easily produced by a randomly triggered monostable multivibrator.

8. Multipath - echoing or ghosting. A result of receiving from two or more paths of significantly different electrical length and of gains of sufficiently comparable amplitude to confuse the receiver. Familiar to most as ghosts on a television screen and also a major cause of intersymbol interference on digital data links. Requires two or more distinct signal paths with delay lines. The difficulties inherent in producing broadband delay lines seem insurmountable for delays in the range of 10 to 1000 milliseconds which would be required dramatic audible effects. The obvious solution, the all-pass network, would require such a large number of units as staggering if a constant delay to 500 KHz is to be obtained.
9. Envelope Delay - variable phase versus frequency characteristic. A cause of intersymbol interference, especially on links where data is being transmitted at near capacity. Realizable as a set of all-pass networks.

10. Continuous Wave Interference - generally a jamming technique but sometimes encountered in crowded portions of the spectrum. The CW itself can be obtained from a VCO but the tracking problem remains.
11. Amplitude Distortion - non-uniform gain versus amplitude characteristic. Often done deliberately as signal compression. Some distortion of this type will occur in any case. May be operated from averaged or instantaneous amplitude in a discrete or continuous manner by adapting any one or combination of numerous expansion and compression circuits already in use.

At this stage, it was apparent that some paring down was required, not only to avoid exceeding the cost ceiling but also to pick a set of functions which could be produced in one term.

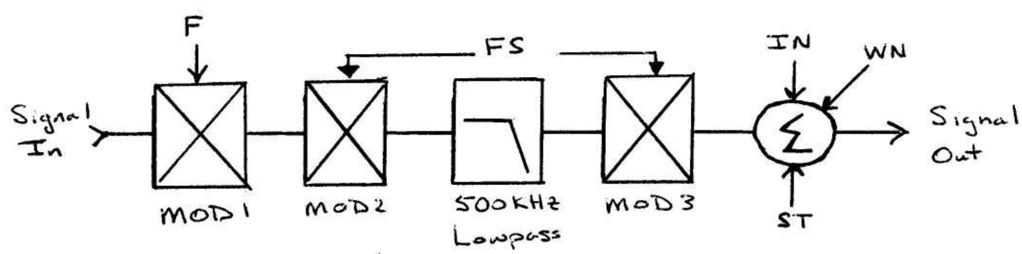
Additive wideband noise, additive impulse noise, and amplitude fading were chosen as very representative and because they share a noise source as a common component. Frequency shifting and swept tone interference were included to be sure of affecting signals which do not carry amplitude information.

IV. BLOCK DIAGRAM SYNTHESIS

Having decided on the terminal impedance characteristics, the basic transfer characteristic, and the degradation functions to be provided, a block diagram synthesis was undertaken. The general design chosen is a main signal path (MSP) with a nominal bandwidth of 500 KHz. To avoid disrupting failures, the MSP remains as unbroken as possible as the various functions are introduced and removed.

Frequency shifting (FS) and fading (F) are multiplicative functions. They can be performed with four quadrant multipliers, but these are fairly expensive and somewhat touchy devices. A cheaper, though more nonlinear device which can be used is the balanced modulator. It is also much easier to bias such a device into fixed gain when it is not in use as a multiplier.

Impulse noise (IN), wideband noise (WN), and swept tone bursts (ST) are additive functions and can be introduced rather easily. Thus, the MSP can assume the rather simple appearance of Fig. 1.



Simplified Simulator Block Diagram

Figure 1

As can be seen from Fig. 1, modulators 2 and 3 and the 500 KHz low-pass filter form an incomplete single sideband modulator. Since it is expected that frequency shifting will be one of the later functions with which the user experiments, it was deemed likely that the user's system would already be equipped with input filters to remove out of band interference. In this case, it is not really necessary to provide a second low-pass filter for the output of modulator 3.

The use of modulator 1 allows for the possibility of randomly invertible phase as well as deep amplitude fading. While phase inversion is a rather special case of the random phase channel, it would at least provide an opportunity for confusing user systems which require a coherent channel.

The nature of the summing point (or points) appeared on the surface to be the most easily soluble problem of the MSP. It was originally hoped that some sort of integrated circuit amplifier would be available for the purpose. As will be seen later, the integrated circuit realization was abandonned for a less expensive although not so immediately apparent solution.

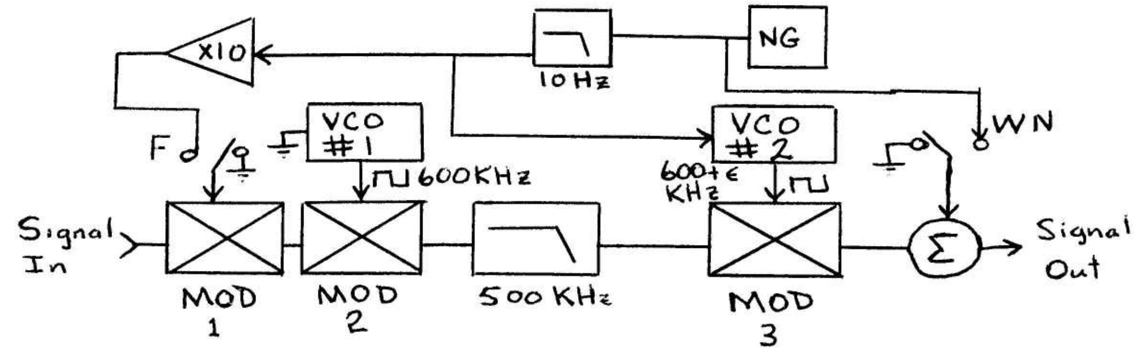
Despite the relative simplicity of the MSP, a not inconsiderable problem arose concerning the peripheral circuitry. The simulator should provide some combinations of the available functions as well as each individually, but what combinations? Clearly, the larger the number of functions to be provided simultaneously, the greater the number of peripheral circuits that would have to be included.

Taking stock of the situation, the primary limiting factor turns out to be the number of oscillators required. Frequency shifting requires two oscillators, swept tone bursts two, and impulse noise one. But, while voltage controlled oscillators are available as plug-in modules, it would be desirable to tie up as few as possible when the simulator is in use. Thus, a limit of two VCO's was set, and frequency shifting, swept tone bursts, and impulse noise were to be mutually exclusive functions. This is because the two VCO's would have to be set up differently for each.

At the same time, it is possible to offer wideband noise simultaneously with fading and/or any one of impulse noise, frequency shifting, and swept tone bursts using a single noise generator without introducing too much correlation. The swept tone function can realistically be made deterministic, hence independent of any of the random functions. Wideband noise would be the output of the noise generator and, if the other three random functions are driven by a version of that output low-passed at 10 Hz, they would be uncorrelated with noise in the user's signal band, which is nominally restricted to frequencies above 20 Hz (100 KHz for frequency shifting). In addition, the driving signals for impulse noise and frequency shifting are effectively integrated in VCO's further reducing the correlation with fading.

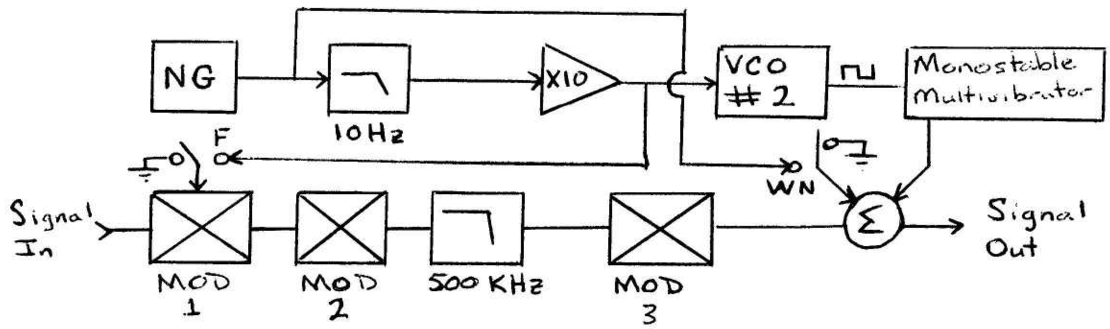
These considerations produce a minimum of peripheral circuits: two VCO's, a 10 Hz low-pass filter, two amplifiers, a monostable multivibrator, and a noise generator (NG). As shown in Fig. 2a, b, and c, the entire system can assume a rather simple configuration for each of the three mutually exclusive functions. It should be noted when examining Fig. 2 that the VCO's produce both triangle waves and square waves. These outputs are identified by conventional symbols for triangle and square waves where necessary.

One observation may be readily made from Fig. 2a: the choice of 600 KHz as the fundamental carrier frequency for frequency shifting restricts the signal input to frequencies above 100 KHz. While this represents a somewhat arbitrary restriction since all other functions may be applied to signals in the 20 Hz to 500 KHz range, it reflects the difficulty of building a filter that will separate signals 40 Hz apart at 500 KHz. Because at least 20 dbV suppression of the unwanted sideband is desired, a basic carrier of greater than 500 KHz is indicated. In addition, the choice of a 600 KHz carrier puts carrier feedthrough outside the nominal signal band. Another



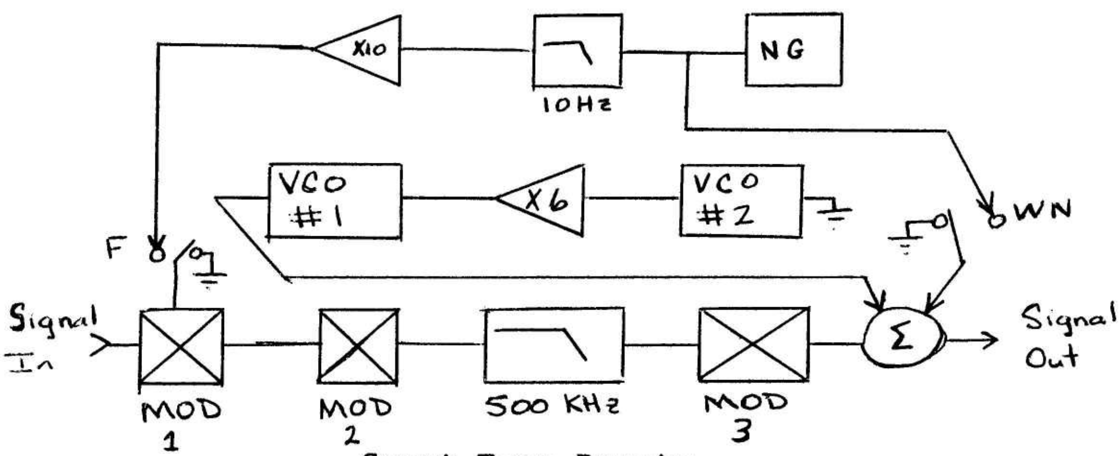
Frequency Shifting

Figure 2a



Impulse Noise

Figure 2b



Swept Tone Bursts

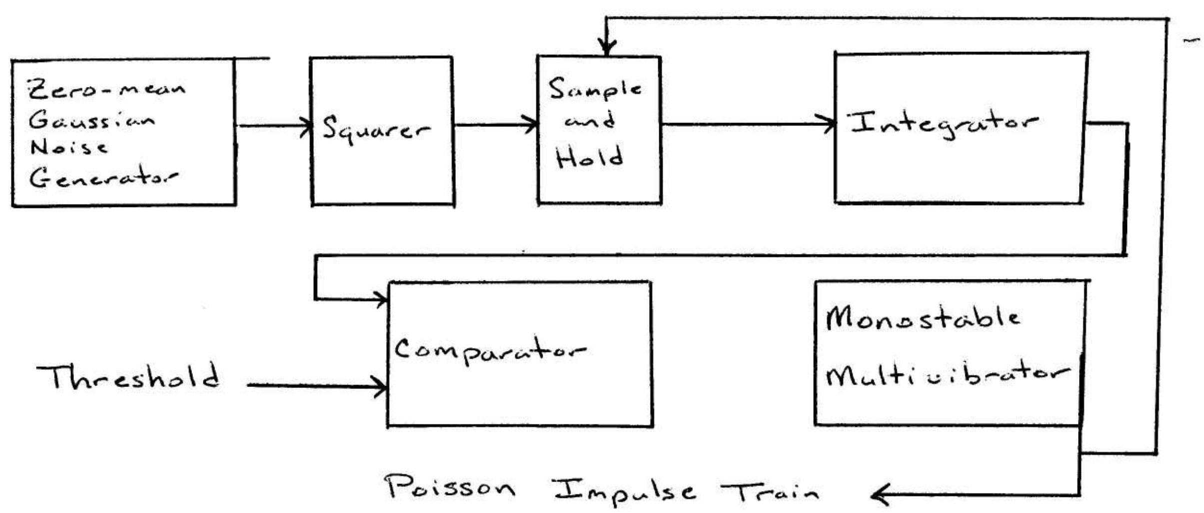
Figure 2c

SIMULATOR BLOCK DIAGRAM

observation that may be made from Fig. 2a is while 100 KHz to 600 KHz signals will be passed, it is also possible to get signals from 700 KHz to 1.1 MHz through.

In Fig. 2b, a question is likely to arise concerning the character of the impulse noise being generated. Impulse noise is commonly modeled as a Poisson process, i.e., exponentially distributed first order interarrival time and all arrivals independent. As conceived in Fig. 2b, the interarrival times would be distributed as a Gaussian process (assuming a Gaussian noise generator) with correlation between them strongly dependent upon the structure of the VCO.

Originally, a method of generating a Poisson impulse train was considered as shown in Fig. 3. Clearly, if the



Poisson Process Impulse Generator

Figure 3

noise generator is sufficiently wide band and the average interarrival time is chosen to be sufficiently long, the output can be said to be a Poisson process. The squarer produces an exponential process and, of course, any two or more samples of white Gaussian noise taken at different times are uncorrelated.

Nevertheless, this appealing solution did not fit into the overall plan. It is basically expensive to implement with sufficient accuracy to justify the complicated design. Many roughly square law devices exist, but a linear multiplier is most likely to provide desired performance. Elementary sample and hold and integrator circuits are common, but the cost factor rises rapidly if low speed performance is required. All things considered, this realization tends to be more expensive and complicated than necessary for a simple simulator.

It should also be borne in mind that a more or less lame theoretical justification may be given for correlated Gaussian interarrival times. If one considers the time to the n^{th} arrival, as n becomes large, the Erlang family of distribution functions which characterize Poisson processes become more nearly Gaussian by Central Limit Theorem arguments. At the same time, as n becomes large, the correlation begins to disappear. But, in fact, it is only necessary that the impulse train be of random period so that it can-not be removed from the user's signal by a simple bandstop filter.

The above considerations being made, and the block diagram finalized, the actual circuit design could proceed apace.

V. CIRCUIT DESIGN

The heart of the simulator project is, naturally enough, the actual circuit design. The author's intent in taking on the project was to determine his ability to produce circuits which fit the requirements of a block diagram systhesis and to gain some experience with current technology devices. As will be seen, a number of compromises were made to accommodate the limitations of present technology and the cost constraint placed on the project.

As previously stated, the circuits to be designed were modulators 1 (fading) and 2 and 3 (frequency shifting), a 500 KHz low-pass filter, a summing point, a monostable multivibrator, a wideband noise generator, a 10 Hz low-pass filter, a dc amplifier for the low-pass noise, and a dc amplifier to drive a VCO for swept tone interference. In addition, some provision would have to be made to program the VCO's because they would be operating in different frequency ranges for different functions.

The subject of this section is then the author's design experience in realizing the simulator. The circuits are present in roughly chronological order as they were produced by the author with no attempt to correlate them specifically to the overall system. The reader may refer to Fig. 2a, b, c as each circuit is discussed if he is uncertain of the circuit's position or function in the system.

Noise Generator

In most commercial applications, the semi-legendary white Gaussian noise is approximated in our range of interest (to 500 KHz) by a low-passed Poisson process. For many years, the Poisson process was the arrival of electrons at the plate of a temperature limited vacuum diode. In order to obtain a degree of uniform spectral density, the electron stream in the diode is subject to a field from an external magnet.

Needless to say, the price associated with the vacuum diode noise generator precluded its incorporation into the simulator project. This problem also excluded other types of commercial white noise sources.

Another possible solution of more recent vintage is the repeatable pseudorandom noise generator using a shift register. An attractive solution because it is more closely controlled than the truly random noise generator, the shift register approach presents other problems. Clearly the clock rate must be higher than the highest frequency at which the noise must present a uniform spectral density. This implies a large capacity shift register so that the pattern does not repeat in a time comparable to the period of any frequency in the band of interest. Further, the longer the time over which it is desired to avoid pattern repetition, the larger the shift register once again.

A compromise on the character of the noise to be provided was the message of the white noise investigation. A semiconductor noise generator seemed the best solution.

Unfortunately, semiconductors tend to exhibit 1/f noise in lower frequencies. This is not so bad in itself because the intent is to encourage the student to transmit his audio rate information outside the audio band. Thus, it was hoped that semiconductor 1/f noise would tend to force the student out of the audio band.

A still more annoying characteristic of semiconductors is the tendency for the noise level to rise at higher frequencies after being relatively constant in some middle range. This is definitely contrary to the intention of the simulator if this rise in noise level occurs below 500 KHz.

Since the noise produced by any likely source would be of such low level as to require amplification and a uniform or at worst decreading noise level was desired outside the audio band, it was decided to attempt to find a wideband integrated amplifier to fit the situation.

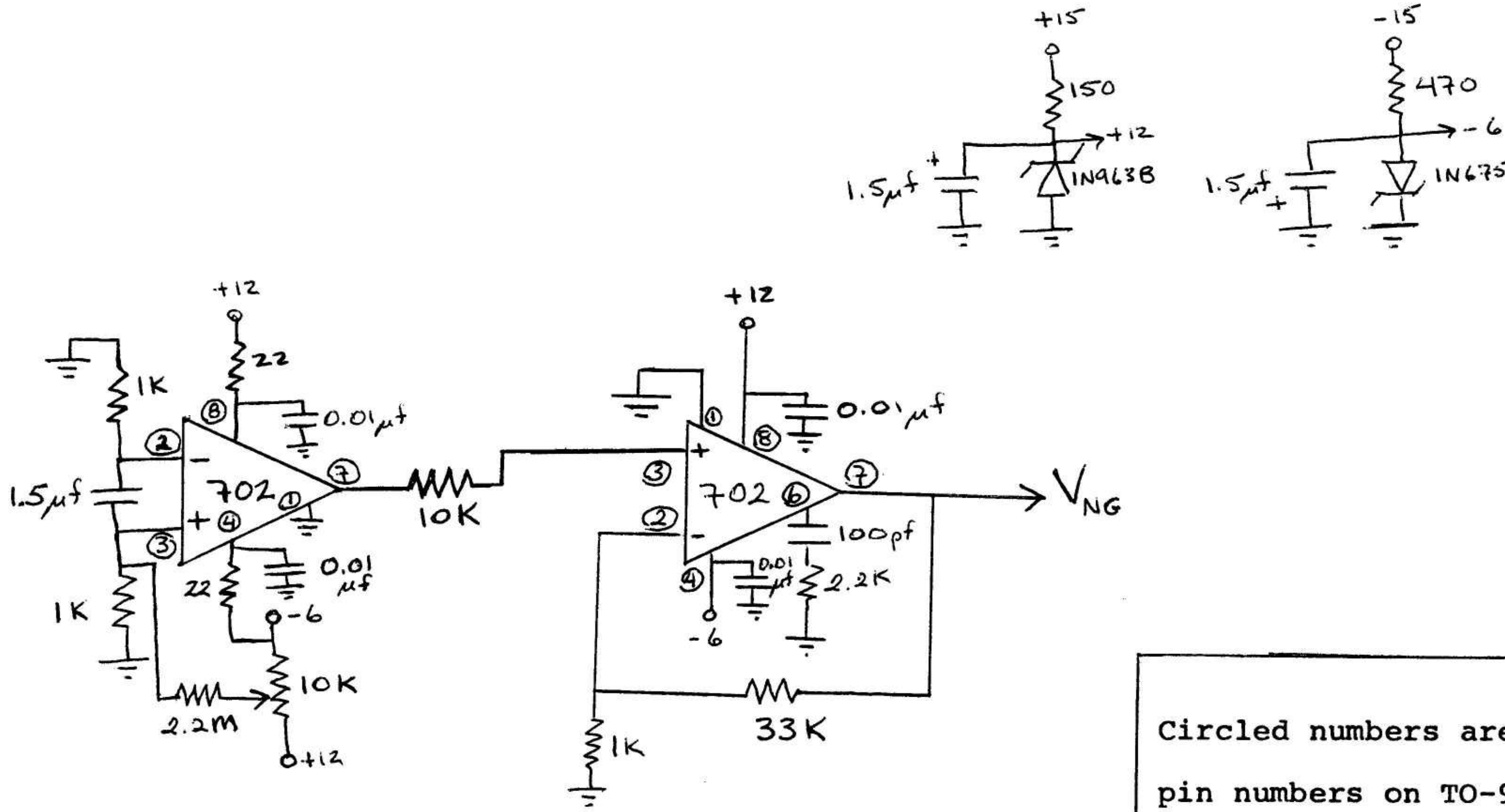
For this purpose, the Fairchild μA 702 amplifier was selected. This amplifier, with an open loop gain of about 3600, exhibits a 1/f characteristic for equivalent input noise only below 10 KHz. Further, the noise level does not rise at higher frequencies.

One drawback of the 702 is that as an older generation device, it requires the odd supply voltages of +12 V and -6 V. This necessitated generation of these two voltages in the simulator because it was planned that the user only be responsible for providing +15 V and -15 V. As the design of the simulator progressed, it became apparent that only the noise generator would require +12 V, -6 V. Thus, a pair of Zener diodes used as regulators was sufficient to produce these two voltages.

Having selected the 702 and disposed of the supply voltage problem, attention turned to the noise level to be produced. In order to allow the user to attain large signal to noise ratios, it was decided to produce an output noise spectrum of the order of a millivolt/$\sqrt{Hz}$. The mid-band equivalent input noise voltage of the 702 (10 KHz to 1 MHz) is around 10^{-8} V/$\sqrt{Hz}$. Assuming an open loop gain of 3600 for a 702, a second amplifier with gain of 30 or so would be required to achieve the desired output. For this second amplifier, another 702 was used.

The final noise generator circuit is shown in Fig. 4 along with the regulator networks used to produce +12 V and -6 V.

In Fig. 4, a few elements bear explanation. The 10K potentiometer and the attendant voltage divider provide control over the offset of the first 702. The 10K resistor between the output of the first 702 and the input of the second was added to provide some additional isolation



Circled numbers are pin numbers on TO-99 package.

Noise Generator

Figure 4

to suppress a tendency to oscillate. (Notice that this could have been done more simply by structuring the second 702 as an inverting amplifier but it had already been built non-inverting before the 10K was added.) The 100 pf - 10K combination on pin 6 of the second 702 composes a compensation network to help minimize the amplification of high frequency pickup. Because the 702 has gain to 100 MHz and spurious signals of 10 MHz and 100 MHz were present in the room where the circuit was built (not, by the way, as a result of oscillation of the noise generator), this was a necessary precaution.

The final noise generator has a noise output of approximately 2 volts peak-to-peak, plus about 2 V peak-to-peak of amplified 10 MHz pickup (under worst conditions). Checks with a General Radio 60 KHz spectrum analyser and a Tektronix 1 MHz spectrum analyser confirm that the noise voltage above 5 KHz is indeed approximately 1 mV rms/ Hz. Thus, around 100 mV rms of noise is available in a 10 KHz band above 500 KHz.

500 KHz Filter

Filters of one kind or another may be found in almost every circuit ever built for any purpose. It was the author's experience that the 500 KHz filter presented the most troublesome problems in designing the simulator.

The intention of the 500 KHz filter is to provide at least 20 dbV suppression of an unwanted sideband in a single sideband modulator. The author's experience with passive filters indicated that filters of this type were plagued by one or more of five problems: gentle slope of the transfer characteristic, high insertion loss, unwarranted mathematical complexity, the presence of inductors, and uncommon and very critical component values.

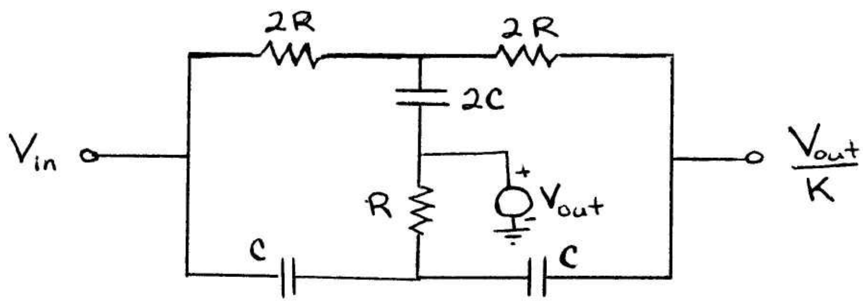
While vainly searching literature on passive filters for some design inspiration, the author stumbled across the world of active filters and was pleased to discover that filter characteristics can be made dependent upon the gain of a voltage controlled voltage source (VCVS). Since an operational amplifier nicely fits the description of a VCVS, this seemed to be a fine solution.

In order to obtain a sharp rolloff in the vicinity of 500 KHz (it will be remembered from section II that the definition of cutoff would be as dependent upon the filter slope as actual attenuation), the best approach appeared to be the use of a notch filter. A particularly simple VCVS realization of this type of filter is shown in Fig. 5.

In the frequency domain, the transfer function of the filter, $T(s) = V_{out}/V_{in}$, is given by

$$T(s) = \frac{(4R^2C^2s^2 + 1)K}{4R^2C^2s^2 + 8(1-K)RCs + 1} .$$

Thus, the width of the notch may be made arbitrarily small by allowing K to become arbitrarily close to 1. (It might also be noticed that if K is greater than 1, the filter is unstable.) This will allow extremely steep skirts in the vicinity of the notch frequency insofar as the passive components are matched.



Active Notch Filter

Figure 5

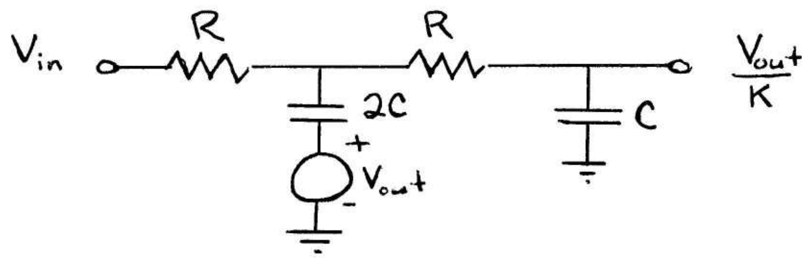
The unwanted effect of the notch filter is the return to non-zero gain above the notch frequency. To suppress this effect, low-pass filter may be used to complete the low-pass function.

Once again, VCVS realizations are available in which filter characteristics are a function of the VCVS gain. The configuration adopted is shown in Fig. 6.

The circuit of Fig. 6 has a transfer function

$$T(s) = \frac{K}{2R^2C^2s^2 + 2(2-K)RCs + 1} .$$

As can be seen, both the cutoff frequency and rolloff slope are functions of K. In fact, as K approaches 2, there is increased peaking in the vicinity of the 3 dbV cutoff frequency. While this is undesirable in itself, it can be removed by allowing the notch filter to begin to roll off at the same point as the low-pass begins to peak. This technique accomplishes the double purpose of sharp skirts at cutoff and good suppression in the stop band of the low-pass filter.



Active Low-Pass Filter

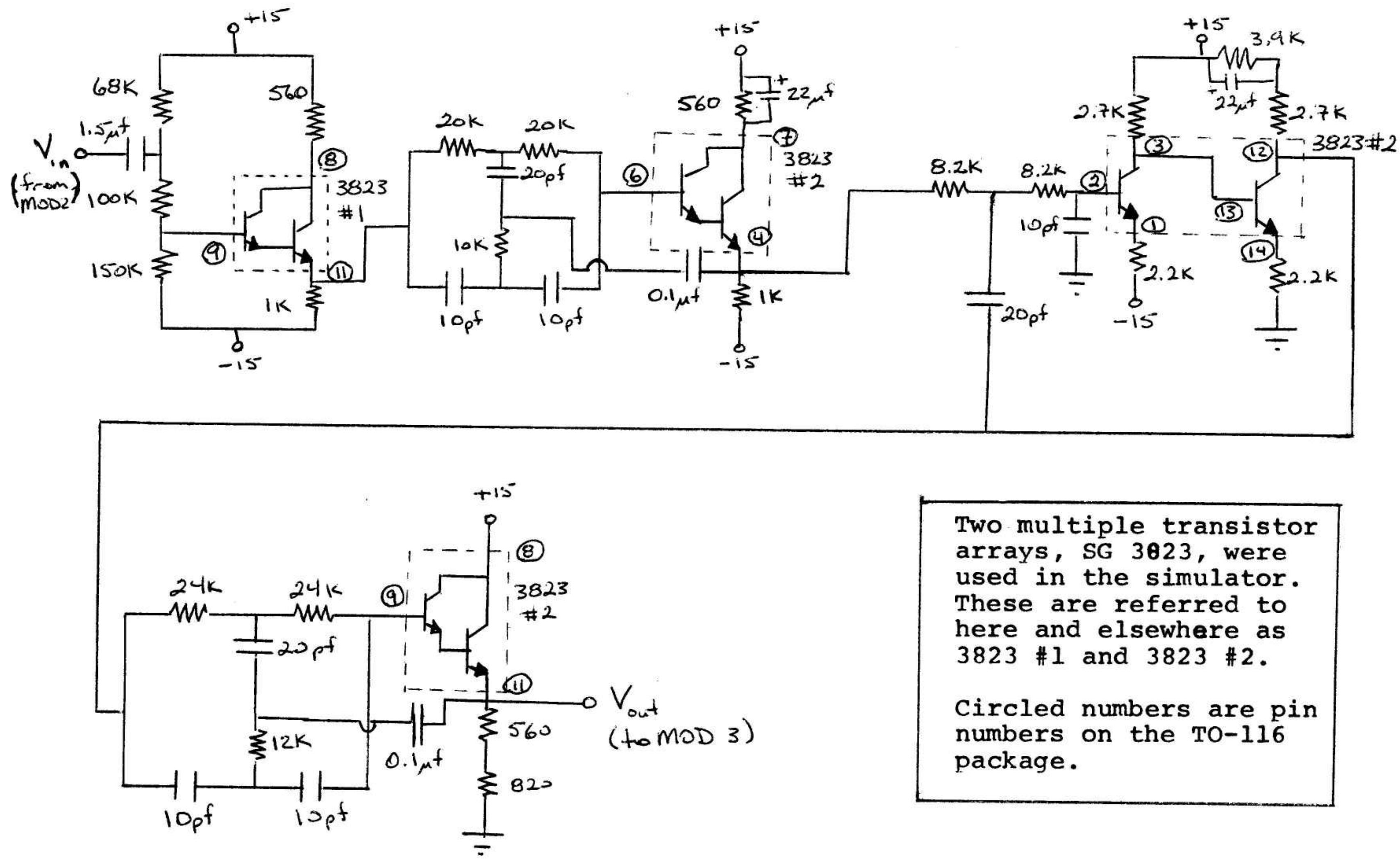
Figure 6

But the biggest setback was yet to come. The device of choice for the 500 KHz filter was again the μA 702. An experimental notch filter was built and tested and found to fail to function as designed. After testing a similar filter at lower frequency with the μA 741, the problem was discovered to be in the input and output impedances of the 702. These were respectively too low and too high to fit the description of a VCVS for the passive component values contemplated.

Fortunately, only low gain was required for the filter realizations chosen: less than but near unity gain for the notch filter; less than two for the low-pass. Less than unity gain with high input impedance and low output impedance describes an emitter follower or more specifically the Darlington follower which was used. Positive gain greater than one but less than two (1.5 was actually employed)with similar input and output impedances can be achieved with a pair of common emitter amplifiers.

An added advantage of transistor amplifiers which was not immediately apparent concerns the impedance as seen by the VCVS output in Fig. 6 and Fig. 5. The input impedance at this feedback point decreases with frequency. This, in turn, decreases the effective VCVS gain, K in the transfer functions if the VCVS output impedance is non-zero. This decrease in impedance is, however, not of great importance anywhere except above the filter break frequency. Thus, the non-zero output impedance of the transistor amplifiers adds to the efficiency of the complete filter as a low-pass.

The final filter is shown in Fig. 7. The first Darlington pair is simply a follower which was added to avoid loading the preceding modulator. The second Darlington pair is a near-unity gain VCVS for an active notch filter at 800 KHz. The direct coupled two stage common emitter amplifier is a VCVS of gain 1.5 for a low-pass filter with 3 dV



Two multiple transistor arrays, SG 3823, were used in the simulator. These are referred to here and elsewhere as 3823 #1 and 3823 #2.

Circled numbers are pin numbers on the TO-116 package.

500 KHz Low-Pass Filter

Figure 7

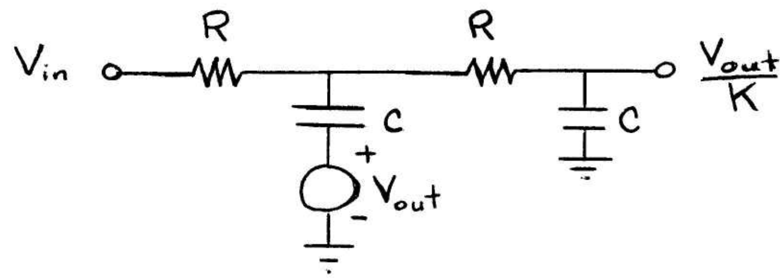
point about 500 KHz. (The bypassed 3.9K in the collector of the second stage is simply to decrease the dc dissipation.) The final Darlington pair is the VCVS for a notch filter at 650 KHz.

The inclusion of a second notch filter at 800 KHz in the cascade was dictated by a desire for a flatter stop band in the 500 KHz low-pass filter. The result was a suppression of better than 20 dbV of the upper sidebands obtained by modulating input signals in the range 100 KHz to 500 KHz with the 600 KHz carrier as shown in Fig. 2a.

10 Hz Low-Pass Filter

Slowly varying random signals are required to drive a VCO for frequency shifting and impulse noise and to drive a multiplier for amplitude fading. Realistic rates of change are not necessarily in order for these functions, because the underlying purpose of the simulator is to provide disturbances which are readily visible on an oscilloscope trace or audible to the listener. The choice of 10 Hz as the maximum rate of change for this slowly varying signal was then arbitrary but reasonable in context. The period, 100 milliseconds, represents a fairly rapid but perceptible event in the sensory context. Because a two-pole active low-pass filter was employed, roughly 30 dbV extra suppression of stray 60 Hz pickup (an almost always undesirable effect) is provided.

The general form of filter chosen for the 10 Hz low-pass is given in Fig. 8. For this configuration, the



Alternate Active Low-Pass Filter

Figure 8

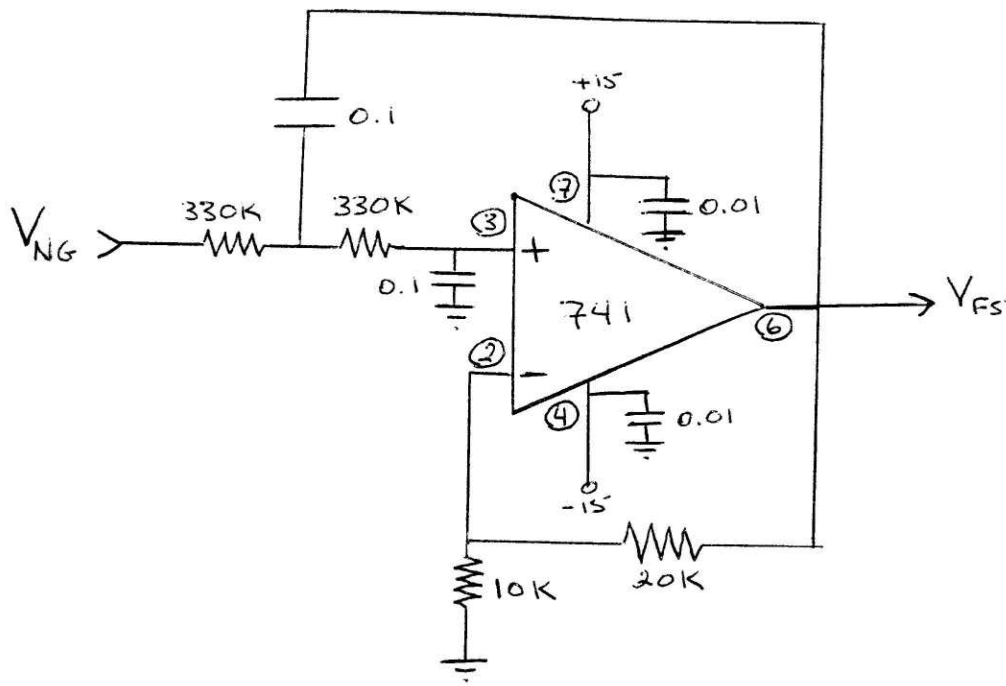
transfer function is

$$T(s) \quad \frac{K}{R^2C^2s^2 + (3-K)RCs + 1} .$$

By letting K approach 3, one obtains a more rapid appearance of 12 dbV per octave skirt slope at the expense of peaking in the neighborhood of $S = j/RC$.

In the final 10 Hz low-pass filter design, it was decided to give some emphasis to 10 Hz components. This decision was reached after subjective experimentation revealed that choosing K = 3 in the configuration of Fig. 8 produced an output with a pleasingly detectable short term variation and larger amplitude long term variation when viewed as an oscilloscope trace at 100 milliseconds per centimeter.

This final design is shown in Fig. 9. The output of this filter called V_{FS} is used to drive VCO #2 (Fig. 2a) when



All capacitors in μf.

Circled numbers refer to pin numbers on TO-99 package.

10 Hz Low-Pass Filter

Figure 9

the simulator is programmed for frequency shifting. When the input is derived from the noise generator of Fig. 4, the output shows short term variations (of the order of 100 milliseconds) of approximately 300 mV amplitude while the long term variations (of the order of seconds and minutes) have an amplitude of about one volt.

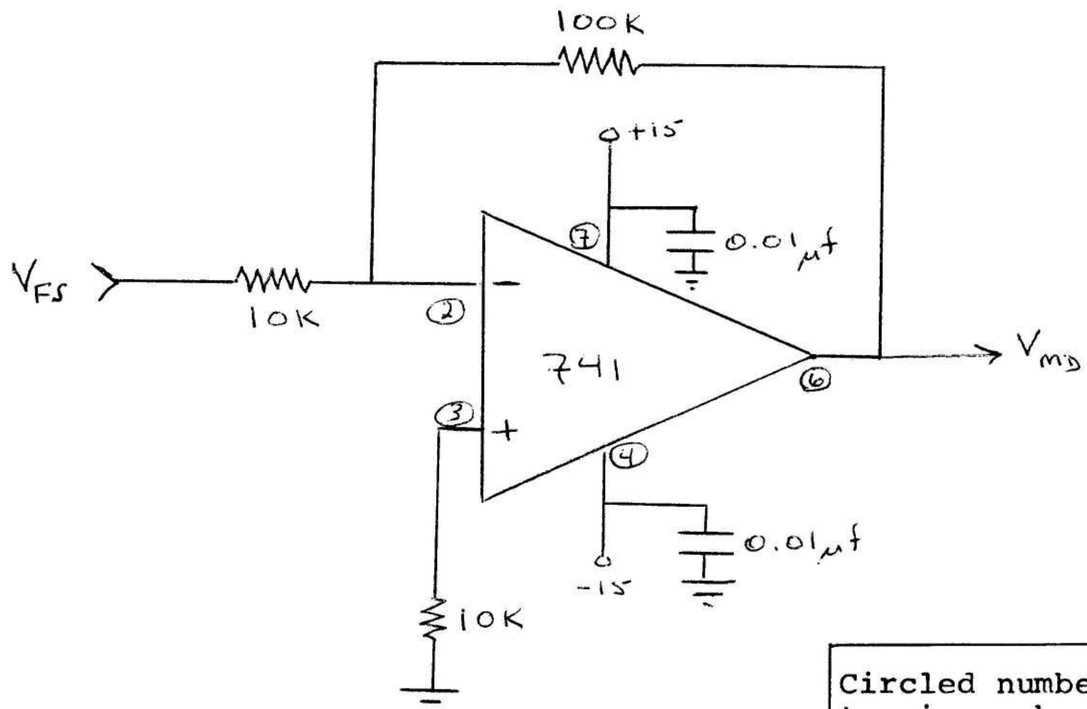
X10 Amplifier

The output of the 10 Hz low-pass filter was not sufficient to drive modulator 1 for fading nor VCO #2 for impulse noise. Principally because a short term (10 Hz) random signal of about 3 V amplitude was required for modulator 1, the gain of the dc amplifier for low-pass noise was set at 10 (hence, the name X10 amplifier).

The X10 amplifier, as shown in Fig. 10, is a common design. Its output, V_{MD}, is used to drive modulator 1 for the fading function and VCO #2 for impulse noise.

Modulator 1

Fading is, like any other real-world random process, not amenable to a simple generalization for all situations. The most common example of a fading channel in introductory communication theory is the Rayleigh channel. The original intent of the author was to attempt to produce a simulated Rayleigh channel.



Circled numbers refer to pin numbers on TO-99 package.

X10 Amplifier

Figure 10

A Rayleigh random process may be obtained by band-passing and envelope detection of white Gaussian noise. Because a noise generator was built which approximates white Gaussian noise above 10 KHz, this seemed a realistic goal. However, at this point, the cost constraint again came into play.

In order to effectively utilize the Rayleigh process which could be generated, it is necessary to use a linear multiplier. Already blacklisted because of its cost and tendency to drift, the multiplier effectively eliminated Rayleigh fading from consideration. As a result, another type of channel model had to be considered.

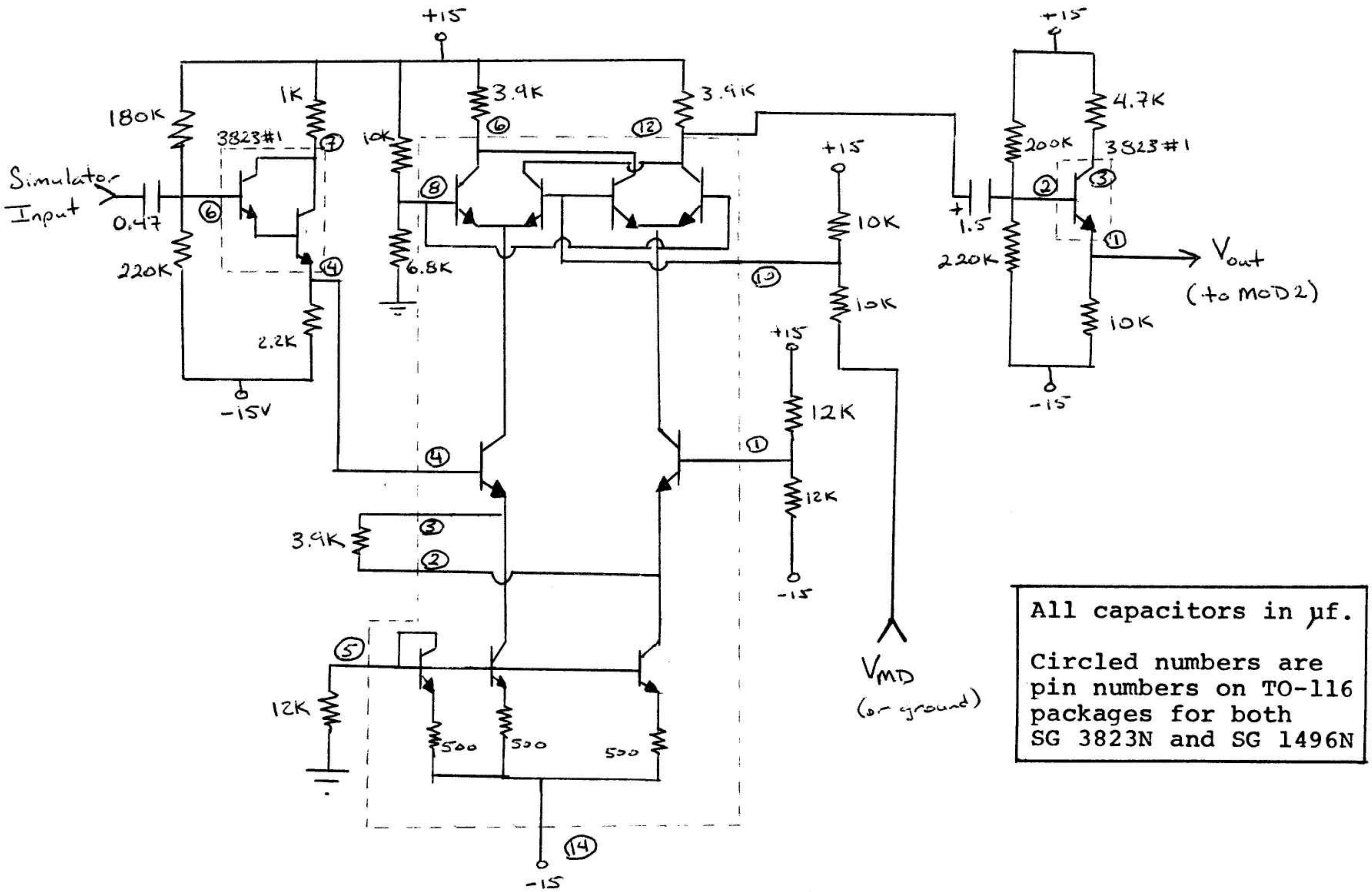
In most systems where fading is important, the limiting factor is usually the receiver AGC threshold. In such a system, the decoder tends to see a constant average signal amplitude for all receiver signals above the lower AGC threshold. As the received amplitude falls below the AGC threshold, the decoder begins to see an amplitude more or less linearly related to the received signal. Thus, the decoder in this type of system sees only deep fades.

Using a Silicon General SG 1496 balanced modulator, this type of random deep fading channel can be modeled rather easily. In addition, if the random drive is permitted to vary through the switching point of the modulator, a Bernoulli random phase channel with shifts of 0° and 180° (nominally) is obtained. While this is not the most realistical random phase channel, it does at least allow some testing of coherent systems under adverse conditions.

The circuit of modulator 1 is shown in Fig. 11. The Darlington pair at the simulator input is used as an input buffer and a dc bias supply for the SG 1496 signal input. Similarly, the emitter follower buffers modulators 1 and 2 and provides the bias point for modulator 2 signal input.

The fading drive, V_{MD}, is input to the switch voltage divider controlling pin 10. The side of the switch is biased at pin 8 to +6 V so that, when V_{MD} is replaced by ground to remove the fading function, fixed gain through the modulator is assured (pin 10 is at +7.5 V in this case). This is the situation that required that V_{MD} have a fast variation of 3 V in order to appear as 1.5 V at pin 10. By use of the trim potentiometer in the noise generator (Fig. 4), the average level of V_{MD} may be chosen for the desired proportion of deep fades. The internal construction of the SG 1496 modulator is such that gain changes in the output signal are effected only over a range of differential inputs to pins 8 and 10 of one volt. Thus, because of the large 10 Hz component in V_{MD}, the fades consist of changes in the output signal from 0 dV to a loss of several tens of dbV back to 0 dbV in a characteristic time of the order of 100 milliseconds.

In order to prevent large swings in the bias levels as it switches, the SG 1496 must be dc balanced at pins 1 and 4.



All capacitors in µf.

Circled numbers are pin numbers on TO-116 packages for both SG 3823N and SG 1496N

Modulator 1

Figure 11

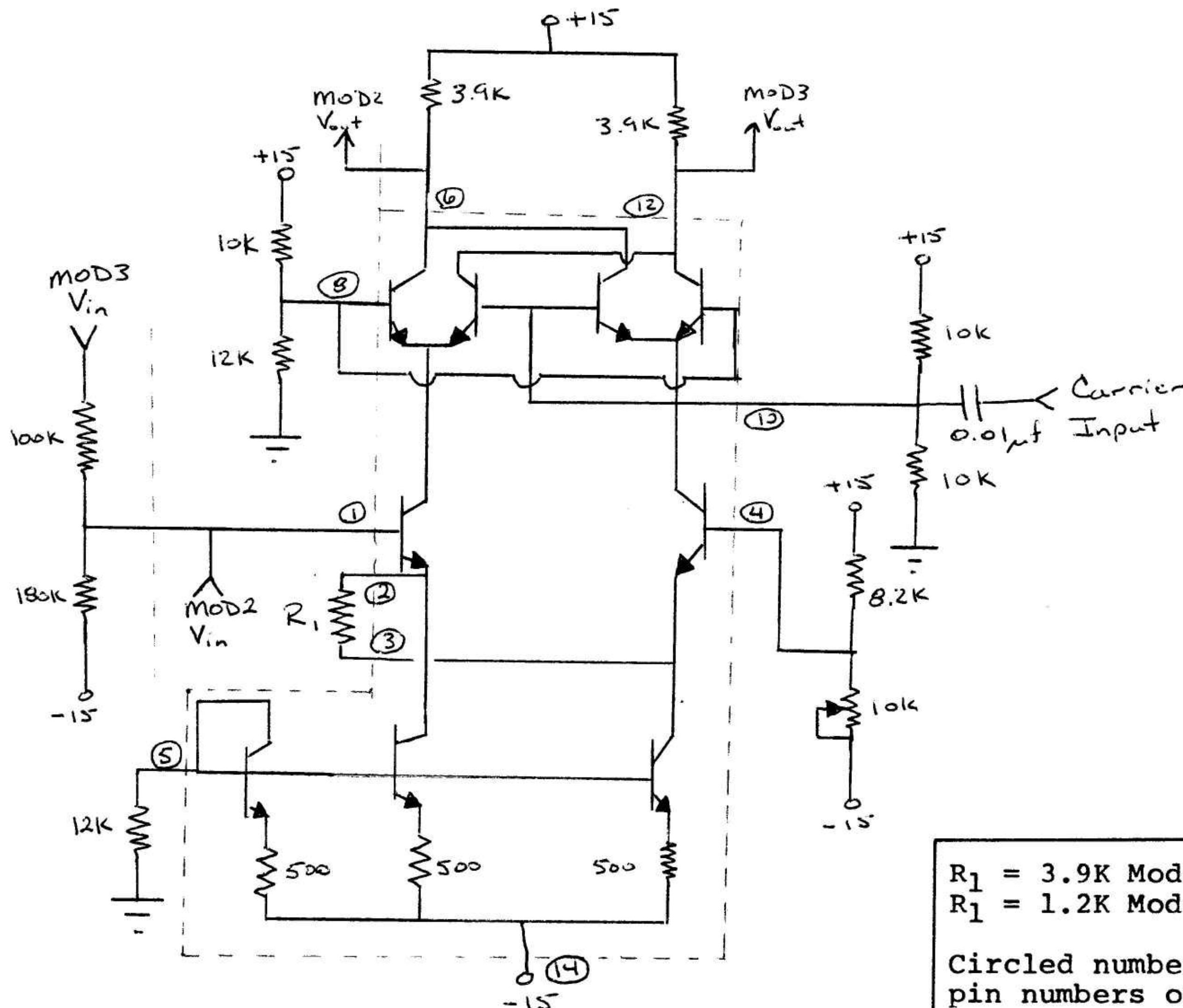
The voltage divider on pin 1 provides a rough balance sufficient to prevent swings great enough to distort the user's signal. The additive feedthrough of V_{MD} is below the nominal lower cutoff of the simulator (20 Hz) and is expected to be filtered out by the user at his receiver.

Because large amplitude user signals (2 V peak-to-peak) are anticipated and the bias current is set at about 1 milliamp by the 12K resistor on pin 12, a maximum of unity gain is provided for the user signal. This is accomplished by using equal value resistors (3.9K) as collector loads for the switch (pins 6 and 12) and as the coupling resistor (pins 2 and 3).

Modulators 2 and 3

These modulators are nearly identical circuits. The few, minor, exceptions are shown in Fig. 12. The carrier inputs are 5V peak-to-peak square waves at 600 KHz and $600+\epsilon$ KHz from the VCO's when frequency shifting is selected. Otherwise, these inputs are open and the modulators are biased into fixed gain by the divider networks on pins 8 and 10.

The fixed gain of modulator 2 is unity using a 3.9K coupling resistor and that of modulator 3 is about three using a 1.2K. Modulator 3 was given greater than unity gain to compensate for insertion losses elsewhere in the



R_1 = 3.9K Modulator 2
R_1 = 1.2K Modulator 3

Circled numbers are pin numbers on TO-116 package.

Modulator 2/3

Figure 12

simulator which amounted to about 15 dbV. While this was not particularly important for most functions, an additional 12 dbV of loss is inherent in the modulation for frequency shifting. It was simply felt that 10 dbV of gain would help negate the effects of carrier feedthrough.

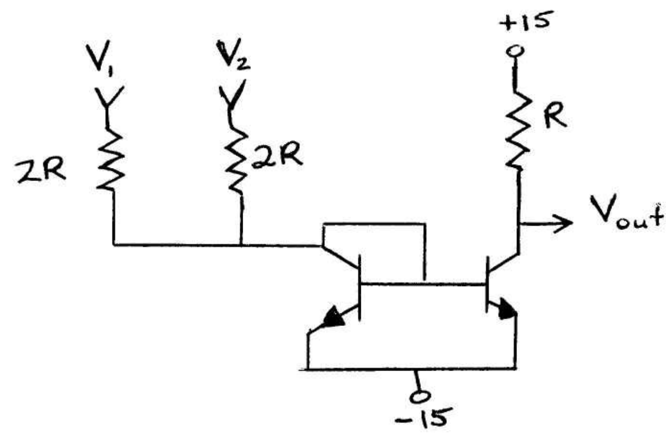
Summing Point

What should have been the easiest of the design problems, a summing point for the additive functions, turned into the most nagging. When the project began, it seemed that a conventional operational amplifier adder would solve the problem quickly and finally. When the difficulty in finding an operational amplifier which was functional at 500 KHz and had high input impedance was discovered, a very nearly frenzied search began.

The greatest hurdle was that the summing point would have to accept inputs down to nearly dc. In fact, capacitive coupling of the noise generator had to be avoided because it frequently caused oscillation.

An apparent savior was the current reflector shown in Fig. 13a. In this configuration, one obtains the relation $V_{out} = 15 - \frac{1}{2}(V_1 + V_2 + 30)$ if the base current of the rightmost transistor is small compared to the input current from V_1 and V_2 and the transistors have the same diode equation.

Several trials were made of the current reflector realization with varying degrees of success. In each case, linearity appeared to suffer more than necessary. Even matched pairs tended to distort signals of the desired amplitude (about one volt peak-to-peak each) inserted into the reflector.

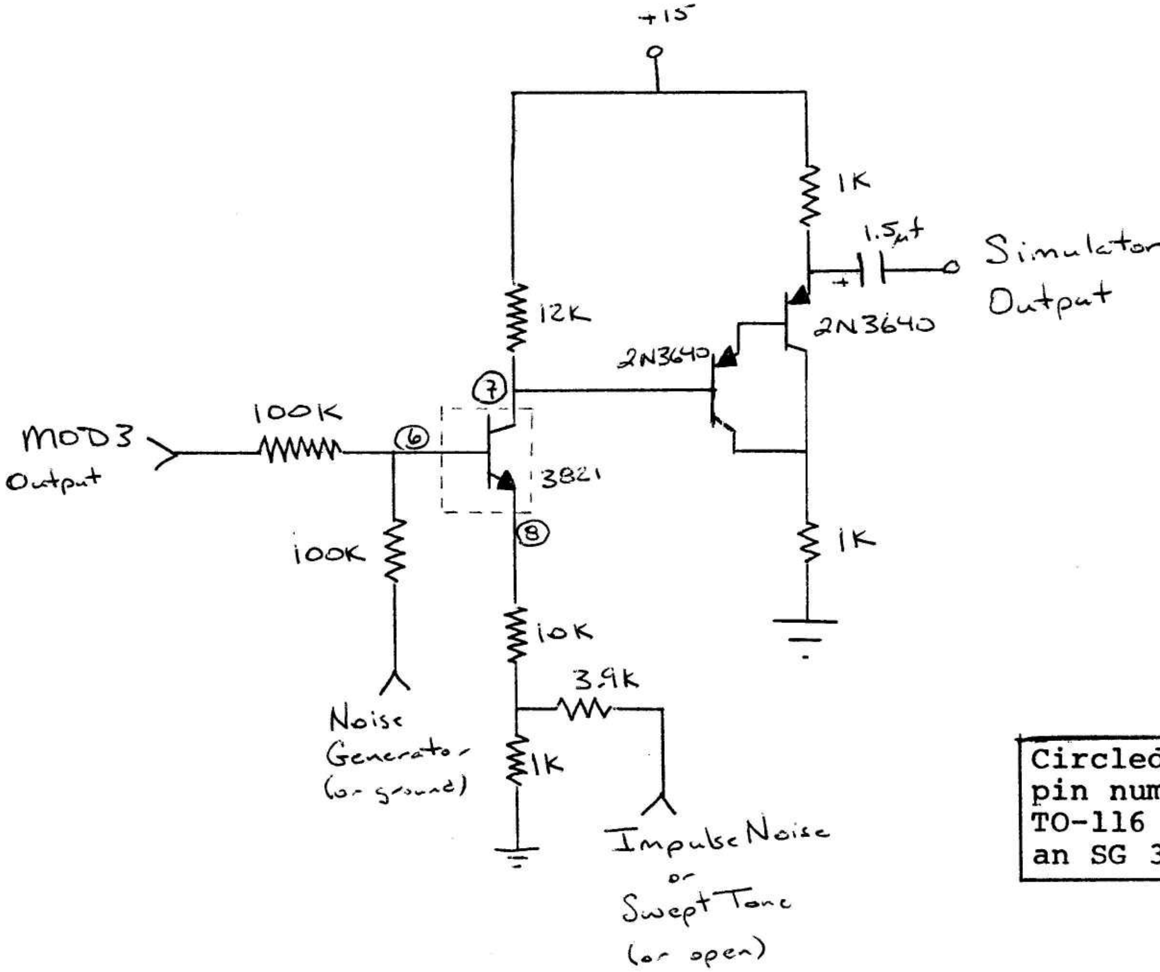


Current Reflector Summer

Figure 13a

The author finally hit upon the technique illustrated in Fig. 13b. While not perhaps historically novel, the noise generator summing point seems to the author to be an interesting way of using the high current capability of integrated circuit amplifiers (in this case the µA 702). This will also be recognized as the scheme used to drive modulator 1 (Fig. 11) for fading.

The output of modulator 3 has a dc bias of 11 V and the noise generator is operated at approximately 0 VDC. The ac



Circled numbers are pin numbers on the TO-116 package of an SG 3821.

Summing Point

Figure 13b

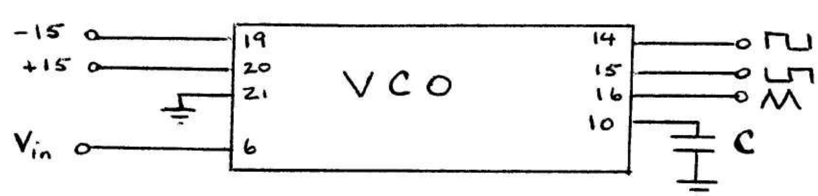
components from the modulator and noise generator are each attenuated by a factor 2 and applied to the base of a common emitter amplifier with gain near one.

In the emitter circuit of the amplifier, a similar summing point is provided for impulses or a swept tone through a 5 times attenuator.

The PNP Darlington follower is the output buffer for the simulator. The 1K collector resistor limits dc dissipation and the 1.5 μf capacitor is made from tantalum and can tolerate dc biasing at the output from about +15 V to -40 V.

VCO Modification

The VCO plays a major role in the simulator, having a hand in frequency shifting, impulse noise, and swept tone interference. Because VCO's were available as plug-in modules under the auspices of the Audio Frequency Communications Laboratory course, these circuits were merely adapted by the author to suit his needs. For this reason, the VCO is shown only as a black box in Fig. 14 where the numbers in the box refer to pin numbers on the printed circuit board on which it is mounted.



VCO

Figure 14

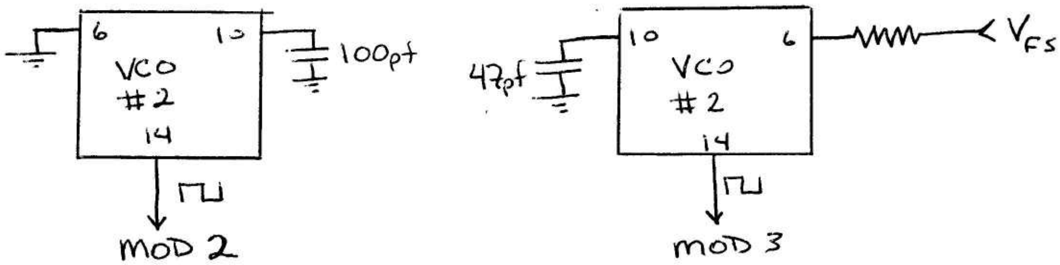
Internally, this particular VCO consists basically of a bilateral current driving the timing capacitor, C in Fig. 14. The driving current is derived from the voltage at pin 6 dropped across an internal 15K resistor which is connected to a current reflector at -15 V.

Two VCO's were obtained for the simulator. Referred to Fig. 2, VCO #1 has serial number 1 and VCO #2 has serial number 8. Both of these devices provide triangle waves (pin 16) of 5 V peak-to-peak (-2.5 V to +2.5 V) and complementary square waves (pins 14 and 15) of 5 V peak-to-peak (0 V to +5 V) from low output impedance voltage sources.

External element values were determined experimentally for the VCO configurations required by the three functions they effect. The results of these experiments are summarized in Fig. 15a, b, c. The 47 pf and 100 pf capacitors of Fig. 15a are hard-wired onto the boards and do not affect the normal VCO configuration in which the nominal capacitors are 300 pf. All other capacitors in Fig. 15b and c are added by a switch external to the VCO's as the various functions are selected.

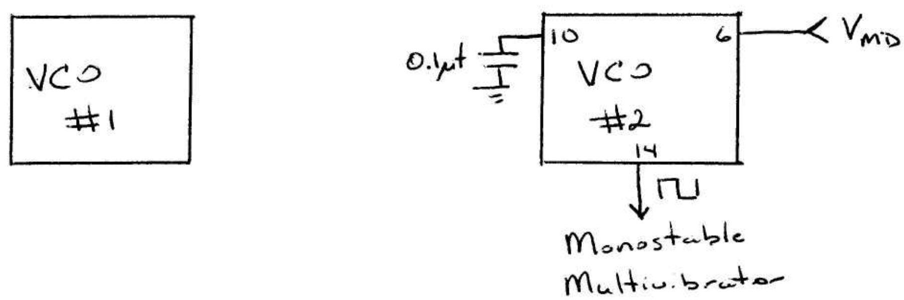
X6 Amplifier

It is necessary to return the VCO input either to -15 V or an open circuit to stop oscillation because of series resistor-current reflector construction. The open circuit



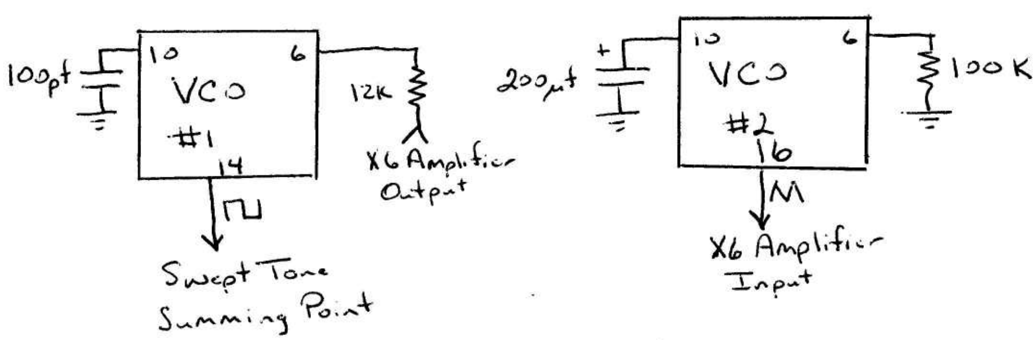
Frequency Shifting

Figure 15a



Impulse Noise

Figure 15b



Swept Tone Bursts

Figure 15c

VCO CONFIGURATIONS

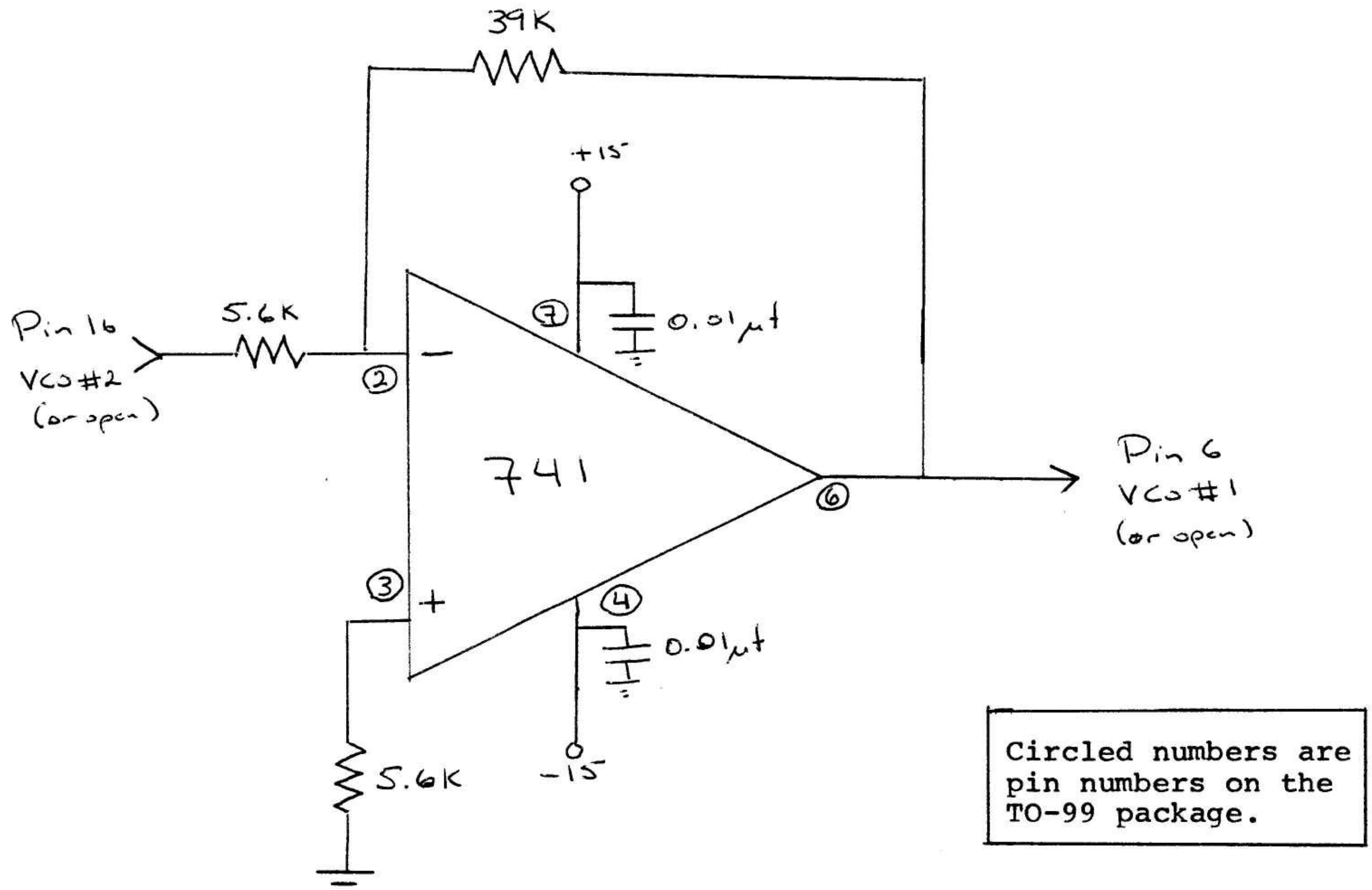
approach is used when the VCO's are not required for the function selected but is not practical for the input of VCO #1 for swept tone interference. For this reason, the triangle wave output of VCO #2 (cf. Fig. 15c) must be amplified from 5 V peak-to-peak to 30 V peak-to-peak (-15 V to +15 V) by the X6 amplifier.

As shown in Fig. 16, the X6 amplifier is a simple fixed gain inverting operational amplifier configuration using a µA 741. The gain of the amplifier is more nearly 7 than the nominal 6. This provides a waiting time at the top and bottom of the VCO #2 triangle when the 741 goes into saturation. This waiting time will provide a brief respite for broadband FM systems under test because the output of VCO #1 will be at constant (except for drift effects) frequency during this time.

Monostable Multivibrator

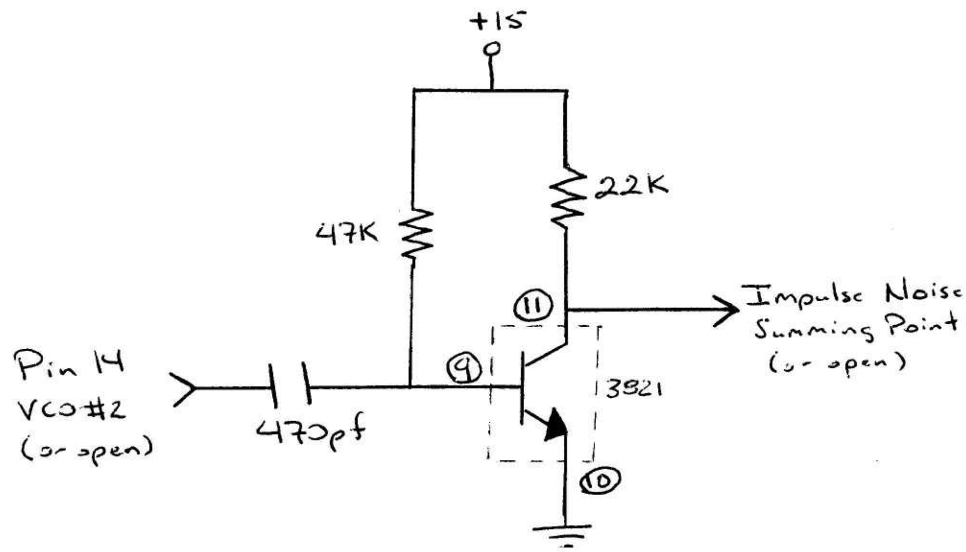
One of the classic circuits of electrical engineering, the single shot is easily realized with one transistor as in Fig. 17. This monostable multivibrator is triggered into cutoff by the negative edge of square waves from VCO #2.

The timing elements of 47K and 470 pf were chosen to provide a pulse width of about 5 microseconds (since the capacitor must charge for about 0.3 time constant to resaturate the transistor) which was deemed long enough to be



X6 Amplifier

Figure 16



Circled numbers are pin numbers on SG 3821 TO-116N package

Monostable Multivibrator

Figure 17

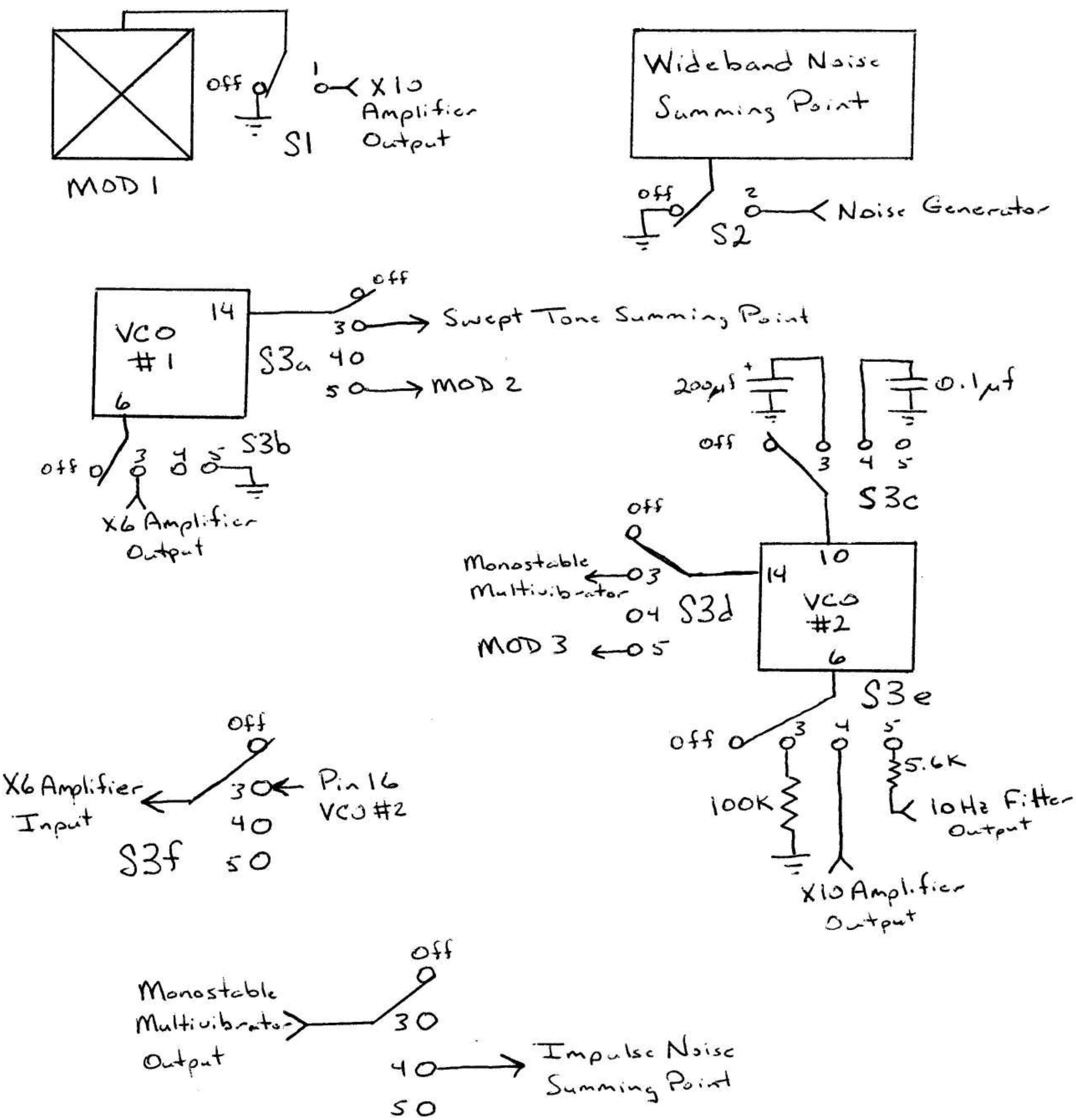
perceptible to most prospective users. A shorter pulse width is not required because the repetition rate is not normally more than 1 KHz.

Switching

A suprisingly major problem arose in connection with the switching of functions in and out of the simulator, i.e., programming. Wideband noise and fading were easily done. A pair of single pole double throw slide switches are used to connect the wideband noise summing point to the noise generator output (or ground) and the modulator 1 drive point to the X10 amplifier output (or ground) respectively. These two functions are independent of one another and the other three functions and thus may have separate controls.

Programming swept tone interference, impulse noise, and frequency shifting was a more challenging task. There are multiple connections to be made for each of these three functions and it is undesirable to have the user make more than one setting per function.

A grueling investigation of the operation of various switch types (relays, rotary switches, thumbwheel switches, etc.) led to the adoption of the non-shorting rotary layer switch. A ten pole six position model (of which seven poles and four positions are actually used) was chosen.



1 = Fading
2 = Wideband Noise
3 = Swept Tone Bursts
4 = Impulse Noise
5 = Frequency Shifting

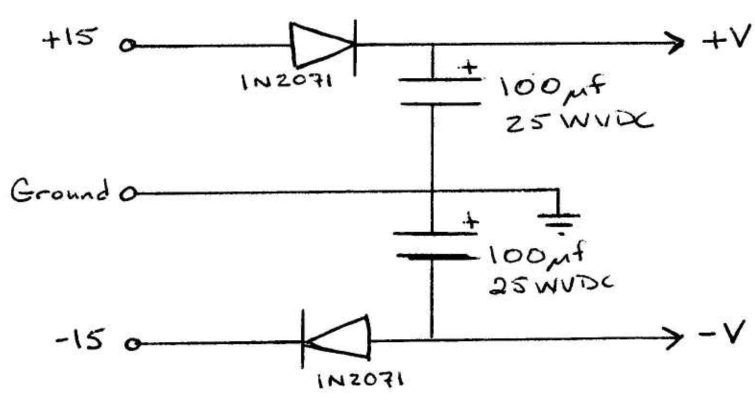
Switching Operations

Figure 18

A complete diagram of the switching operations is given in Fig. 18. Switches S1 and S2 are two pole double throw slide switches while S3 is the ten pole six position rotary switch.

<u>Power Supply Protection</u>

The omnipresent danger to systems which require external power is that the user will connect it incorrectly to his power source. In operating the simulator, the user is asked to supply three power leads: +15 V, -15 V, and ground. To provide minimal protection and filtering, the diode capacitor combinations of Fig. 19 are placed at the power inputs. This makes that actual internal supply voltages more nearly 14.3 V when ± 15 V is the input. However, ± 15 is easier to obtain and the only devices for which the operating point might be adversely affected by the 0.7 V drop, the μA 702's, are Zener regulated.



Power Supply Protection

Figure 19

VI. SYSTEM PERFORMANCE

Testing is an integral part of any design process and each circuit described in the preceding section was tested, modified, and retested many times. Rather than give results of individual circuit tests, this section will present the results of tests on the entire system as the various functions are programmed. This will give an indication of the performance of the individual circuits involved although any circuit which primarily limits the performance of the system will be mentioned specifically.

Main Signal Path

With all function switches in the off position, the main signal path should act as a 500 KHz low-pass filter with good frequency response as low as 20 Hz. A series of tests with input signals of various amplitudes produced the frequency response curve of Fig. 20.

For input signals of amplitude 1, 1.5, 2, and 2.5 V peak-to-peak, the response of the main signal path is so similar as to make it fruitless to plot a family of curves for these inputs. At input amplitudes of 0.1, 0.2, and 0.5 V peak-to-peak, the curve of Fig. 20 is confirmed between 20 Hz and 500 KHz. Past 500 KHz, these signals are so swamped by extraneous signals at 10 MHz and 100 MHz that seem to inhabit the author's laboratory at all hours

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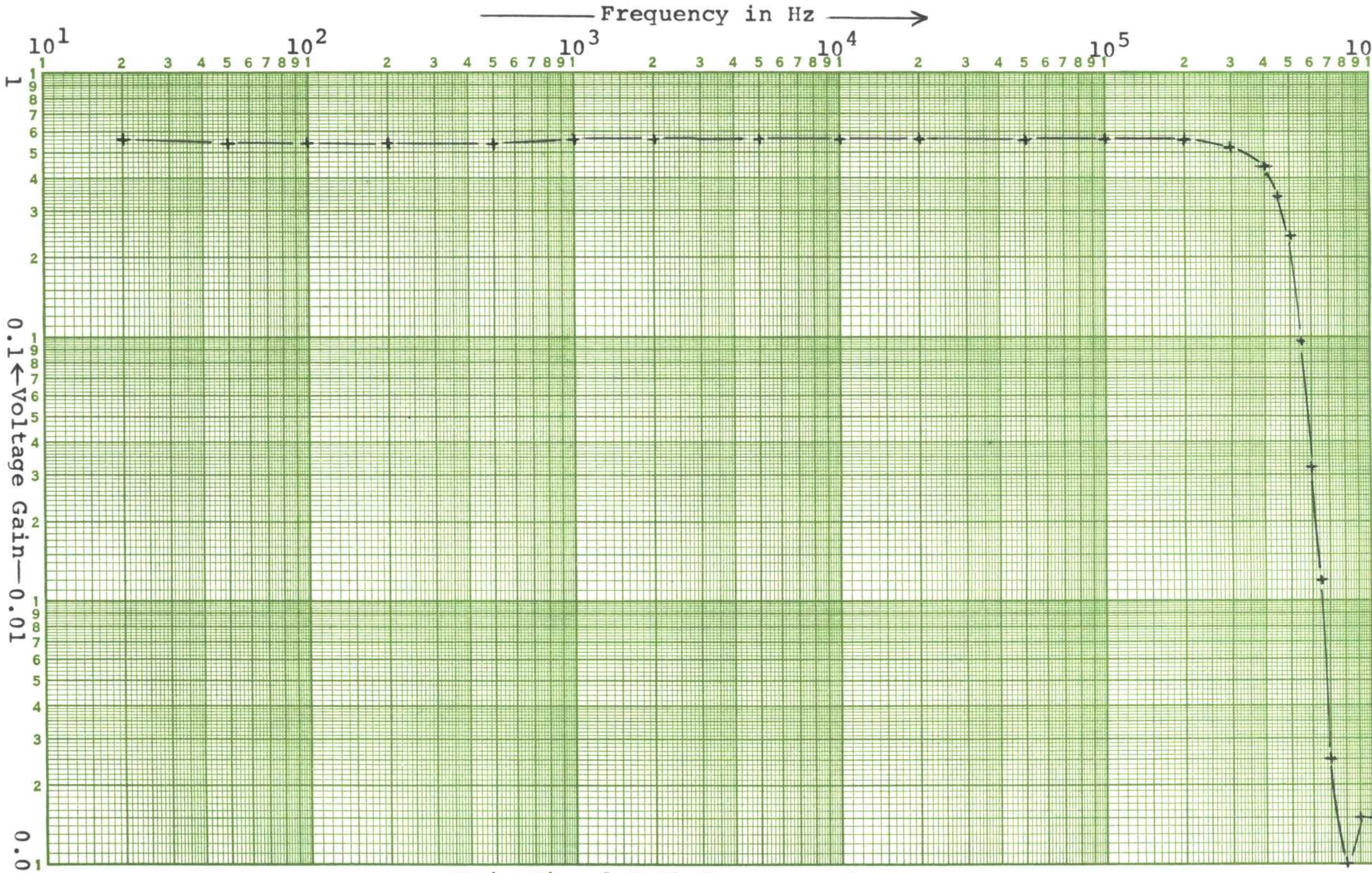

Main Signal Path Frequency Response
Figure 20

that they were unresolvable. (Oscillation of the μA 702's in the noise generator was the suspected source of the 10 MHz and 100 MHz. Measurements at various hours of the day, with the simulator and all other devices off, with several oscilloscopes, and by various different persons demonstrated that the 100 MHz and 10 MHz signals were in fact produced by some unknown external source. The 100 MHz signal is believed to be an FM radio station. The origin of the 10 MHz signal has been the subject of wild speculation.)

The maximum signal amplitude (in the 500 KHz filter pass band) that may be applied to the simulator is 2.5 V peak-to-peak if minimal distrotion is desired. This amplitude is dictated by the 500 KHz filter (Fig. 7). More specifically, the second amplifier stage of the low-pass section nears cutoff at this amplitude. The amplitude limiting is provided at this point so that additive functions introduced farther along the signal path will not cause clipping.

The main signal path thus acts as a 500 KHz low-pass filter with 5 to 6 dbV insertion loss. The relative gain at the cutoff frequency is -6 dbV but it will be noted that this is the point at which the gain variation across a 10 KHz band (the anticipated user bandwidth) approaches 3 dbV.

Wideband Noise

With the output of the noise generator added to the main signal path (cf. Fig. 13b), further spectrum analyser tests showed that a relatively flat spectrum is still obtained above 10 KHz.

In the absence of sophisticated noise measuring equipment, one simple estimate of the character of the noise as observed at the simulator output was made. After extended viewing of oscilloscope traces, the maximum noise amplitude observed was 2.5 V peak-to-peak. This was taken to be roughly the maximum amplitude obtained.

Fading

The SG 1496 balanced modulator is capable in static tests of attenuating the input signal in excess of 30 dbV. A dynamic test of the fading function is difficult to perform because it is a random process. Thus, the technique of extended viewing of oscilloscope traces was again employed to develop an estimate of system performance. On this basis, it appeared that about 26 dbV maximum relative attenuation was given input signals of one to two volts peak-to-peak amplitude.

More easily observable is the small additive change, caused by the modulator unbalance, as the fading drive takes the control input through the switching point. This jump is about 250 mV in amplitude and does not cause apparent distortion of the signal. Because the control input has a linear range, this level shift tends to follow the control input frequency and is thus outside the nominal bandwidth of the simulator.

A drawback is associated with the additive level shift in the fading function. As stated, it will have a characteristic frequency around 10 Hz and can be separated from the signal by the user. But, one intention in the use of the balanced modulator was to produce some uncertainty in phase at the output for coherent user systems. The clever user will thus have a key for his estimate of received phase if he should notice this effect. Nonetheless, it was not considered useful to provide a potentiometer adjustment in place of the fixed balancing network of Fig. 11.

The critical adjustment for fading is the offset voltage of the noise generator (Fig. 4) which has experienced a gain of several thousand by the time it reaches the output of the X10 amplifier. The dc bias at the output of the X10 amplifier has proved to be repeatable in numerous trials extending over a period of several days after it was first set. Thus, the fading function should require no user adjustment.

Swept Tone Interference

Probably the easiest function to evaluate is swept tone interference. The swept tone appears at the output of the simulator at an amplitude of approximately one volt peak-to-peak. Its maximum period is 48 microseconds (or roughly 20 KHz) and minimum period 1.6 microseconds (about 600 KHz). The time required for the transit from minimum to maximum frequency or vice versa is 9 seconds with a waiting time at either end of 3 seconds. Despite this long sweep time, it must be remembered that this amounts to an average of only 180 milliseconds to pass through a 10 KHz band.

The square wave output of VCO #1 was chosen for use as the swept tone so that energy could be supplied to a wider spectrum for a given amplitude in hope that the effect of the relatively rapid passage through user signal bands could be partially offset by having more transits per unit time at higher frequencies (albeit not all at the same amplitude). Along with the VCO square wave, one also gets a more or less unwelcome addition. Because the VCO square wave is derived from a logic chip with nanosecond risetime, ringing is a definite problem to be considered. For any VCO loading tried, about 8 cycles of decaying 50 MHz were produced at the simulator output.

Radiation is so real and pressing a problem at 50 MHz that it was more practical to let the 50 MHz run rampant in the simulator and allow the simulator case to shield the external world while the user can take steps to filter it out if his system does not already do so. This option was made possible by the fact that the maximum amplitude of the 50 MHz disturbance is one half volt and 50 MHz is two decades above the highest nominal user frequency.

Impulse Noise

The impulse train at the output of the simulator consists of 5 microsecond positive pulses 0.6 V in amplitude varying in repetition rate from 1 to 2.4 milliseconds (about 400 Hz to 1 KHz) with an average period of 1.6 milliseconds (600 Hz). No ringing is present on these pulses because they are produced by the multivibrator of Fig. 17. Smaller negative pulses (commonly not more than 0.1 V) are observed between the positive pulses as the multivibrator is pushed toward breakdown by positive swings of the VCO output.

Frequency Shifting

As previously explained, the output of the simulator is a randomly varying square wave, of nominal frequency 600 KHz, amplitude modulated by the lower sideband, taken about 600 KHz, of the user's signal. Tests of the frequency shifting function were conducted with and without an external 500 KHz low-pass filter (similar to that used in the simulator) used to remove the upper sideband.

Carrier feedthrough at 600 KHz from modulator 1 was found to be minimal at the output of the simulator when

VCO #2 was disconnected. With VCO #2 in the simulator, its carrier feedthrough became a problem as did the 50 MHz encountered in swept tone interference. The random 600 KHz can be suppressed to 50 mV peak-to-peak by balancing modulator 2 (a planned screwdriver adjustment) but the 50 MHz still appears at the output with an amplitude up to 500 mV. Again, the 50 MHz was left as a filtering problem for the user, for the reasons previously discussed.

When the simulator output was put through an external 500 KHz filter, two interesting effects were observed. First, 20 Hz input signals can be recovered relatively undistorted when frequency shifting is programmed (these are, of course, tremendously attenuated and unshifted). Second, input signals of up to 20 v peak-to-peak can be handled by the simulator without destruction (all amplitude discrimination is lost above 4 V peak-to-peak).

Tests with a spectrum analyser disclosed that the frequency shift obtained has a range of 30 KHz (with rapid variations over 10 KHz). The average shift is, of course, dependent upon proper setting of the noise generator offset. Because this setting has been shown to be repeatable, the dependence is not considered a critical problem.

VII. CONCLUSIONS

The simulator project spanned a period of some 18 weeks of which the first five were spent on formulation of a general plan and simplification of the proposed block diagram. Designing the simulator proved to be, above all, an introduction to the practical world. Large blocks of time were invested in scouring commercial literature for devices capable of providing the best compromise between performance and cost. A certain familiarity with the professional literature was also acquired in the search for design ideas to realize various system blocks.

In the end, the circuits built for the simulator were neither unique nor particularly exotic. None of the concepts employed are too far beyond the typical undergraduate and are, hopefully, presented in a manner comprehensible to such a reader.

Many of the simulator circuits are derived from the principles of the early part of the electrical engineering education, although dredging some of these ideas from the memory was often a frustrating experience. Others, while not so basic, yield readily to straightforward analysis.

In all, the most striking result of the simulator project is the manner in which the author's advisor's dictum was demonstrated. "You will find that you are going to waste a lot of time in this business," was painfully borne out in hours of aggravation and near despair when failure to consider something so basic as a turn-on transient spoiled a design or a model was insufficiently detailed to predict the behavior of a circuit. Nonetheless, it is felt that a successful design was finally achieved. Certainly, a large file of ideas and devices has been acquired which, while not applicable to the simulator, may well be of use in the future.