AN IMPULSE NOISE SUPPRESSOR
THAT OPERATES
IN THE TIME DOMAIN

By

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PREFACE

Noise reduction and suppression is a subject of extremely widespread interest and a tremendous volume of literature exists concerning it. The more limited subject of the reduction of impulse noise has largely been approached from the point of view of the suppression of impulsive radio interference. Some effort has been put forth in attempts to reduce the amplitude of this interference below the amplitude of the signal, but this effort has been concentrated in that part of the radio receiver near the antenna. Few attempts have been made to reduce the amplitude of impulse noise or to mitigate the effects of impulse in the purely audio section of an audio system. With one exception, such work as has been done in this area has been concerned with limiting the noise amplitude to the maximum signal amplitude. This one exception is the work of Mr. D. T. N. Williamson. This study is an attempt to improve his time gate and to show that its use may be extended from the gating of impulse noise due to record surface damage to gating any impulse noise which has a frequency spectrum wider than that of the signal.

Indebtedness is acknowledged to Dr. Herbert L. Jones for his guidance and encouragement; Mr. Paul A. McCollum for his assistance in regard to certain circuit features; and to Mr. H. H. Scott, Maynard, Massachusetts and Mr. D. T. N. Williamson, Edinburgh, England for information concerning their circuits and for their kind efforts to help in this study.

TABLE OF CONTENTS

LIST OF FIGURES

LIST OF TABLES

CHAPTER I

INTRODUCTION

The problem of noise reduction appears in practically all electronic systems. How much of a problem it is depends upon the signal and noise power levels and the system requirements as to the signal-to-noise ratio. The higher the noise level, the lower the signal level and, the more stringent the signal-to-noise requirements, the more difficult the system design becomes in respect to noise. Audio systems are one group of electronic systems in which this problem is of importance and one in which the problem appears in all degrees. For the purposes of this thesis an audio system may be defined as "A transmission channel originating with some source of audio-frequency intelligence and terminating with some electroacoustic transducer, such as a loudspeaker." To some extent the noise problem involves the response of the human ear. But to extend the definition of the transmission channel to include the path from the transducer to the ear would allow the inclusion of noise originating in this area. While the sound of a woman's voice during the playback of Beethoven's Ninth Symphony may be considered noise, such consideration is outside the subject of electronics and beyond the scope of this thesis.

Noise may be considered as any system-output power which tends to interfere with the use of the applied signal, with the exception of harmonics and intermodulation products. (1). Noise, therefore, is due to spurious signals originating within the system, but not to distortions resulting from system nonlinearities, though it may be modified by these system characteristics. There are extremes to the infinite variation of noise allowed by this definition. At one extreme is low-level noise which covers the whole frequency-spectrum of interest in the system under study and which is of equal amplutude at all frequencies. This is sometimes referred to as "white" or "background" noise. (2).

The reduction of relatively low-level background noise has been achieved more or less successfully by various means. It is reduced at its source by the use of low noise components, better recording media (the replacement of shellac by vinyl plastic in records) and by proper design methods. It may be attenuated in the system by the use of noise-suppression circuits of a nonlinear nature. The circuits used are not based upon any principle of recognition and elimination of the noise itself, but upon certain characteristics of human hearing. These characteristics make it possible to eliminate, or at least attenuate, all low-level signals (including the low-level audio signal being transmitted) without introducing new and more objectionable noise and distortion.

The principle of considering the response characteristics of the ear as the eventual receiver of the audio signal has only rarely been applied to the problem with which this thesis is concerned...the attenuation of impulse noise. From this point of view, the most general problem is not that of reducing impulse noise, but that of reducing or modifying it so as to be as unobjectionable as possible. Of course, if complete filtering of impulse noise without modification of the audio signal were possible, that would be the perfect solution, But, insofar as this may be impossible, filtering and modification in light of the characteristics of the ear will lead to the optimum solution. Fortunately, it is possible to almost completely filter impulse noise, but even so, the characteristics of human hearing indicate why it is necessary to do so.

In order to design a suitable filter it is necessary to determine the characteristics (amplitude, frequency and time) of the most common types of noise and how these characteristics are modified by the transmission channel. Though system nonlinearities do not contribute to the noise they do modify it, usually for the worse. It is possible to filter radio noise in the radio frequency or intermediate frequency stages of a superheterodyne receiver. But as it is the purpose of this study to design a single filter which will be useful for all impulse noises that may enter any audio transmission channel, and as the audio amplification stage is common to all audio transmission channels, the filter will be placed in the audio-amplification stage. Therefore, modifications of the impulse noise by stages prior to the audio amplifier will have to be taken into account.

There are two principal types of audio systems and a type of impulse noise more or less characteristics to each. One of the systems involves the radio-frequency transmission of the intelligence. Impulse noise is introduced into this system at the receiving antenna and is the result of electrical discharge in thunderstorms (sferics), automobile ignition radiation and other radio-frequency noise sources. The first of these is the most common and bothersome and will be considered here at length. The second system involves the recording and playback of audio signals. Impulse noise here is the result of imperfections in the recording media, dirt on the record surface or damage to the recording. The latter commonly takes the form of scratches on a record surface. This source of noise will also be considered as typical.

Assuming that everything else is constant, the amount of interference which may be tolerated and, therefore, the necessary signal-to-noise ratio, depends upon the audio signal being transmitted and the use for which it is intended. When the transmission of voice communications is involved and intelligibility is all that is demanded, considerable noise may be tolerated. On the other hand, any audible noise may be considered objectionable or, in the extreme, intolerable in the reproduction of music. The noise level in the first case is usually of much greater amplitude and repetition rate than in the second, and the same noise-suppression circuit proves useful in both cases. The filter or noise-suppressor circuit arrived at could be called an amplitude-sensitive, frequency-triggered time gate.

CHAPTER II

IMPULSE NOISE

Atmospheric Noise

The two types of impulse noise to be herein considered as typical (static and record clicks) are also the most familiar. While most people merely mutter to themselves when radio static is heard and either bear it with some irritation or turn off the radio, it is a real and immediate problem to those who have to communicate by radio; the military, aviation industry, and those advertisers whose public has decided that a good book is preferable to static.

The source of most static is electromagnetic radiation originating from electrical discharges within and from that type of cumulonimbus cloud known as a thunderstorm. These discharges are referred to as atmospherics or "sferics". When this radiation impinges upon an antenna, that part which falls within the bandwidth of the receiver connected to that antenna is amplified and distorted by the receiver and ultimately gives rise at the speaker to audio noise or static. The following discussion is largely a summary of several of the published articles of Mr. S. V. C. Aiya (3, 4, 5) of the University of Poona; Poona, India.

Thunderstorms are of widely variant types, the characteristics of any one depending upon the location, time of year and many variables of weather. Each is a highly dynamic phenomena which phases of buildup, maturity and decay. The electrical discharges associated with them may also be expected to be variable. Any analysis and general discussion of the storms and the associated phenomena must be of a more or less statistical nature and carried on in terms of median values of the parameters involved. Aiya refers to these as typical or "Idealized statistically valid representation(s)". A distinction may be made between two typical types of discharge, intracloud and cloud-to-ground.

When the electrical field in a cloud exceeds the disruptive strength of the dialectric, an electrical discharge occurs with resultant visible radiation called a lightning flash. At the peak of its activity a storm may radiate well in excess of ten flashes per minute. A typical flash consists of four strokes. Each stroke probably starts with a pilot streamer which travels with a speed of approximately 10^7 centimeters per second and establishes an ionized path for the rest of the stroke. (6). This is followed by a leader streamer, herein referred to as a leader, which travels at approximately 10^9 centimeters per second. Following the leader is a recoil of low intensity and long duration when the discharge occurs in the cloud only. When the discharge is to ground, there is a return stroke of great intensity and short duration in place of the recoil occurring in the intracloud discharges.

All leaders in intracloud discharges are discontinuous and consist of a number of steps following one another in time sequence. Each step in such a stepped leader is of short duration, normally lasting less than one microsecond. A step is a high-intensity discharge over a distance of from 40 to 100 meters traveling from the end of the previous step to the head of the pilot streamer. The rate of change of current in such a step has been estimated as high as 10^{10} amperes per second.

The step path acts as a radiating dipole in free space. The growth of current during the discharge in a step gives rise to the radiation of an impulse, and the series of steps in a stepped leader therefore radiates a train of impulses. This radiation from leaders within the cloud is largely limited to frequencies above 2.5 megacycles per second. Aiya (5) has theorized that the ionized region through which the discharge takes place forms a pocket within a negatively charged cloud, and that this pocket acts as a waveguide with a cutoff frequency of 2.5 megacycles per second and is responsible for the limitation of the radiation spectrum. The recoil is of long duration and low intensity and therefore radiates little significant power at radio frequencies.

Ordinarily, strokes originate in the negatively charged regions of the cloud and involve a length of one kilometer. $1/\sqrt{2}$ times this length remains ionized after each stroke and the process continues, one stroke following another to the center of the negatively charged region. This involves a distance of three kilometers. Thus, a typical intracloud flash may be expected to consist of four strokes and the stepped leader of each stroke to be composed of from ten to fifteen steps. The process varies largely, and, in rare instances, as many as twenty-eight strokes have been observed. (7).

Figure 1 is that given by Aiya (3) for the time relations of a typical flash made up of four strokes. The strokes (and therefore the stepped leader) are separated by 0.04 seconds, and the leaders consist of steps of less than one microsecond duration separated in time by 74 microseconds. The actual time taken by the radiated interference is less than forty microseconds. Why such a short duration of noise should cause the great interference-effect observed will be discussed later.

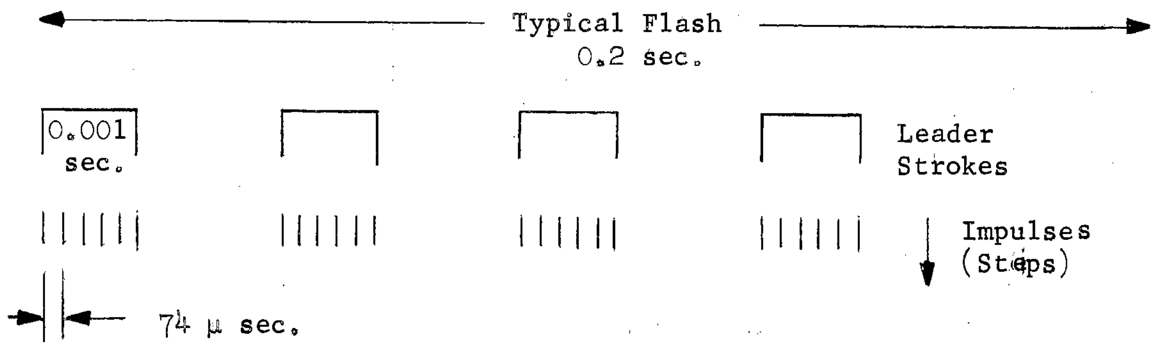

Figure 1

Typical Time Relations in an Intracloud Flash

When the cloud base-to-ground distance is of the order of 2.2 kilometers, conditions are favorable for a cloud-to-ground stroke. The mechanism of such a stroke differs somewhat from that of an intracloud stroke, and the radiation, therefore, also differs.

Generally, in cloud-to-ground discharges, only the first leader is stepped and the recoil or return stroke is of great intensity and of short duration. The leaders subsequent to the first leader are not stepped and are often referred to as "dart" leaders. Aiya (5) says;

> The return stroke, the recoil and unstepped portion of the dart leaders do not radiate any significant power in the short and medium wavebands. It follows that the steps in the stepped leaders are the principle sources of noise in the short and medium wavebands.

The cloud-to-ground leader takes place in three phases, not all of which are necessarily present. The three phases correspond to the discharge through three regions of the cloud and the volume between it and the ground. These regions are illustrated in Figure 2. (5).

Stage one of the cloud-to-ground stroke takes place in the nega-

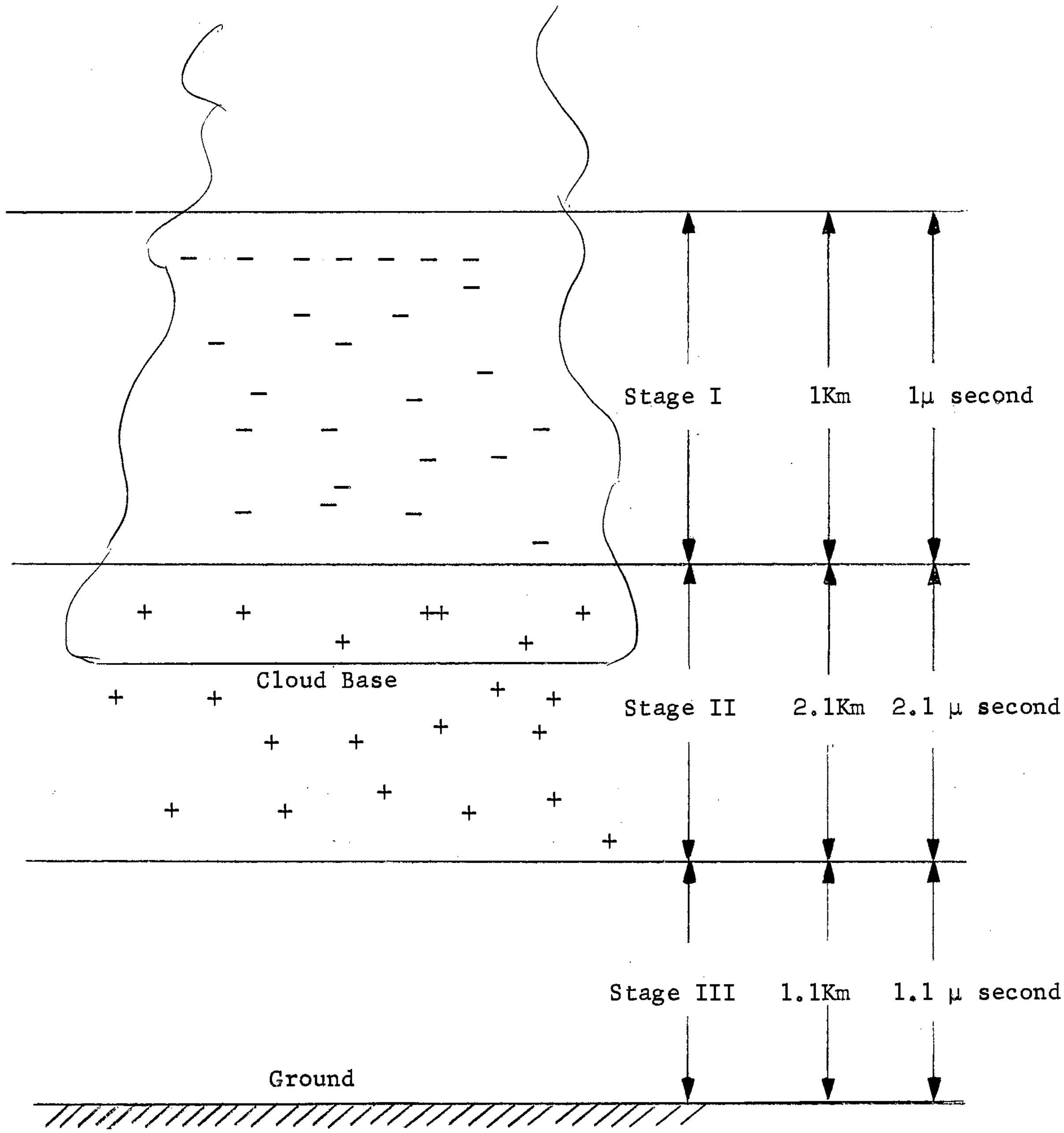

Figure 2

First Leader to Ground and Its Stages

tively charged portion of the cloud and is the same as the intracloud discharge described above.

In stage two the stroke progresses through the positively charged lower section of the cloud and the air below the cloud. The steps are longer here, lasting up to 2.1 microseconds. The discharge "pocket" exists in a region of positive ions and evidently does not have the waveguide characteristics of the pocket in the negatively charged volume, as there is considerable radiation below 2.5 megacycles in the medium wavelength spectrum.

Stage three takes place in the uncharged free air directly above the ground. The resistivity of the air is higher here than in the charged regions above. The rate of change of current in the steps is much lower (1/10 of that above) and the steps and the intervals between them are shorter. In leaders subsequent to the first, this stage and most of stage two are absent; that is, the leaders subsequent to the first are dart leaders having no steps in these regions.

When the base of the cloud is much lower than 2.2 kilometers, all or most of stage two and all of stage three are missing. Therefore, stage two contributes most of the medium wavelength noise, and that is contributed principally on the first stroke. A typical flash from the base of a low cloud to the ground would have the time diagram shown in Figure 3 as given by Aiya. (5). Most of the noise is in the first stroke which lasts longer than strokes of intracloud flashes.

Aiya bases most of this theory upon the data and theory presented in various publications during the 1930's by Schonfield and his coworkers. The mechanisms involved are not fully understood, and other theories have been presented. Notably, Bruce and Golde (7) rejected

Schonfield's theory that strokes subsequent to the first travel from deeper within the cloud to the origin of the previous stroke and then over the previous stroke's "aged" or partially deionized path. They advanced the idea that original path remains fully ionized, and that strokes subsequent to the first are from other charge centers within the cloud to the origin of the first stroke and then over its ionized path to the ground.

In any event, the theory is designed to explain observed data which includes the electromagnetic radiations in the medium and short wavebands. These radiations remain the same and are typically as shown in the figures, independent of the theory of their origin. Any given flash may vary considerably from the typical. Large numbers of strokes may occur in a flash, but the relative frequency of occurrence diminishes as the number of strokes increases.

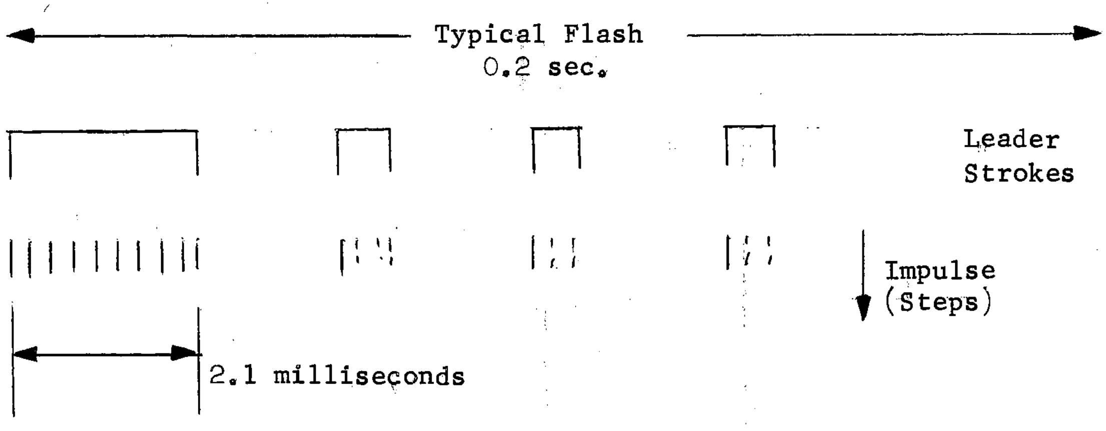

Figure 3

Radiation from Stage II of a Cloud-to-Ground Flash

Noise Modification by the Superheterodyne Receiver

A radio receiver may be considered essentially a bandpass filter. This viewpoint is exemplified by Mr. B. T. Newman's (8) definition of the ideal receiver.

"...Terminus of a communications system which employs a transmission medium wherein there exists an extremely large number of unrelated signals combined with an overwhelmingly large variety of interference. The prime function is to filter from this confusion one particular signal and decode it in a form understandable to the observer or listener. Ideally, the receiving antenna feeds a rectangular filter suitably located in the frequency domain and having an acceptance characteristic exactly suited to the desired signal. The filter is terminated in a decoding device which will ignore all perturbations caused by co-existance of interference in the filter. The decoding device output should be an exact replica of the original intelligence encoded at the transmitter."

Unfortunately, even the most well-designed receivers depart greatly from this ideal. Impulse noise, which could be clipped after passage through an ideal receiver and retain little energy in infinitesimal pulses with fundamental frequencies far above the range of human hearing, is so modified by an actual receiver that the decoding function is often incapable of presenting a useable facsimile of the original intelligence. This is a result of signal modification within the receiver due to several receiver characteristics.

(1) Oscillatory circuitry: The IF and RF amplifiers of the receiver are coupled by oscillatory circuits, and any high-amplitude impulse triggers them into a decaying oscillation. It has been shown that, as a result of this, the impulsive signal delivered to the detector is a function of the transmission characteristics of the receiver viewed as a filter and the total impulsive energy only. It is independent of the bandwidth, duration or shape of the pulse as long as the pulse duration is shorter than the receiver bandwidth. (9). The final duration of the impulse at the

detector is inversely proportional to the amplifier bandwidth, while the amplitude is directly proportional to the receiver bandwidth. (9). Thus, the time integral of the pulse power output is independent of the receiver bandwidth.

(2) Amplifier Nonlinearities: Amplifier and circuit nonlinearities cause intermodulation of the pulse noise and the signal. Limiting of the noise pulse late in the circuit cannot remove this distortion from the signal. For reasons of stability, the input to the filter is of wide bandwidth. (8). Interference outside the frequency spectrum of the desired signal therefore enters the receiver and mixes and intermodulates with the signal, thus increasing the effective distortion without the full power of the noise passing through the receiver.

The amplitude of the noise pulses may be sufficient to cause grid-limiting and plate-saturation in the various amplifiers of the receiver. During the recovery time of these circuits, the signal level is depressed and there is a resultant loss of signal, even though no noise is present. (8). This may be of considerable importance in the audio amplifier where grid current charges the interstage-coupling capacitors, with the result being that a stage may remain cut off for a considerable period of time.

(3) Detector Nonlinearities: The diode detector used in most receivers is a nonlinear device with the transfer characteristics of Figure 4. (9). The detector discriminates against low-level signals; that is, high-amplitude signals are attenuated less than low-amplitude signals. (9). When the noise is of lower amplitude than the signal, the result is an increase of the signal-to-noise ratio at some cost in fidelity. But when the noise is of high amplitude, the noise is attenuated less than the signal,

and the signal-to-noise ratio suffers accordingly. The curvature of the transfer characteristic is the greatest for low signals, so the effect is worst where it can least be afforded.

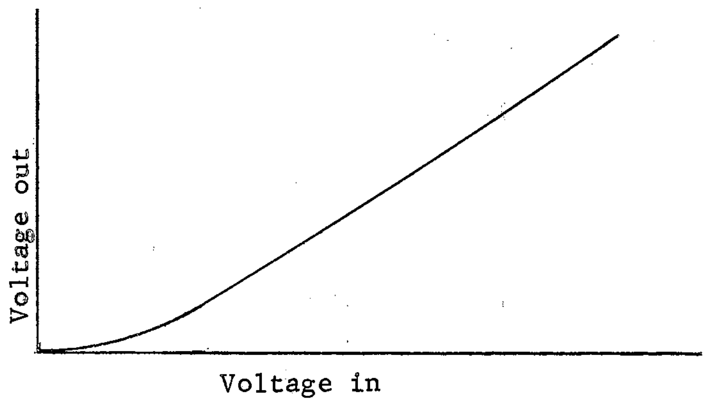

Figure 4

Diode Transfer Characteristics

Almost all receivers are equipped with an automatic volume-control (AVC) circuit in order that the carrier-amplitude be kept relatively constant at the detector. The RC filter network in the detector circuit is then designed to avoid amplitude-clipping of the modulating signal. This is achieved by making the filter time constant small enough so that the capacitor can discharge from the peak voltage of one carrier wave to that of the next when the peak values of the varrier voltage are decreasing at the greatest rate. This occurs when the highest modulating frequency to be passed is at its greatest average amplitude and an individual modulating waveform is decreasing in instantaneous value through zero. The filter design equation is;

$$1/RC = m\omega_m \qquad (1)$$

where m is the modulation index and does not exceed one, and ω_m is the highest modulating frequency to be passed by the receiver. (10).

AVC circuits are slow-acting and individual noise pulses do not appreciably affect the AVC action. The pulses are therefore of much greater amplitude than the maximum instantaneous signal voltage after detection. As a result, the pulses are lengthened in passage through the detector filter and, if they occur at a frequency or repetition rate much greater than the maximum modulating frequency for which the filter is designed, the pulses are run together and form one long pulse with ripple of the noise frequency riding it.

(4) AVC Action: AVC circuits perform their function by feeding back a control voltage to earlier stages in the receiver. This control voltage is obtained by filtering the audio-voltage output from the detector. A series of high amplitude pulses need occupy only a very small proportion of any time period in order to add significantly to the average control voltage. Such an increase of AVC voltage acts to decrease the receiver gain. (10). It is probable that the pulse amplitude has already been limited by tube saturation, so the decreased gain has little or no effect upon the pulse amplitude reaching the detector while the signal gain is noticeably reduced. Thus, the AVC action is such as to reduce the signal-to-noise ratio for impulse noise.

The degree to which a pulse or series of pulses is modified in passage through a given receiver is, of course, dependent upon the design of the individual receiver. To some extent, all of the distortions discussed above will apply. The result is that an individual pulse is considerably lengthened. The pulses radiated from a stepped leader are separated in time by an average of 74 microseconds. This corresponds to a fundamental frequency of 13,000 cycles per second ($1/74 \times 10^6 \cong 13,000$). Superheterodyne receivers are rarely designed for a modulating frequency exceeding 7,500 cycles per second. (11). The pulses of the stepped leader are therefore run together into a single pulse approximately one milli-second long in the case of intracloud stepped leaders and 2.1 millisec-onds long for the cloud-to-ground stepped leaders. If these pulses are regarded as corresponding to one-half a complete 100 percent modulated envelope, it follows that the equivalent modulating frequencies will include 500 or 250 cycles per second and their harmonics. Thus, pulses of extremely short duration with fundamental frequencies in excess of 250 kc are rendered audible by the action of the receiving filter.

Phonograph Record Noise

While impulse noise of radio origin is of extreme military and commercial importance, that which is experienced in music playback is of little importance except as an irritant to the listener. Many music lovers, however, are more than slightly irritated, as a perusal of the letters-to-the-editors section of any of the magazines devoted to record reviews and high fidelity will prove. The impulse noise experienced in the playback of a damaged phonograph record is familiar to everyone and is referred to in terms of "clicks", "pops" and other such onomatopoetic words.

Though of short duration and taking only a minute portion of the playing time, these noises are extremely annoying, and some people con-sider a record with but a few such noises to be unacceptable. The noises result from imperfections in the record surface caused by dragging some sharp object across the record surface (more often than not, the playback stylus), or the record rubbing over a small particle embedded in the record envelope. The imperfections are for the most part less than 1/200th of an inch wide.

The peripheral velocity with which the record passes under the stylus is lowest toward the center of the record where the circumference of the record groove is the smallest. Thirty-three and one-third rpm records are rarely recorded to within $2\frac{1}{2}$ inches of the center of the record. The peripheral velocity here is;

$$v = \frac{rpm}{60} 2\pi r = 8.7 \text{ inches per second} \qquad (2)$$

If the scratch or other surface imperfection is 1/200th of an inch wide, it lasts for;

$$T = \frac{D}{v} = 0.000575 \text{ seconds} \qquad (3)$$

where D is the width of the scratch. This corresponds to a fundamental frequency of 1740 cps.

The frequency spectrum of the electrical impulse resulting from the stylus passing over such an imperfection is dependent upon the shape of the imperfection and upon the mechanical and electrical characteristics of the system made up of the record, the stylus assembly and the voltage generator to which the stylus is attached. The lowest frequency present can be no lower than the fundamental frequency corresponding to the width of the deformation or imperfection in the record surface. Until worn down by repeated playings, the leading edge of the scratch caused by dragging a sharp object across the record surface is apt to be rather abrupt. This results in concentration of the higher noise frequencies at the leading edge of the impulse. In any case, such damage often results in noise which is impulsive in nature and which contains significant power at frequencies beyond the range of human hearing.

The mechanical characteristics of the stylus assembly and record material are rather involved. In general, they form a mechanically resonant system at the point of contact. (12). In this mechanical system the mass of the stylus assembly together with the record material displaced by the pressure of the moving stylus is analogous to inductance in an electrical system. The stylus assembly is designed to return to a center position when not under pressure and the displaced record material tends to spring back against pressure. In mechanical systems this restoring force is referred to in terms of a spring deflection constant in units of force per unit displacement. (13). When speaking of stylus assemblies, it is more common to use the reciprocal of this constant which is called compliance and which is analogous to capacitance in electrical systems.

The motion of the system becomes mass-controlled above the resonant frequency, and the output falls rapidly with increasing frequencies. High quality transducers are designed with low moving mass so that this resonance falls at as high a frequency as possible and the resonance itself is friction damped. Nevertheless, peaks up to 14 db or more at frequencies exceeding 20 kc have been found in even the best units. (14). A sharp deformation in the record surface will shock-excite the system with the result that there will be a considerable output from the best quality transducers at frequencies above 20 kc. This is above the normal upper limit of human hearing, and only one recording company claims to record frequencies above 20 kc. (15). Therefore, any impulsive output at frequencies exceeding this will be noise.

Response of the Human Ear

The audio system output is intended to be heard by human ears. If it is impossible to completely filter out impulse noise, then the aim should be a type of filtering which results in the least irritating noise and smallest loss of intelligibility. The characteristics of the hearing mechanism are therefore of some interest when designing such a filter.

At the relatively high audio frequencies contained in the pulses in which we are interested, the ear acts in accordance with the approximate logarithmicity of the Weber-Fechner law of sensation intensity; that is, a seeming linear increase in loudness requires a logarithmic increase in the actual sound intensity. (16). In this manner the ear is capable of sensing a tremendous range of sound intensities exceeding 130 db at 1,000 cps without overloading. This may be due to the manner in which sound vibrations are transmitted through the middle ear by the small bones therein. As the actuating force transmitted to these bones by the eardrum increases, their stiffness or resistance to motion also increases. Moreover, the recovery time of these bones is large. As a result, the ear acts as an integrating device for sounds of short duration and remains relatively insensitive to low-level sounds for some time after a pulse is over. Pollock (17) states;

> The recent experimental literature on the aftereffects of auditory stimulation show that the threshold changes are obtained for intervals many times the critical duration of 55 milliseconds even for moderate... intensity sounds.

The critical duration referred to above by Mr. Pollock is another auditory phenomena; that of the seeming persistence of a sound for a period after the sound has been terminated. Again, quoting Mr. Pollock (17);

> After a sound is terminated there are aftereffects of excitation of the auditory system which persist in time. The various methods (of investigation) all point to an estimate of about 55 milliseconds for the duration of the persistence of auditory sensation.

One result of these hearing characteristics is that the seeming loudness of an impulse noise is much greater than what might be expected from the actual pulse power. But, there is another result which is advantageous. Even though the duration of an impulse noise is magnified by the persistence of auditory sensation, complete silences of short duration are hidden by the persistence of sensation of the desired signal. Due to this and to the great redundancy of human language and music, a considerable portion of a communication may be blanked out without loss of information if the blanking periods are considerably less than 55 milliseconds in duration. The only real loss would be if the blanking pulse coincided with a signal transient.

The auditory persistence is a decaying phenomena, and the rate of decay seems to be logarithimic in nature. Therefore, the original sound intensity seemingly persists for a much shorter time than 55 milliseconds. Mr. D. T. N. Williamson states that interruptions are not noticeable, provided that the time does not exceed 300 microseconds. (18). Experiments by this author with magnetic recording tape have shown that a barely discernable effect is present when blank tape of three milliseconds duration is inserted in prerecorded tapes. It would seem that one millisecond should be a sufficiently short blanking period so as to be practically unnoticeable, and that longer periods of silence, though not noticeable, would be preferable to the noise that is banked.

CHAPTER III

NOISE FILTERS

Any number of filters have been designed for the specific purpose of suppressing or limiting impulse noise in superheterodyne receivers. Few have been strictly designed for use in audio amplifiers.

Filter Types

Filters are commonly thought of as acting in the frequency domain. With greater generality, circuits which act selectively on the parameters of amplitude and time may also be considered as filters operating in the amplitude and time domains respectively. (19). Such circuits are more often referred to by other names; limiters, suppressors, coincidence amplifiers, etc. The use of these particular names has perhaps tended to hide the essential relationships of the circuits involved.

Such type-classification of filters into those that act in the frequency, amplitude and time domains is synthetic and idealized. Actual filters may combine aspects of two or more of these types and may operate in any combination of the domains simultaneously. It is interesting to note that the most frequently used type of filter, that in the frequency domain, is the one with which it is most difficult to approach the ideal.

Frequency filtering is most effective where the undesired perturbations are confined to a relatively narrow frequency-spectrum or where they exist at the extremes of the frequency spectrum of the desired signal. An idealized filter of this type would have a frequency-versus-amplitude characteristic similiar to that of Figure 5 where the frequencies between f_1 and f_2 and those above f_3 are completely blocked, while the rest are passed without attenuation.

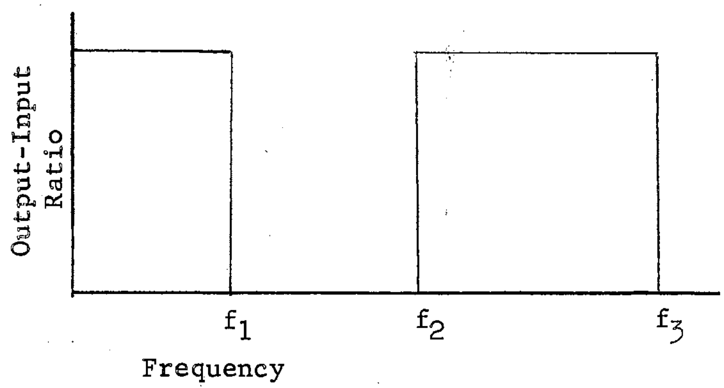

Figure 5

Ideal Output Characteristics in the Frequency Domain Filter

Amplitude filtering is effective where the noise is of a considerably different amplitude than the signal. Its greaters use is in limiters and squelch circuits. A filter of this type might have amplitude characteristics such as those in Figure 6. Time filtering may be used when the undesired perturbations are of intermittent duration and have a higher amplitude or wider frequency spectrum than the signal. This is often referred to as "gating". A time filter has characteristics similiar to those of Figure 7 in which the signal is completely blocked during the times from t_1 to t_2 and from t_3 to t_4.

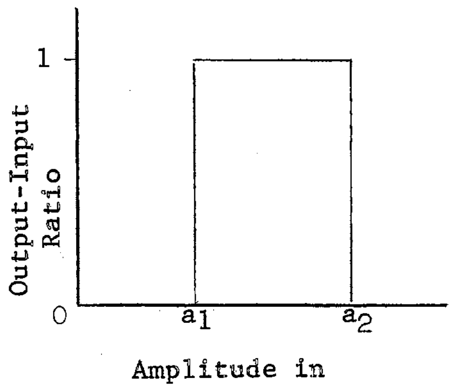

Figure 6

Ideal Output Characteristics
Amplitude Domain Filter

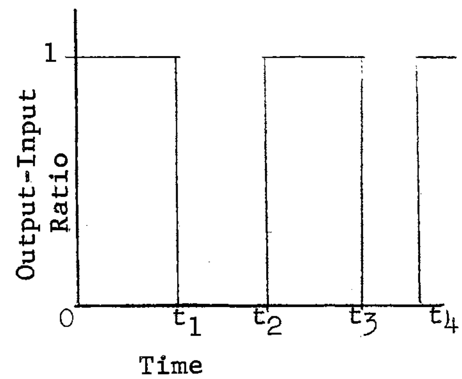

Figure 7

Ideal Output Characteristics
Time Domain Filter

Radio Noise Filters

Radio noise filters, more often called limiters, are of two general types. (20). One operates in the amplitude domain and is an amplitude limiter proper, limiting the maximum output voltage to a value not appreciably greater than some predetermined level. The second acts in the time domain and operates as a gate, which, in effect, punches a hole in the signal by curring the system off for the time that the noise exceeds some predetermined level. (20). In both cases, the best results are obtained when the action takes place early in the receiver circuit before the noise has had an opportunity to distort and lengthen in the circuitry. If placed in the first stage of the receiver, both types of limiters operating optimumly allow the same amplitude of noise to pass through the receiver. The limiter allows noise up to the maximum modulating peak-to-peak amplitude to pass, and the gate places a hole in the signal with the same peak-to-peak value. This is somewhat simplified, and arguments in favor of each have been presented. In either case, the impulse noise passed through the receiver is not of sufficient amplitude to overload the receiver or to be greatly modified by the receiver, and is of too low duration (too high a fundamental frequency) and amplitude to be bothersome. (21).

Our problem, however, is the noise which is passed to the audio amplification stages when the receiver either has no limiter circuit in the early stages or has an ineffective one. Radio noise filters which are designed to operate in and following the detector stage do fall within the scope of this thesis.

The filters most commonly used in and after the detector are the series-diode and shunt-diode limiters. A simple embodiment of the series-diode limiter is shown in Figure 8. The diode plate is connected to the output of the detector and the cathode is connected to a voltage divider which biases the cathode negatively. The detector provides a signal riding a negative d-c voltage. The voltage E is set by the potentiometer, and signals which drive the diode plate to a more negative voltage than E will not be passed by the diode, which then ceases to conduct. Many more sophisticated versions of this circuit are available. (22). The most commonly used is one in which the output from the detector is filtered by a relatively long time-constant circuit to provide the biasing voltage. It is then possible to set the limiting voltage at the value of the maximum carrier voltage and to obtain the maximum noise limitation possible without clipping the signal. In cases where the average carrier modulation is low and clipping may be tolerated, the limiting voltage is set lower with resultant high distortion on signal peaks.

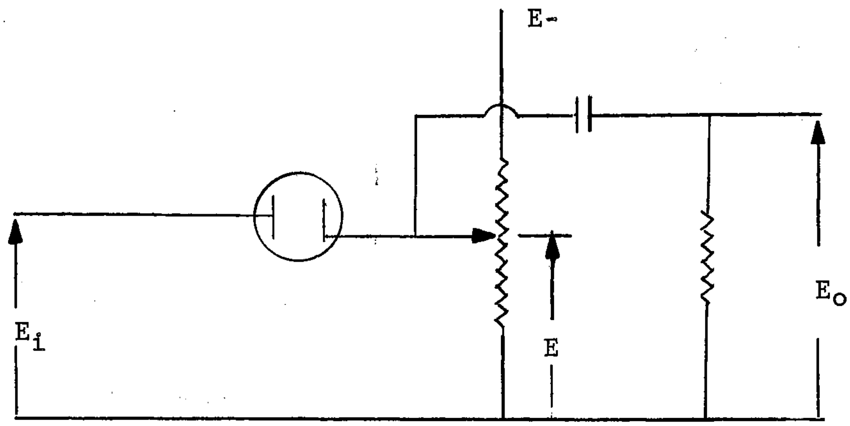

Figure 8

Series Diode Limiter

The shunt-diode limiter of Figure 9 operates in much the same manner, except that the diode is in shunt with the output and limits by conduction rather than by cutoff. The diode plate is biased at the limiting voltage. Any time that the negative value of the detector output exceeds the voltage E, the diode conducts, causing any signal voltage in excess of the bias voltage to be dropped across the series resistor R_1. The bias or limiting voltage, E, may also be derived from the carrier voltage for this circuit.

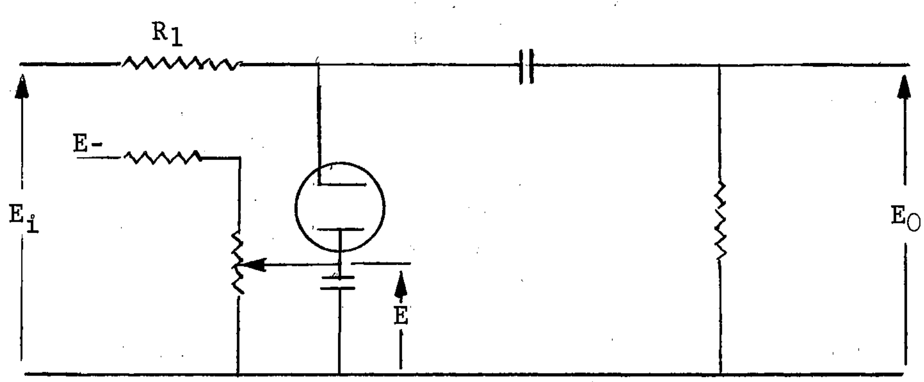

Figure 9

Shunt-Diode Limiter

There are triode variations of these circuits which decrease the loading effects upon the detector while at the same time providing gain, but the essential operation is the same and practically all limiters operate on these principles. There is one interesting variation of the shunt limiter which operates as a gate in the time domain rather than merely limiting the noise. This is the circuit of Figure 10 due to A. A. Gerlach. (19).

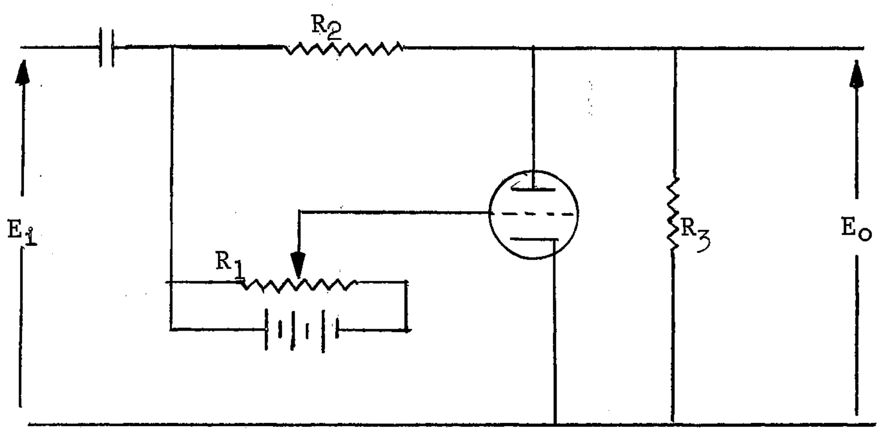

Figure 10

Shunt Triode Limiter

In this circuit, the resistances R_2 and R_3 are much larger than the resistance of the tube when the grid voltage is above cutoff. The grid is biased either as shown or by the AVC voltage so that the tube is curoff for voltages up to the maximum peak signal developed across R_3 of the voltage divider made up of R_2 and R_3. A positive pulse of noise exceeding the maximum signal drives the grid positive, and the tube resistance drops to a low value. For the duration of the pulse, practically the entire signal is dropped across R_1 and no output is developed across R_3. The original noise pulse is entirely eliminated and another pulse is introduced in the form of a hole in the signal. Thus, this circuit generates a pulse equal in amplitude to the instantaneous signal voltage at the time of the noise, while shunt and series limiters, if operated for maximum limitation with minimum distortion, allow a portion of the noise equal to the maximum peak-to-peak to pass. Using amplitude-probability functions and the standard deviation curve, Gerlach (19) has shown that the noise-power output from this circuit is 1/5 of that from a peak limiter.

None of these circuits will limit or suppress noise of a lower amplitude than the peak-signal amplitude without distorting the signal. They will considerably reduce high-amplitude impulse noise, but can do nothing to mitigate the effects of the noise on the operation of the previous stages. Furthermore, the impulse noise from a phonograph record is, more often than not, of lower amplitude than the loudest passages of the music recorded, and so cannot be gated by circuit of this type which are triggered by the over-all amplitude of the noise.

Audio Noise Suppressors

Only three types of noise suppressors have been specifically designed for operation in audio circuits. Oddly enough, each operates in its strict suppression action in a different domain (frequency, time or amplitude), and each is either fed from or controlled by a filter operating in a different domain than the suppressor. All three were originally designed for the suppression of phonograph noise only.

The first of these in a historical sense operates in the frequency domain and is the only one to enjoy any considerable commercial success, having been marketed as the Scott Dynural Noise Suppressor. (23). The primary purpose of this circuit is the suppression of background noise resulting from the motion of the pickup stylus across minor and almost continuous imperfections in the record material. This noise is concentrated in the higher audio frequencies and is of sufficiently low amplitude to be completely masked by music in the louder passages. At lower levels, analysis of the threshold of hearing over the audio-frequency spectrum and the probable distribution curves of music frequencies shows that the higher frequencies contained in the music are inaudible. (24). The record surface noise, however, is often far from inaudible. Therefore, the bandwidth of the playback equipment may then be restricted without reduction of the music but with considerable reduction of the noise. Since a considerable portion of the energy contained in a noise impulse is concentrated in the upper frequencies, this circuit is also to some degree effective on low-level impulsive noises which may also be masked by the louder passages. Therefore, the operation of a simple version of this circuit will be investigated.

This circuit, a simple version of which is shown in Figure 11, acts as a variable low-pass filter in the frequency domain, the cutoff frequency being determined by the signal amplitude integrated over a certain time. The circuit is an active four-terminal network consisting of a series and a shunt arm, each contributing infinite attenuation at a different frequency. The components of the series arm C_1 and L_1 are of fixed value and are designed to have a frequency of infinite attenuation higher than the highest useable audio frequency. The shunt arm consists of L_2 and the reactance tube V_1 with its associated circuitry.

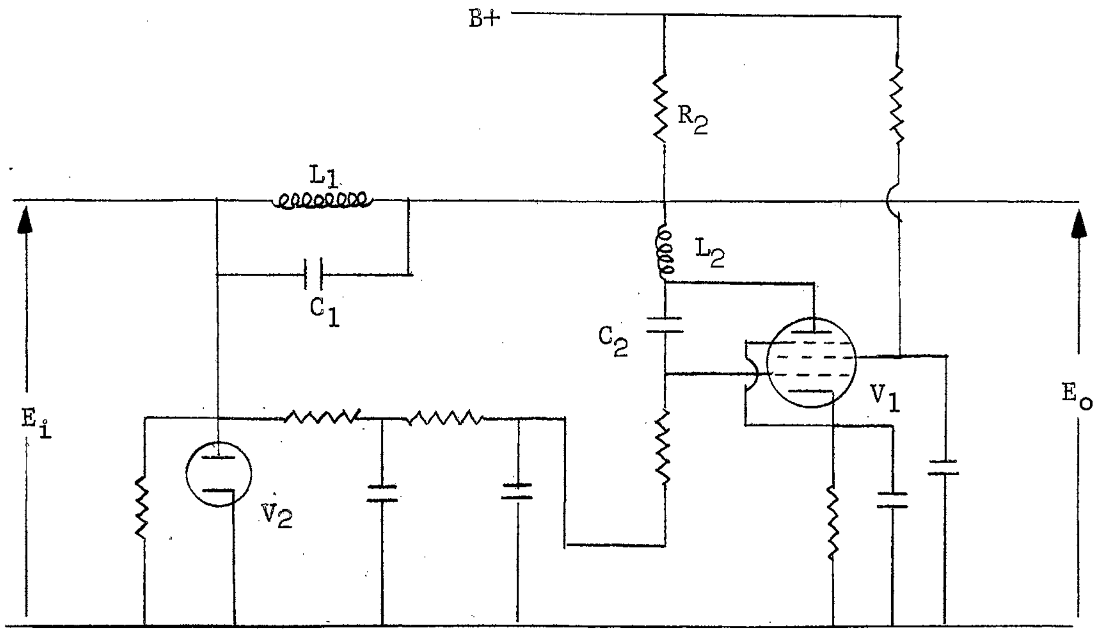

Figure 11

Scott Noise Suppressor, A Variable Low Pass Filter

The plate voltage of the reactance tube appears across C_2 and R_0 where R_0 is the equivalent resistance between the tube grid and ground. The grid voltage is that part of the alternating-plate voltage which is dropped across R_0. The voltage across this resistance leads the plate voltage, and the plate current therefore leads the plate voltage. Thus, the tube and associated circuitry act a a parallel combination of resistance and capacitance. It may be shown that the effective value of this capacitance, C_e, is (25).

$$C_e = \frac{C_2}{\omega^2 R_0{}^2 C_2{}^2 + 1} \quad (g_m R_0 + 1) \tag{4}$$

V_1 is a variable mu pentode, and the effective capacitance of V_1 as a reactance tube changes directly with its transconductance, as may been seen from Equation 4. The audio signal voltage is rectified by V_2 and filtered by the RC circuit following it. The time constant of this RC filter is adjusted so that the negative d-c voltage output is a function of slow-varying audio amplitudes but is insensitive to transients. This d-c voltage biases the reactance tube. As the audio voltage decreases, the transconductance of the tube increases and so, in accordance with Equation 4, the equivalent capacitance also increases. The resonant frequency of L_2 and the capacitance of the reactance tube drops, thus providing a lower cutoff frequency in the shunt arm of the filter network. Thus, the lower the signal level, the narrower the bandwidth of the circuit. For this reason the inventor calls this circuit a "frequency gating" circuit. (26).

It is obvious from the description of the circuit operation that only impulse noise of sufficiently low intensity to share the record surface-noise characteristic of being masked by high-intensity signals is suppressed in any part of its frequency spectrum. The lower-frequency components of the noise are unaffected, and high-amplitude pulses are attenuated only in their high-frequency components during low-level passages. Thus, this circuit and its many variations are of very limited effectiveness in the suppression of impulse noise.

A second noise-reduction circuit operates in the amplitude domain and discriminates against all low-level signals by the use of the nonlinear transfer-characteristic of two opposed diodes connected in parallel, as in Figure 12. (27). It may be seen from reference to Figure 13 that the nonlinearity of the circuit transfer-characteristic is such that low-amplitude signals and the low-amplitude portions of high-amplitude signals are greatly attenuated. The resultant crossover-distortion is eliminated to a large degree by using a group of these circuits in parallel, each being fed from one of a parallel group of one-octave bandpass filters. The center frequencies of these filters are separated by one octave so that the whole group passes the entire audio-frequency spectrum. Each diode circuit feeds a one-octave bandpass filter. The second filter tends to filter out the crossover distortion which consists of frequencies higher than the passband of the filter. The high-amplitude audio signal is thus reformed. This circuit is extremely effective in discriminating against noise of lower amplitude than some predetermined value, and is completely ineffective in discriminating against impulsive noise of high amplitude.

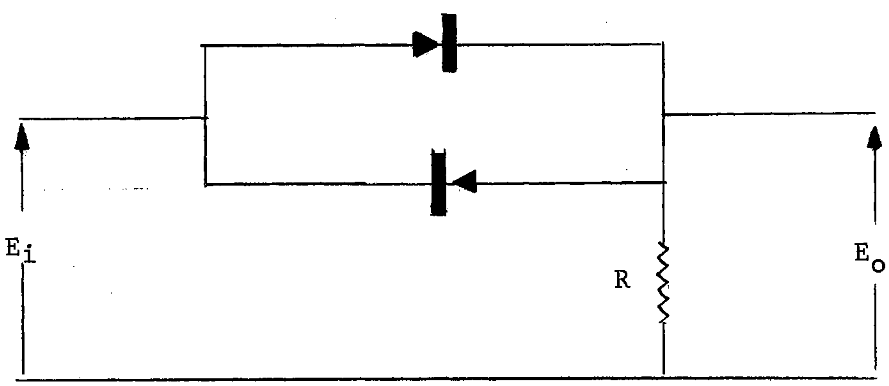

Figure 12

Low-Amplitude Rejection Circuit

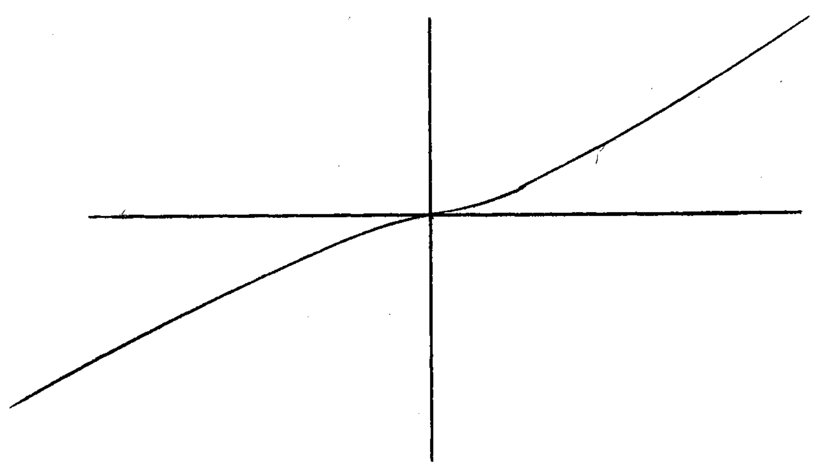

Figure 13

Amplitude Characteristics of Diodes of Fig. 12

CHAPTER IV

A FILTER IN THE TIME DOMAIN

The two filters (or "gates", as their inventor prefer to call them) presented at the end of the last chapter were not presented in order to prove that filters in the frequency and time domains are ineffective against impulsive noise, but merely to show that the existent audio noise suppressors of these types are ineffective. This, of course, is not unexpected, as these filters were designed with a view toward noise of a completely different aspect than impulsive noise. However, these filters are bound to be ineffective against impulse noise of approximately the same amplitude as the signal. It has been noted that filters in the frequency domain can discriminate against noise which is outside the frequency spectrum of the signal to be transmitted or against noise of narrow bandwidth within that spectrum. Neither criterion applies to impulse noise. Filters in the amplitude domain are effective against noise which is of greater or lesser amplitude than the signal, but they are obviously of little use when there is little or no difference between the signal and noise amplitudes. In many cases, limiting the impulse noise to the maximum signal amplitude may be sufficient. Impulse noise so limited may not impair the intelligibility of voice communications to a great extent, but it is irritating.

Time filters alone remain to be discussed. An extremely simple example of such a filter has been described and was shown in Figure 10.

Unfortunately, this filter requires for its action that a noise impulse be of greater amplitude than the signal, and it acts only as long as this condition prevails. It does not in any way lower the consequent distortion. Furthermore, the hole that it punches in the signal is itself a low-level impulse. While this impulse is of lower amplitude than an impulse resulting from a gate operating in the RF or IF stages of a radio tuner, the impulse is of unnecessary amplitude.

Any time filter or gate that is operated by the total impulse-amplitude for the period of the impulse only will have the characteristics of the circuit discussed above. Some other characteristics of impulse noise not possessed by the signal must be used to trigger the gate if improved operation is to be achieved. One such characteristic is the high-frequency component of an impulse. As has been noted, the signals obtained from records or from superheterodyne tuners contain no frequencies above 20 kc, while impulse noises do. The high-frequency components of a noise impulse may be used to trigger a gate which is sensitive to the amplitude of these frequency components. Since the duration of impulse noises is generally less than one millisecond and since the ear is insensitive to silences of this duration, a gating period of one millisecond should eliminate the noise with little, if any, aural effect.

Such a gating circuit has been patented by Mr. D. T. N. Williamson. (18). The patent gives only a block diagram of the circuit with two embodiments of the gating circuit itself. For reasons which are unknown to this author but which are probably economic in nature, the circuits covered by this patent have not been incorporated in any commercial unit, not has anything been published in reference to it. Mr. Williamson has stated that he knows of no other work being done or in progress in respect to this. (28). The circuit to be presented here is an embodiment of Mr. Williamson's block diagram with an improved lower noise and lower distortion gating circuit.

A block diagram of the noise suppressor is shown in Figure 14. It consists of delay and gate circuits in the main audio channel. In the control channel the signal is fed through a high-pass filter and amplifiers to the gate-control pulse generator. High-frequency impulses trigger this generator. The generator in turn operates the gate which momentarily blocks the signal from passage through the system.

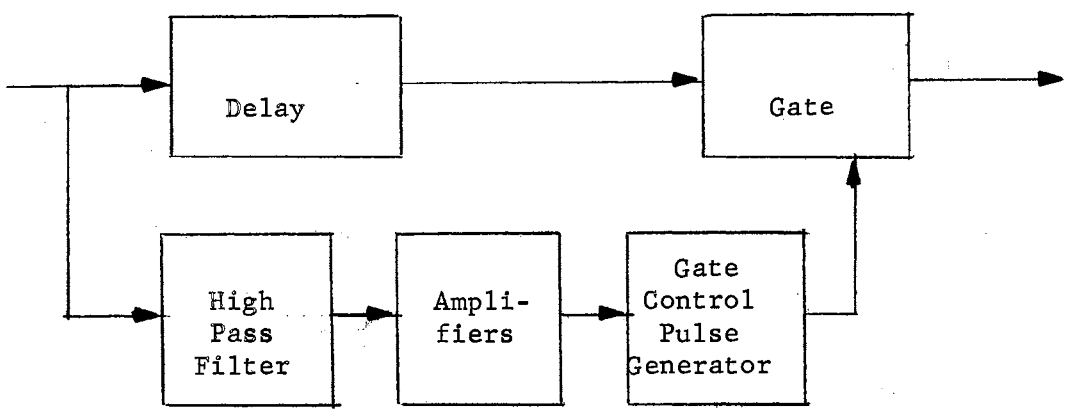

Figure 14

Block Diagram of an Impulse Noise Filter in the Time Domain

A gate circuit described in Mr. Williamson's patent is shown in Figure 15. The audio signal is introduced on the grid of V_1 and is amplified by that tube. The gate-control pulse generator provides a negative-going square wave at the suppressor grid of V_1 through C_1. This cuts the tube off and prevents the passage of the impulse through the circuit. Were this the entire extent of the gate, the positive gate pedestal, which is produced by the plate of V_1 rising to the supply voltage when the tube is cut off, would be greater than the original noise. Therefore, V_2 is added to the circuit and the output is taken from the tap on the potentiometer R_3 between the plates of V_1 and V_2. V_2 is held cutoff by a negative voltage on its suppressor grid. A positive gating pulse is introduced at this grid through C_2 simultaneously with the application of the negative pulse to the suppressor grid of V_1. When the plate voltage of V_1 rises, that of V_2 falls, and at some point on R_3, the d-c potential remains constant through the switching operation.

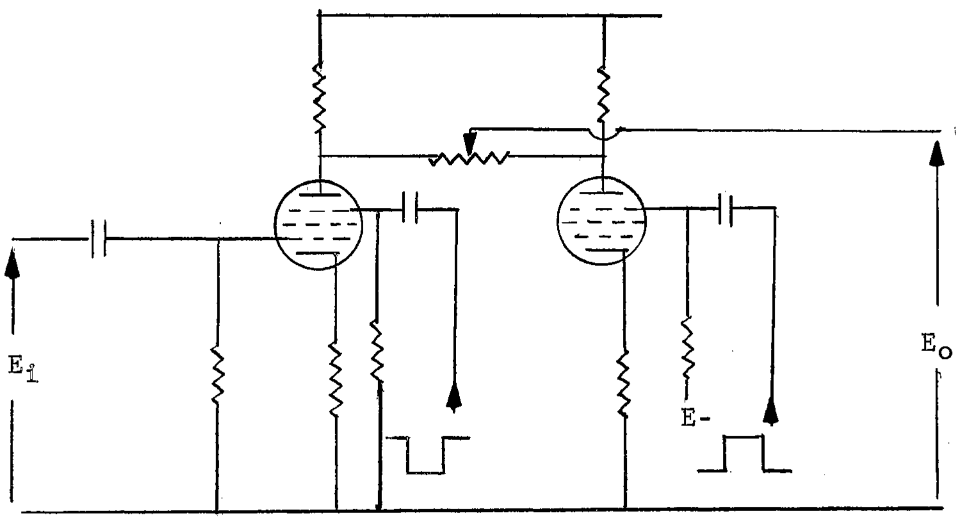

Figure 15

Pentode Antinoise Gate

Operation is relatively sensitive to changes in the values of the circuit parameters, and frequent adjustment of R_3 may be necessary. This is a minor fault. Of greater importance is the fact that a gate pedestal is produced even when R_3 is perfectly adjusted. As in the Gerlach circuit, the hole punched in the signal is equivalent to a noise pulse which is the negative of that portion of the signal which is removed by the gate. In case of a one-millisecond gating period, this pulse consists of a 1000 cps fundamental with a harmonic spectrum which is dependent upon the amplitude and waveform of the signal during the gating period.

It is possible to considerably reduce this pedestal. An extremely simple method is to use the diode circuit of Figure 16. In normal operation, positive portions of the signal are passed by the diode, D_2, and negative portions are passed by D_3. During the gating period, a negative square voltage-pulse is injected at the junction of R_{40} and D_2 and a positive square voltage-pulse is injected at the junction of R_{41} and D_3. These square waves are of greater amplitude than any audio signal encountered, and thus cut both diodes off for the gating period. R_{40} and R_{41} are of equal resistance and are large enough not to overload the gate control pulse generator. The time constant of these resistors and C_{31} is made sufficiently short to pass the highest audio frequencies desired. R_{42} must then be made large enough so that the time constant of it and C_{31} is long enough so that the voltage on C_{31} remains constant throughout the gating period. Thus, the pedestal produced is of much lower amplitude than was the case with the previously described circuit. This may be clearly seen from Figure 17 which shows comparisons between waveforms passing through the two gates and between the pedestals produced.

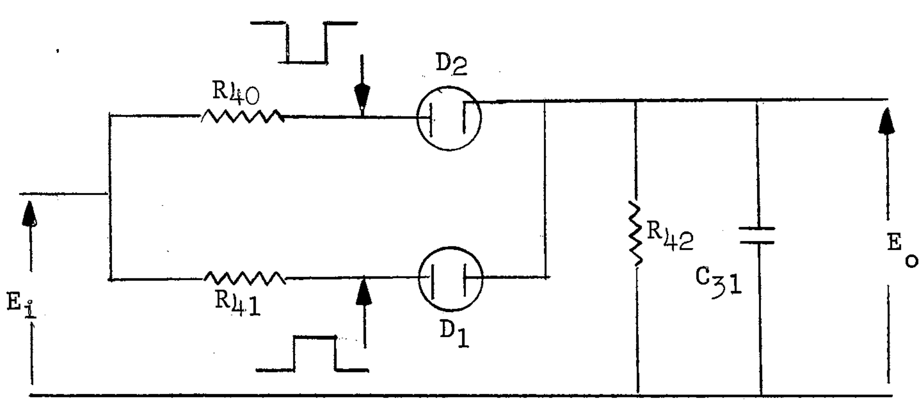

Figure 16

Diode Antinoise Gate

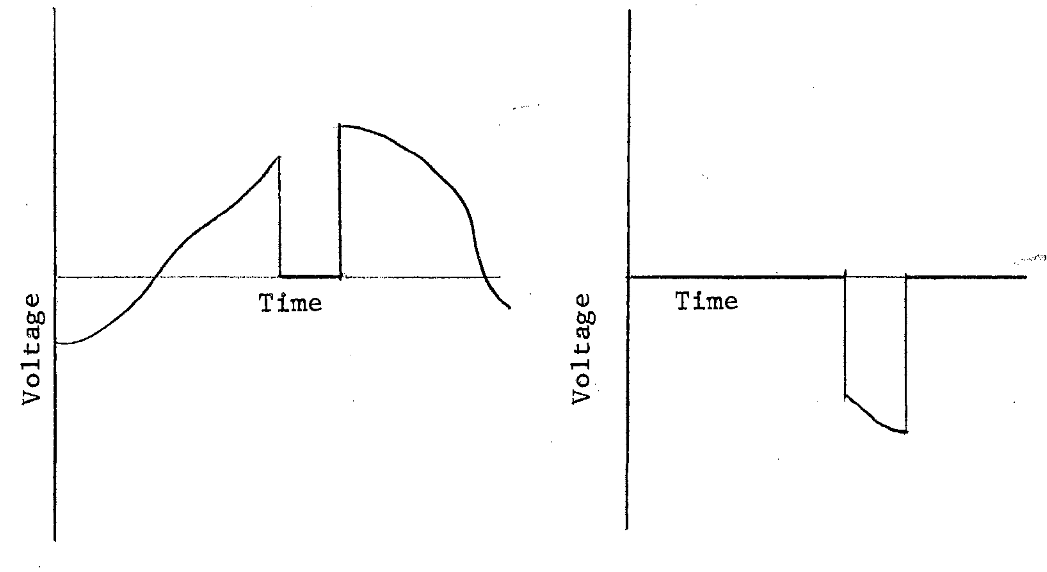

Figure 17A

Waveform Output and Gating Pedestal of
Pentode Antinoise Gate of Figure 15

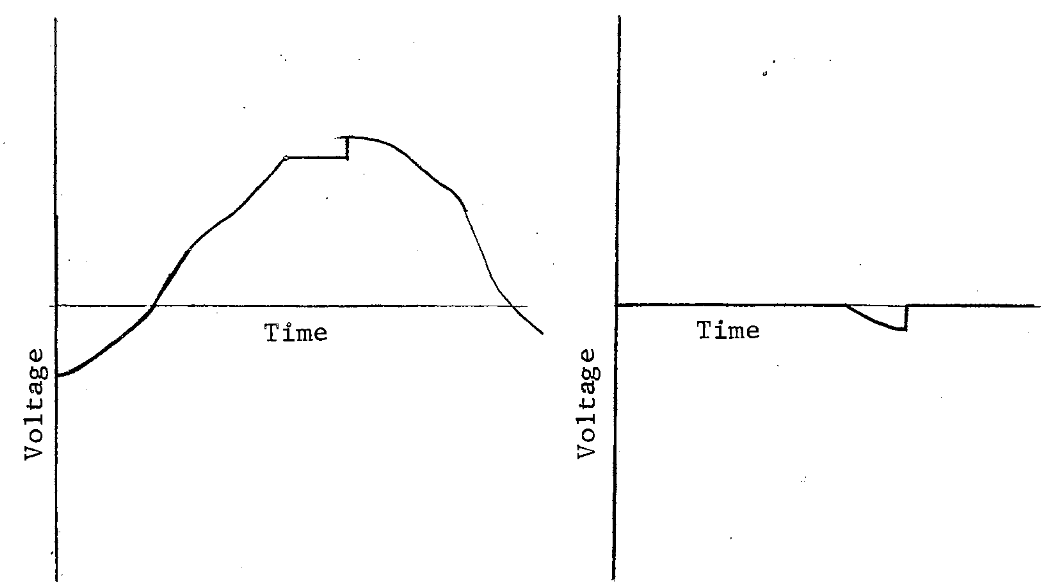

Figure 17B

Waveform Output and Gating Pedestal of
Diode Antinoise Gate of Figure 16

This embodiment of the gating circuit also has a drawback. The diodes, when conducting, are operating on the lowest part of their transfer curves. This is the area of maximum nonlinearity, and the output is somewhat distorted. This distortion may be quite small if the series resistors R_{40} and R_{41} are of much greater resistance than the diode forward-resistance, and if the applied signal voltage is large. The design considerations bearing on the relative values of R_{40}, R_{41}, R_{42} and C_{32} require that the resistance of R_{40} and R_{41} be relatively low.

The same principle of diode operation may be used and distortion avoided by forward biasing the diodes as in Figure 18. This is the circuit used in the final suppressor design. The design requirements concerning the relative values of R_{40}, R_{41}, R_{42} and C_{32} are the same as they were for the unbiased diode circuit discussed above.

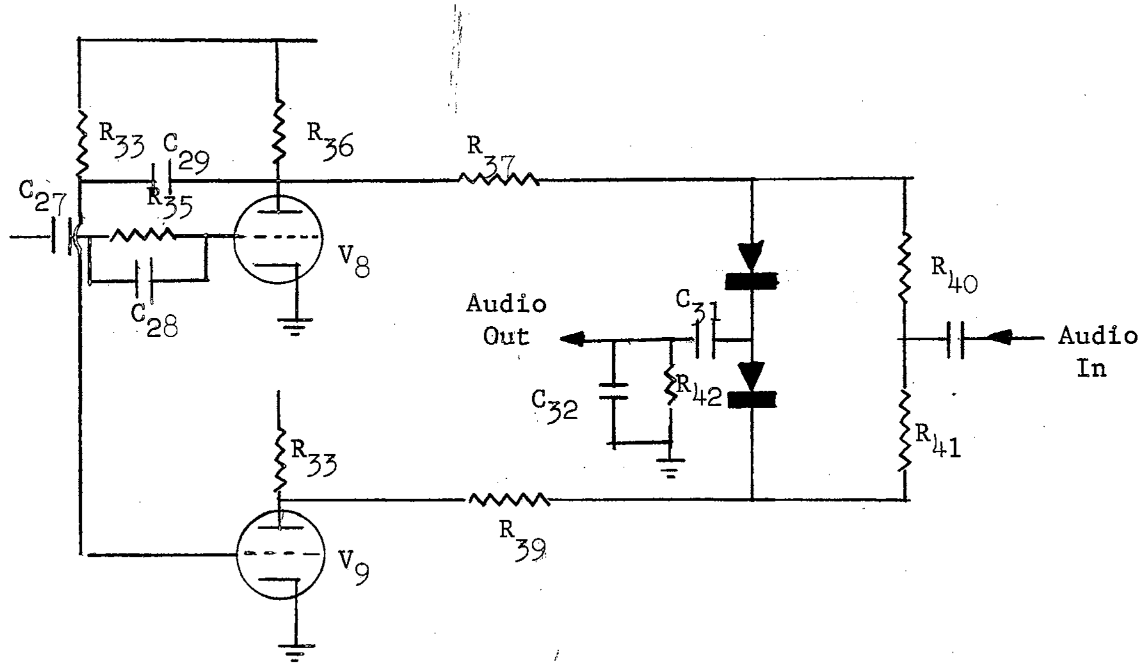

Figure 18

Forward Biased Diode Antinoise Gate

Bias is provided by placing the gate circuit between the plates of the two triodes V_8 and V_9. These tubes are operated as overdriven amplifiers. V_8 is normally biased far below cutoff by applying a negative voltage to its grid through the large resistor R_{34}. The grid of V_9 is connected to the supply voltage through the very large resistor R_{33}. R_{33} is so much larger than the cathode-to-grid resistance of the tube that practically all the voltage is dropped across R_{33}, and the grid of V_9 is normally very slightly positive. With V_8 cut off, the total series resistance of R_{36}, R_{37} and R_{39} in parallel with R_{33} forms the plate load on V_9. Part of the plate current of V_9 therefore flows as bias current through the diodes.

R_{37} and R_{39} are of equal resistance, as are R_{33} and R_{36}. V_8 is driven by a large positive-going square wave. The grid-limiting resistor R_{35} allows the grid of V_8 to go only slightly positive. The negative square wave output from the plate of V_8 is coupled through C_{29} to the grid of V_9, and cuts that tube off for the gating period. Thus, the voltages that normally exist at the plates of V_8 and V_9 are interchanged for the duration of the gating period. The diodes are reversed biased and cut off for this period.

R_{40} and R_{41} are small in relation to R_{33}, R_{36}, R_{37} and R_{39}, but are large enough so that the reverse bias applied to D_2 and D_3 is greater than any audio signal fed to the gate. The audio signal could be fed in through just one of the resistors R_{40} and R_{41}, but if it were, the entire gating pulse would be fed back into the preceding audio circuitry. The transient at the trailing edge of the gate pulse would then re-trigger the gate-control pulse generator. The result would be oscillation. With both resistors in the circuit, the opposing square waves produced in the gate cancel one another at the junction of the resistors, and the d-c potential remains constant at this point.

The gate-control pulse generator is the d-c coupled gate of Figure 19. This is a monostable version of the familiar plate-coupled multivibrator. V_6 is normally biased beyond cutoff by supplying its grid with a negative voltage from the voltage divider formed by R_{27}, R_{29} and R_{31}. V_7 is normally conducting heavily due to the positive voltage supplied to its grid through R_{30}. A positive pulse at the grid of V_6 causes it to conduct. Its plate voltage drops and this negatively going pulse is applied to the grid of V_7 through C_{26}. The plate voltage of V_7 rises and, as it is coupled to the grid of V_6, this rising voltage augments the positive trigger to V_6. This action continues until V_6 is conducting heavily and V_7 is cut off. When C_{26} has charges through R_{30} to the cutoff voltage of V_7, V_7 once more conducts. Its plate and the grid of V_6 go negative and the plate of V_6 goes positive. This positive-going voltage is fed through C_{26} to the grid of V_7. This action progresses cumulatively around the loop until the stable condition is reached.

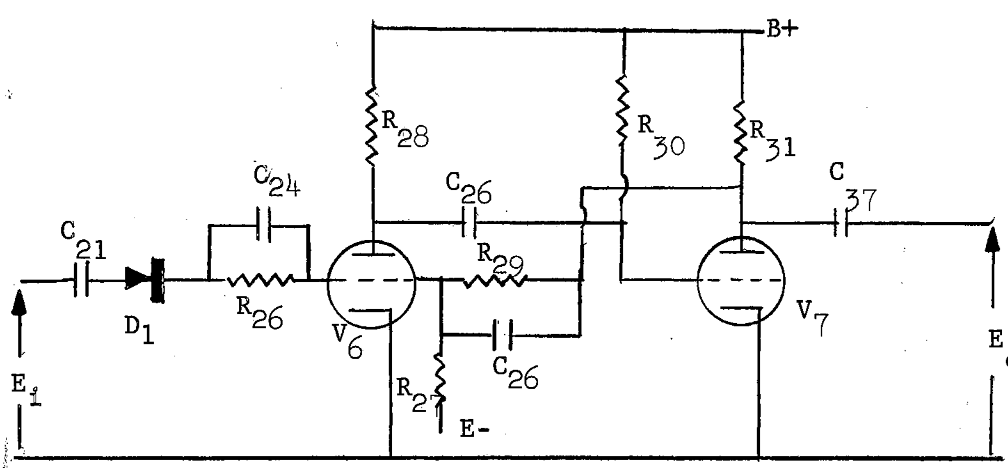

Figure 19

D-C Coupled Gate

The capacitor C_{25}, which might appear to be unnecessary at first glance, serves to decrease the switching time. The interelectrode capacitances of V_6 form a total capacitance in parallel with R_{27}. This capacitance tends to hold the voltage across R_{27} (and therefore, the grid voltage of V_6), constant as the plate voltage of V_7 changes. If the ratio of C_{25} to this interelectrode capacitance is made the same as the ratio of R_{27} to R_{29}, the change of the plate voltage of V_7 is immediately felt on the grid of V_6. The diode D_1 allows only positive triggers to pass to the grid of V_6 so that any negatively going waveform following a positive trigger cannot interrupt the multivibrator action. R_{26} is a grid current-limiting resistor and prevents grid current from charging C_{21}. C_{24} speeds the triggering of V_6 in much the same way as C_{25}.

The time duration of the pulse output from the circuit is given by

$$T = C_{26}R_{30}\log_e \frac{2\,E_{bb} - E_{b6}}{E_{bb} - E_{o7}} \qquad (5)$$

where E_{bb} is the supply voltage

E_{b6} is the plate voltage of V_6 when that tube is conducting

E_{o7} is the cutoff voltage of V_7 (10).

It may be seen that if the various voltages remain constant, the pulse width is controlled by $C_{26}R_{30}$. The gating period may therefore be set for the minimum time consistant with the type of noise to be gated by making R_{30} variable.

It might seem that the gate circuit may well have been placed between the plates of the multivibrator tubes without adding the two overdriven amplifiers to the circuit. There are several reasons for not doing this. The gate circuit requires fast, simultaneous, equal and opposite square waves. The outputs from the plates of the multivibrator are not exactly equal, nor are they simultaneous. C_{26} loads V_6, and causes some rounding of the leading edge of its output pulse. When the circuit switches back to the stable condition, C_{26} charges in the opposite direction through the cathode-to-grid resistance of V_7. This causes a negative pip to appear at the trailing edge of the output pulse of V_7. At the leading edge of the pulse, V_7 is being driven by and follows V_6, and at the trailing edge the opposite condition prevails. Thus, the output pulse from V_7 is very slightly shorter than the output pulse from V_6.

The overdriven amplifiers need only one-seventh of the multivibrator output to be driven from cutoff to zero bias, and so their switching time is much less. Furthermore, if one of the overdriven amplifiers acts more quickly than the other, its action may be slowed by capacitively bypassing its plate-load resistor.

In order that the noise-suppressor may operate correctly when placed in an audio amplifier with inputs from various sources, it is necessary that the high-pass filter have a cutoff frequency higher than the highest signal frequency obtained from any of the inputs. As has been noted, AM broadcast stations seldom use modulating frequencies in excess of 7 kc. However, there is often a 10 kc output from an AM tuner which results from crossmodulation of the carrier frequencies of transmitters adjacent to one another in the broadcast-frequency spectrum.

Spectrum analysis of recordings of various musical selections shows that the energy in the one-third octave centered at 12 kc may occasionally be as great as that in any one-third octave centered at any lower frequency, and that appreciable energy may be present up to 17 kc. (29).

It seems, therefore, that signal frequencies in excess of 20 kc will only very rarely be encountered, so 20 kc was selected as the filter cutoff frequency. Attenuation below the cutoff frequency should be as rapid as possible in order that the signal may not trigger the gate. The requisite rate and degree of attenuation can be achieved by the use of an LC filter. While the design of such a filter presents no difficult problems, the construction of high Q inductors to the design specifications and acquisition of capacitors within close tolerance of the design values is difficult and time-consuming. Were this circuit to be produced in numbers, it would be profitable to overcome these difficulties. For a single trial circuit, it was felt to be more efficient to build an active RC filter using feedback to achieve the desired rate and degree of attenuation, while at the same time giving the amplification required to trigger the gate-control pulse generator.

Any desired rate of attenuation may be obtained by using a sufficiently long array or ladder of simple RC sections, provided that each section is of much higher impedence than its predecessor, or that the sections are isolated so not to load one another. Unfortunately, the ultimate rate of attenuation is approached slowly, and there is a three db loss at the turnover frequency for each 6 db per octave of ultimate attenuation. A greater rate of attenuation involving the use of fewer components is possible using the parallel T RC network of Figure 20. With the relative values shown, this circuit gives infinite attenuation at $\omega_0 = 1/RC$ and 6 db attenuation at $0.4\omega_0$ and $\omega_0/0.4$. (30).

If several simple RC sections, each having a turnover frequency of 18 kc, are cascaded with two parallel T networks tuned to 17 kc and 12 kc respectively, the RC sections give a high attenuation at low frequencies, and the parallel T network provides the required rapid attenuation just below 20 kc. There is a rather slow rolloff with the considerable loss of 26 db at 20 kc with such a circuit. This may be eliminated by the use of some circuit with a peaked response at 20 kc. Any of several RC oscillator circuits operating just short of oscillation will provide such a response. The phase-shift oscillator of Figure 21 is such a circuit. It may be shown that this circuit will oscillate at the angular frequency $\omega_0 = \sqrt{6}/RC$, provided that the amplifier gain is -29 and that the amplifier-plate resistance is of a much lower value than the individual resistors of the phase-shift network. (25).

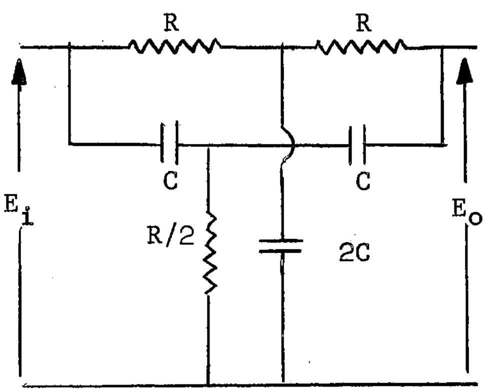

Figure 20

Parallel T Filter

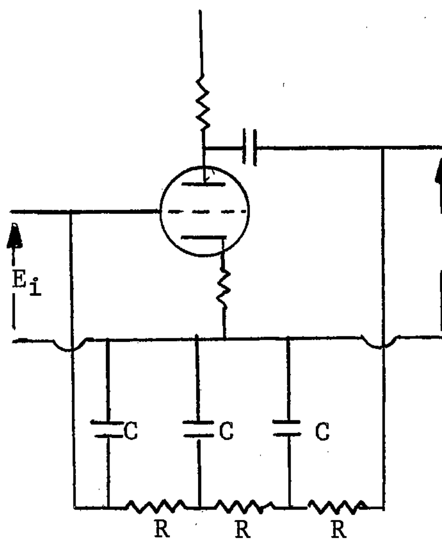

Figure 21

Phase Shift Oscillator

The phase-shifting network shown uses series resistors and shunt capacitors rather than the reverse combination usually employed. This configuration has two advantages. First, the circuit usually employed oscillates at $\omega_0 = 1/\sqrt{6}RC$, while this configuration oscillates at $\omega_0 = \sqrt{6}/RC$. Therefore, for a given frequency, a value of RC six times as large may be used. This allows the load resistor of the tube to be small in relation to R yet large enough so that the tube gain is high. At the same time, C may be large enough so that interwiring capacitances are relatively small and may safely be ignored. Secondly, there is little feedback at high frequencies, as the shunt capacitors ground the feedback signal. There is negative feedback at low frequencies. This helps to achieve the desired filter-transfer characteristic of high relative gain above 20 kc.

The circuit ceases oscillation when the loop gain around the loop from the tube grid through the tube and feedback network back to the grid is reduced below one at the frequency at which there is 180° phase-shift around the loop. The circuit then acts as an amplifier with a response that is peaked at ω_0. This is done by taking the feedback from a voltage divider which is connected across the output of the phase shifting feedback network.

Due to the loading effects of one circuit upon another, these several RC circuits cannot be merely cascaded in any order and the resultant circuit then be expected to have a transfer function equal to the product of their individual transfer functions. For example, the parallel T network presents a load of R/2 at resonance and, as the frequency increases above resonance, the input impedance decreases approaching R/4. Loading the T decreases transmission on both sides of resonance, though the decrease will be less than one db if the load exceeds 6R. (30). Thus, the equivalent Q of the parallel T network is decreased unless it is driven from a low-impedance source and loaded by a high impedance.

The gate-control multivibrator requires a trigger of approximately ten volts to insure positive triggering. Using a damaged record and a phonograph pickup with a peak output of approximately 0.15 volts on high intensity passages, it was found that a circuit-gain slightly in excess of 80 db at the frequencies above 20 kc was necessary in order to secure this triggering voltage. Considering this along with the impedance-matching problems, the circuitry of the first five tubes of Figure 22 was arrived at for the active filter and amplifier. This figure shows the final design of the noise-suppression circuit.

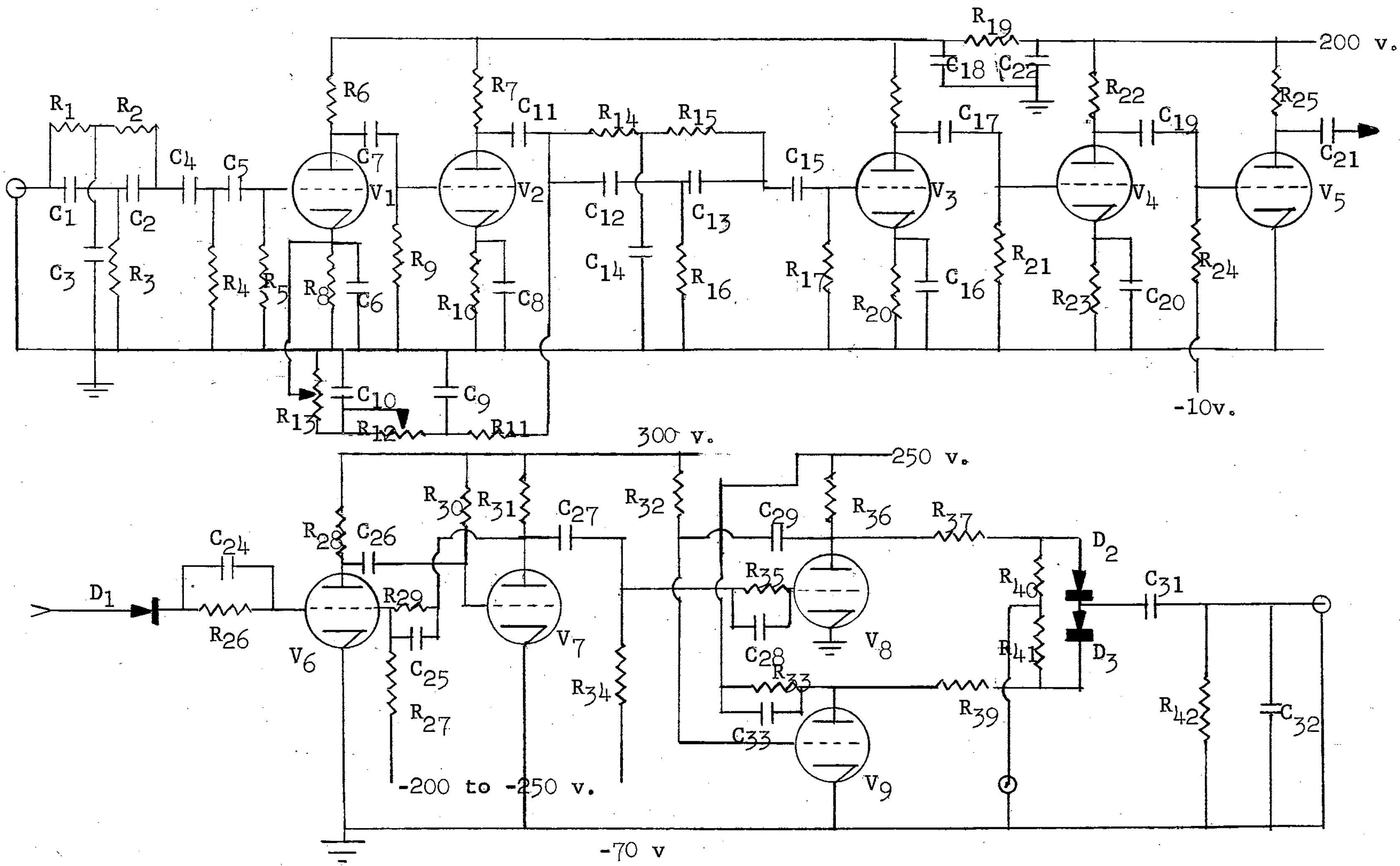

Figure 22

Circuit Diagram of Noise Suppressor

TABLE I

COMPONENT VALUES FOR FIGURE 22

Resistors (All resistors $\frac{1}{2}$ w, 10% unless otherwise specified.)

R1, 2	100K ohm 1%	R22	68K ohm
R3, 14, 15	50K ohm 1%	R23	820K ohm
R4	47K ohm	R25	56K ohm
R5	90K ohm	R26	180K ohm
R6, 11, 18, 21, 24	100K ohm	R27, 34, 35	1M ohm
R7	220K ohm	R28, 31	10K ohm, 2 w.
R8	5.6K ohm	R29	690K ohm
R9, 17	470K ohm	R30	18M ohm
R10	2.2K ohm	R32	8M ohm
R12, 13	100K ohm pot.	R33, 36	25K ohm, 2 w.
R16	25K ohm	R37, 39	33K ohm, 1%
R19	50K ohm	R42	2M ohm
R20, 40, 41	1K ohm		

Capacitors (All capacitors 600 v, 20% unless otherwise specified.)

C1, 2	100μμfd 5% mica		C18	10μfd	electrolytic
C3	200μμfd 5% mica		C20, 21	.0047μfd	
C4	330μμfd 5% mica		C22	16μfd	electrolytic
C5	.001μfd		C24	56μμfd	mica
C6	2590μfd	mica	C25	47μμfd	mica
C7, 8, 11, 15, 16	.01μfd		C26	13μμfd	mica
C9, 10, 28	470μμfd		C27, 31	.25μfd	
C12, 13	130μμfd 5% mica		C29	.5μfd	
C14	250μμfd 5% mica		C32	3600μμfd	
C17, 19	75μμfd 5% mica		C33	110μμfd	

Germanium Diode

D1

Tubes

D2, 3	6AL5
V1, 2	12AX7
V3, 4, 5, 8, 9	12AT7
V6, 7	12AU7

CHAPTER V

RESULTS AND RECOMMENDATIONS

Any time that an impulse noise occurs, the input filter and amplifiers are required to deliver to the multivibrator a trigger at least ten volts in amplitude at a frequency exceeding 20 kc. At the same time, the signal to be transmitted should never be allowed to reach the multivibrator at this amplitude. It was assumed that the peak signal from any source that would feed the circuit would be no less than 0.15 volts. Any signal of higher amplitude may be attenuated to this level before being fed to the circuit. A gain of 30 db is therefore allowable for signal frequencies without their reaching a triggering amplitude. It was found that an additional gain of 40 db at frequencies exceeding 20 kc would insure the necessary trigger to the multivibrator.

The frequency-response curve of Figure 23 is that of the active filter and amplifier. It may be noted that the relative low frequency attenuation required is obtained between 22 kc and 17.5 kc. The rate of attenuation between these frequencies approximates 150 db per octave. The response at 23 kc could have been increased by 10 db by tuning the phase-shift oscillator circuit closer to self-oscillation. However, such an increase in the Q of this circuit only produces a sharp peak at 23 kc and decreases the gain at higher frequencies. Furthermore, this brings the circuit so close to self-oscillation that every time the circuit is shock-excited it produces a long-decaying oscillitory wave-

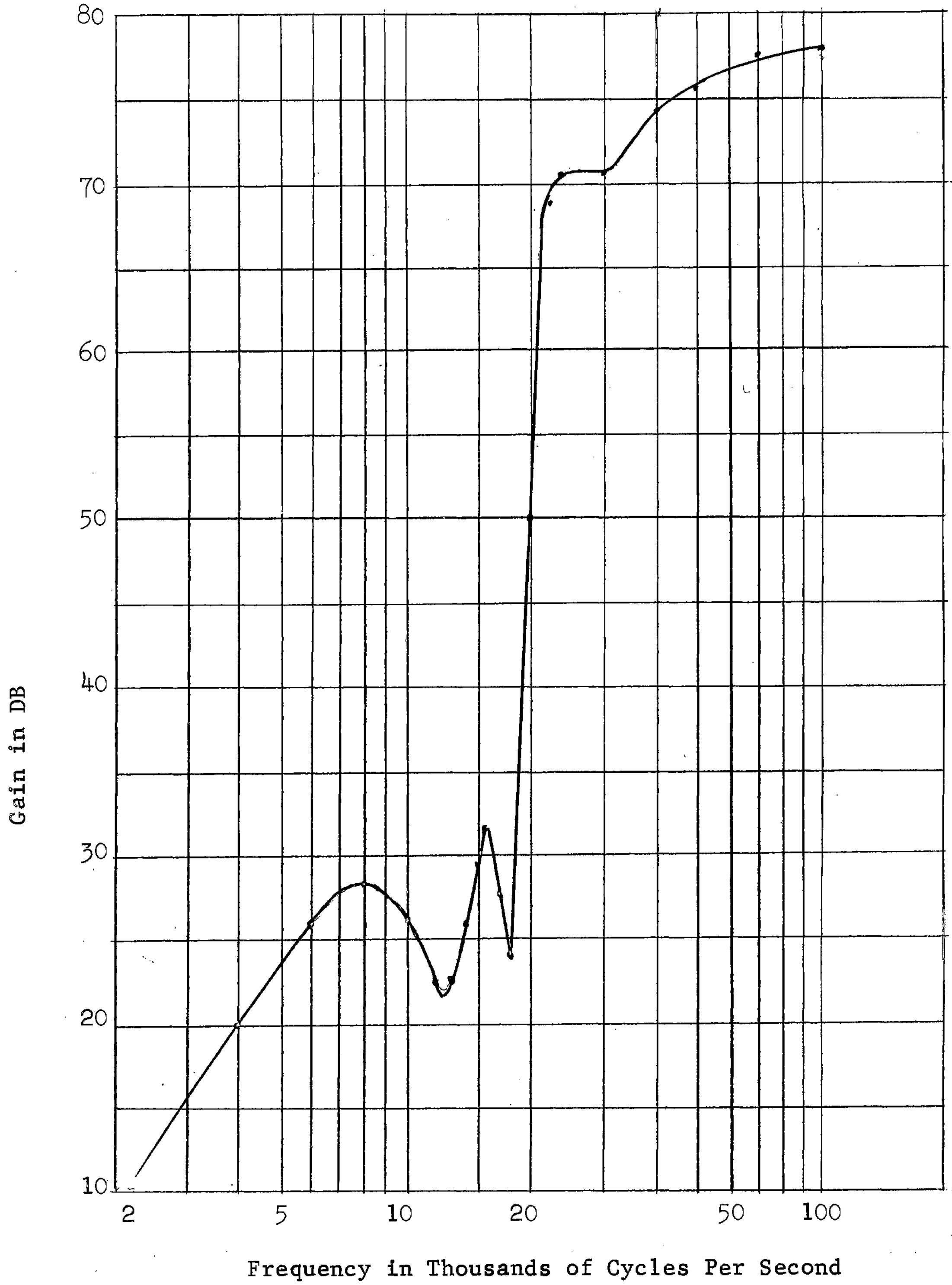

Figure 23

Frequency Response of Suppressor Input Filter form, and that the small background noise from the record surfaces and circuit components keeps triggering the circuit into low-amplitude oscillation.

Certain difficulties were encountered with this circuit. It was found that when a single tube was used and the plate-load resistance and phase-shift network impedance were made large enough so that the tube gain approached -29, the impedance of the driving source was low enough to overload the phase shift network. Therefore, insufficient voltage was fed back to obtain the required response from the circuit. Two tubes were threfore used with the feedback being taken to the cathode of the first tube. The cathode resistor of this tube was bypassed by a small capacitance, as were all the cathode resistors, to increase the gain above 20 kc. This capacitance and the coupling, interelectrode and load capacitances introduced sufficient phase-shift into the circuit, so that only two RC combinations were required in the phase-shift network instead of the three normally used.

There is a 0.5 percent change in the null frequency of a parallel T network for each 1.0 percent change in a component value. (30). Precision capacitors were not used in these networks. Further detuning was caused by the partially capacitive loads placed upon them. Therefore, the parallel T null points fall at 12.5 kc and 17.6 kc respectively rather than at the design frequencies of 12 kc and 17 kc.

It was noted in reference to the gate design that simultaneous, equal and opposite square waves are required for the proper operation of the gate. If one of the square waves arrives at the gate earlier than the other, the capacitor C_{32} charges in the direction of the earlier square wave until the other square wave arrives. C_{32} holds that charge for the duration of the gating period. The voltage output due to this is, in effect, a gating pedestal which may be of great amplitude. This problem was easily solved by slowing the operation of the amplifier which provides the earlier square wave. V_8 drives V_9, and so V_9, which is overdriven by only a small part of V_8's output, provides the fastest-rising square wave. Therefore, the plate of V_9 was bypassed by the capacitor C_{33} in order that its switching action might be slowed. Since the plate voltage of V_8 starts to drop before that of V_9 starts to rise, there is a slight drop in the voltage of the gate output at the start of the gating period. This counter acted when the plate waveform of V_9 catches up with that of V_8.

If one of the square waves ends or even starts to drop before the other, there will be a high-amplitude pulse at the output of the gate. This pulse is somewhat longer in duration than the time difference between ending of the two square waves, due to the discharge time of C_{32}. No way was found to time the square waves so as to eliminate this pulse. The problem was made more acute by the fact that the plate-bypass capacitor C_{33} used to slow the rise time of the positive-square waveform from the plate of V_9 also acts to delay the end of this waveform. The final circuit produces a pulse of 2.5 volts in amplitude and 50 microseconds in duration.

Figure 24 is a representation of the output of the gate when triggered with no signal passing through it. The two impulses mentioned above may be seen in addition to a third departure from a noiseless output. This is a 0.2 volt dip following the gating period. This negative voltage pip decays logarithimically with a time constant of 15 milliseconds. The exact origin of this is unknown, though it is evidently due to the discharge of some capacitance which charged during the gating period. Every effort was made to eliminate this dip. It was greatly accentuated when the output of V6 of the multivibrator was used to drive V_9. Due to the relatively slow switching of the multivibrator, all the gate noises were increased when this was done. After several minor circuit variations were attempted, limiting diodes were introduced at the plates of V_8 and V_9. These were connected to regulated voltages of 150 volts and 255 volts obtained from the plates of a series of two VR tubes. Unfortunately, it was discovered that VR tubes are relatively show-acting and are entirely unsuitable for filtering short-duration, fast changes of current. An attempt to filter the high-frequency changes of current occuring during the switching action by placing a capacitor across the VR tubes merely resulted in the VR tubes oscillating on and off.

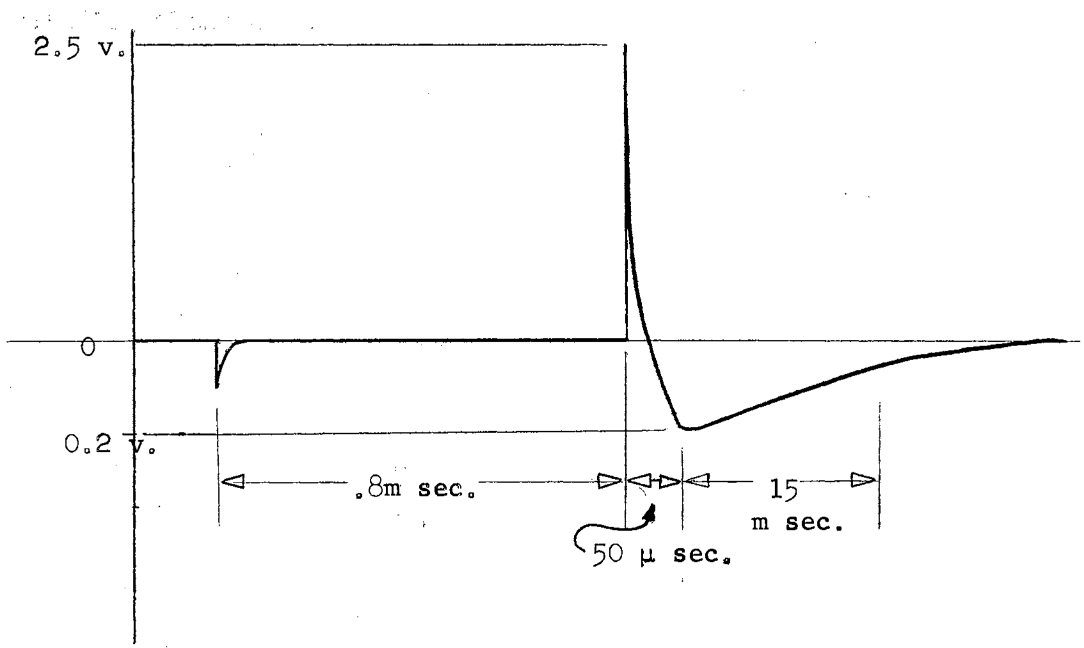

Figure 24

Noise Voltage Output Produced by the Gate

Excluding these three exceptions, the gate operates as it is supposed to. Impulse noise with the relatively low fundamental frequency of 2,000 cps is gated out. The three noises produced by the gate evidence themselves as an audio sound of lower amplitude and of higher frequency than the gated noise.

Intermodulation-distortion measurements using a Heath Audio Analyzer showed no distortion above the residual readings of the meter for any input voltage from 0.3 volts to 10 volts. Of course, when the gate operates, there is a considerable amount of distortion introduced during the gating period, but this is impulsive and unmeasureable. It is of short duration and preferable to the noise which is eliminated.

While the circuit as described and constructed does not operate perfectly, it does fulfil its intent. It shows that impulse noise may be time filtered and that the gating pedestal produced can be of considerably lower amplitude than the noise it replaces, even though that noise is of no greater amplitude than the signal.

Certain improvements are possible which were not incorporated in this circuit.

(1) Faster and more synchronous square waves at the gate would improve operation. This may require the use of diode limiters at the plates of the gate drive tubes, V_8 and V_9. These diodes would have to be biased from a regulated power supply which would provide effective regulation with rapid changes of load.

(2) The circuit may be considerably simplified by the use of an LC filter and pentode amplifiers.

(3) The triggering output of the amplifier which drives the multivibrator is a waveform composed of several oscillations. These may start with

either a positive or negative polarity, depending upon the direction from which the original noise starts. Gate operation on some impulses would, therefore, be faster if some method of triggering the multivibra- were incorporated which would trigger the multivibrator on negative-going as well as positive-going pulses.

(4) Some further reduction of the noise due to the gating pedestal may be achieved by dividing the signal channel into high and low-frequency channels before the gate. The low-frequency channel would carry the frequencies below the lowest fundamental frequency of the noise, and the remainder of the signal frequency spectrum would be transmitted by the high-frequency channel. Only the high-frequency channel need then be gated, and there would be no gating pedestal as such.

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