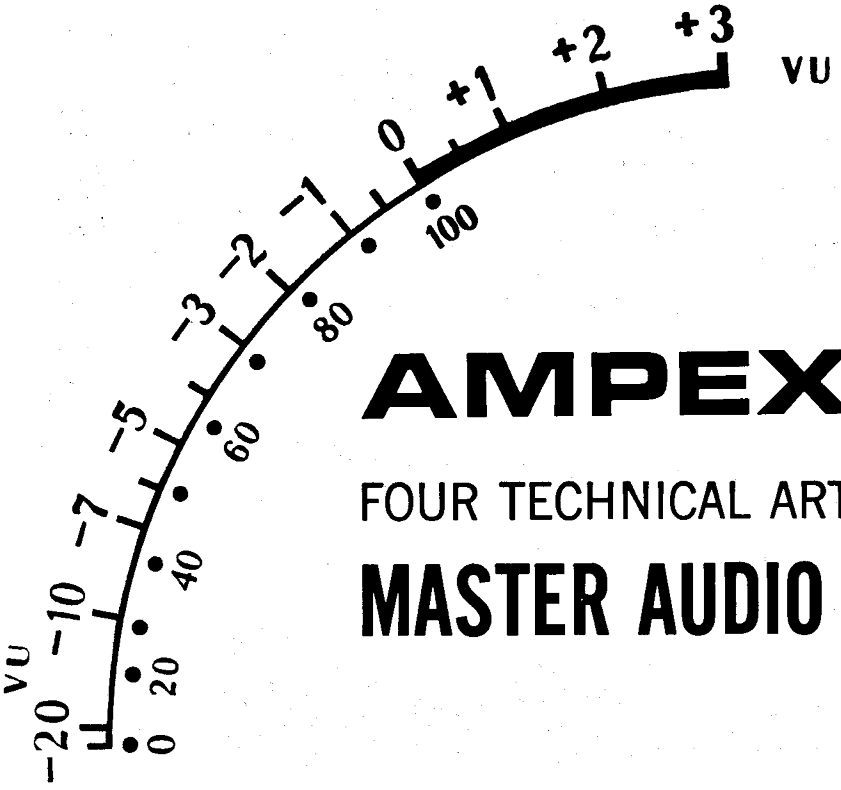
To: Tom Fine
with my compliments
J Mc Knight
2002-06-18



AMPEX

FOUR TECHNICAL ARTICLES ON
MASTER AUDIO TAPE RECORDERS

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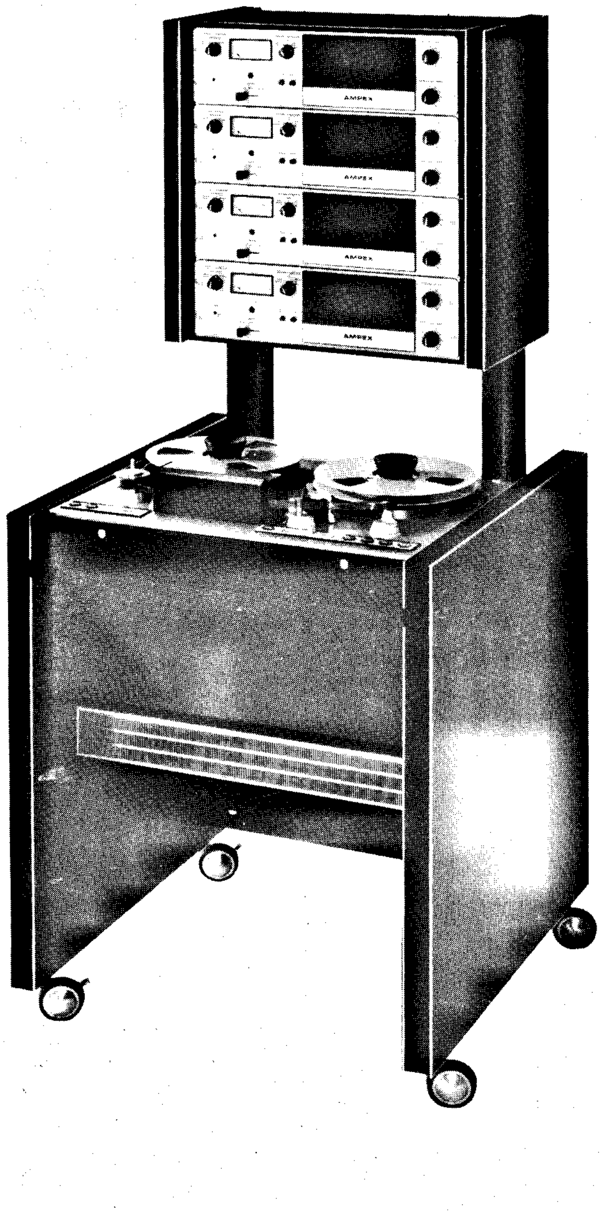

Performance and Reliability Requirements for a Master Tape Recorder. October 1964

Noise Limitations in Tape Recorders. October 1964

Mechanical Damping in Tape Transports. April 1964

Dynamic Range Limitations in Tape Recording. October 1964

Performance and Reliability Requirements for a Master Tape Recorder[*]

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In view of the maximum performance and reliability demands of the record industry for tape recorders, the current state of the art is evaluated, and the performance of a new tape recorder is compared to the industry's requirements and to existing tape recorders.

INTRODUCTION

When the design program for a new master recorder was initiated at Ampex Audio during 1960, discussions with key individuals in the master recording industry revealed that common requirements existed for a master recorder, even though each person interviewed had a personal order of priority for those requirements. Two general demands of the industry were particularly evident: *1.* An improved signal-to-noise ratio, and *2.* Higher reliability and better maintainability.

Other requests (generally of slightly less importance) included those relating to *1. Electronics: a*) Less distortion, especially in recording electronics, *b*) More stable gain and frequency response, *c*) Adjustments possible from the front panel, *d*) Modular design, *e*) Compensation for minor variations in line voltage, *f*) Reduction of D-C (AM) noise; *2. Head Assembly: a*) Reduction of frequency modulation noise (scrape flutter), *b*) Full accessibility to heads through the head-gate, *c*) Power-operated tape lifters, *d*) Greater erase efficiency; *3. Tape Transport: a*) Better stability of azimuth alignment, *b*) Faster starts to stable tape motion, *c*) Reduced speed variations, both average speed and flutter, *d*) Better transport controls, *e*) Variable speed spooling, *f*) Editing features, *g*) Adaptability to European CCIR reels; *4. System: a*) Adaptability of transport and electronics for more channels, *b*) Extreme flatness of response over the range of 50 cps to 15 kc.

The attempt to satisfy these requirements resulted in the master tape recorder designated as Model MR-70.

RELIABILITY

Downtime is always very expensive, and technically qualified service personnel scarce and costly. Any failure in the middle of a recording session can easily cost many times the price of the recorder. Reliability, therefore, was emphasized in the design of the new master recorder. One major decision had to be made early in the program: conventional vacuum tubes had failure rates which were obviously too high, so a choice was necessary between transistors and military/industrial nuvistors as active components.

Table I shows failure rates of active components at published maximum ratings. The apparent failure rates of military grade transistors and military/industrial nuvistors as shown on this Table are very similar, with nuvistors actually slightly better. Since there are definite perform-

Table I. Failure rates of active components used at published maximum ratings.*

	Failures per 10^6 hrs	% Failures per 1000 hrs
Industrial/Military Nuvistors (7586 and 7895)	1.3	0.13
Industrial/Military Transistors	2.0	0.2
12AT7 (A or B)	12.4	1.24
12AU7 (A or B)	13.5	1.35
12AX7 (WA)	9.5	0.95
12BA6	17.75	1.75
6X4	42.5	4.25
5Y3	21.9	2.19
5V4	84.8	8.48

* Data taken from: Department of Defense Military Standardization Handbook, *Reliability Stress and Failure Rate Data for Electronic Equipment* (MIL-HDBK-217, 8 August 1962).

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[*] Presented October 17, 1963 at the Fifteenth Annual Fall Convention of the Audio Engineering Society, New York.

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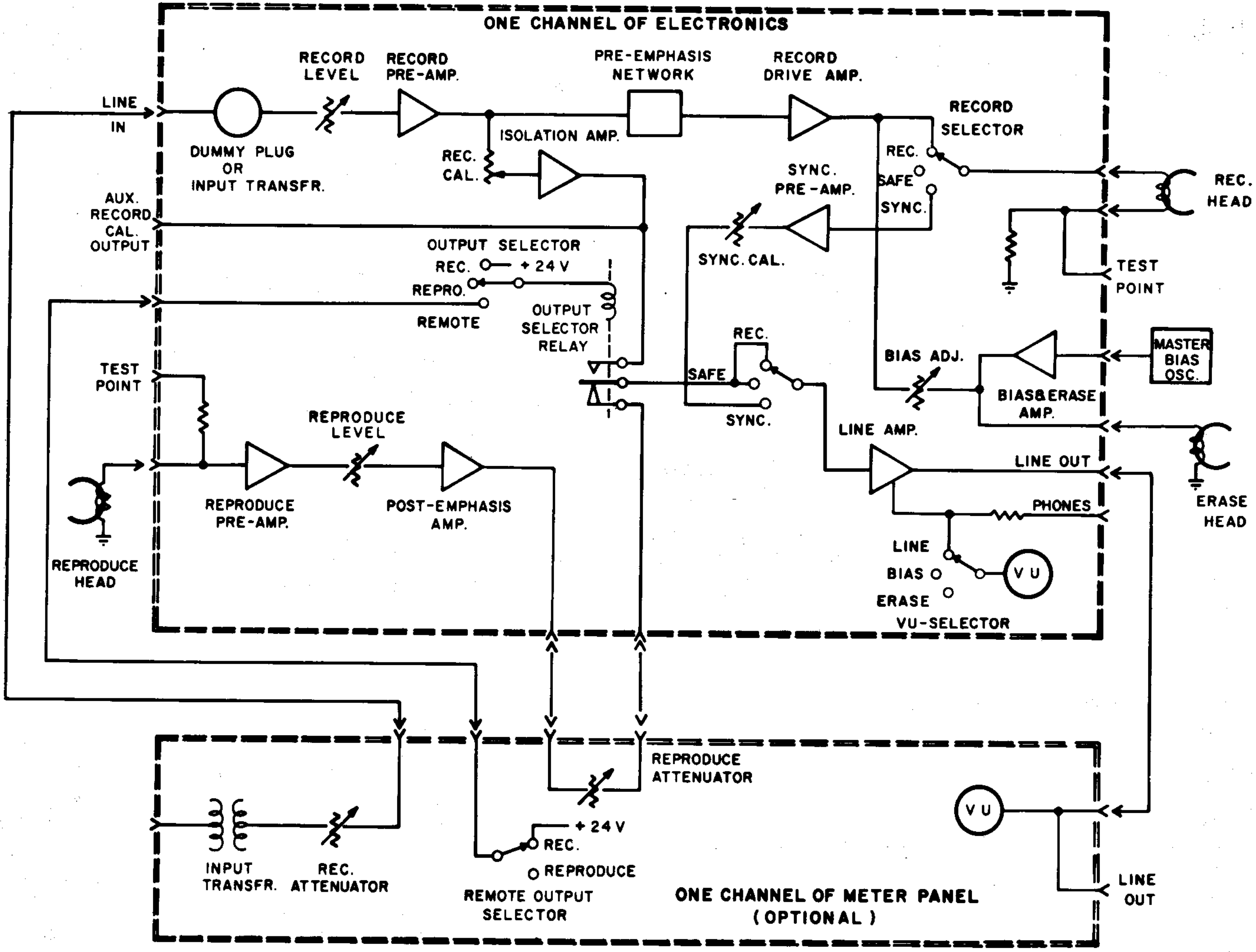

FIG. 1. Simplified block diagram of one channel of the Ampex MR-70 electronics.

ance advantages with nuvistors, they were chosen as the active elements, along with silicon diodes and rectifiers in the power supplies. All other electronic components were chosen for maximum reliability/cost ratio (Table II).

In the tape transport, high-grade (Class 7) ball bearings were used wherever possible. The solenoids were improved by using polished chrome-plated plungers and higher pull-in currents with lower holding currents. The insulation in motors was increased, or changed, to reduce failures.

ELECTRONICS

The block diagram of the recorder (Fig. 1) shows a number of innovations. A master bias and erase oscillator (part of the tape transport) feeds a bias and erase amplifier in the electronics for each channel. This makes it possible to add channels as desired.

Provision has been made to add a remotely located meter panel containing step attenuators, vu meters, input and output terminals and A-B switching; this permits ganging of various controls for multichannel use. A plug-in equalized "Sel-sync" preamplifier module makes each electronics section self-contained and provides better synchronization.

The electronics for each channel (Fig. 2) occupy a 5¼-

TABLE II. Failure rates of passive components at 60°C, 60% RH, 70% of maximum ratings.*

	Failures per 10^6 hrs	% Failures per 1000 hrs
Dipped Mica Capacitor (Mil-C-5B)	0.013	0.0013
Ceramic Capacitor (General Purpose) (Mil-C-11015B)	0.03	0.003
Aluminum Electrolytic Capacitor (Mil-C-62)	1.5	0.15
Tantalum Foil Electrolytic Capacitor (Mil-C-3965B)	0.26	0.026
Solid Tantalum Capacitor (Mil-C-26655A)	1.65	0.165
Paper Capacitor (Mil-C-14157B)	0.17	0.017
Composition Resistor (Mil-R-11)	0.06	0.006
Accurate Wirewound Resistor (Mil-R-93)	1.52	0.152
Power Wirewound Resistor (Mil-R-26)	1.2	0.12

* Data taken from: Department of Defense Military Standardization Handbook, *Reliability Stress and Failure Rate Data for Electronic Equipment* (MIL-HDBK-217, 8 August 1962).

in. rack space. Each set contains five plug-in modules (Fig. 3): two identical output amplifiers (one to serve as the recording drive amplifier, and the other as the line amplifier), a bias amplifier, a Sel-Sync amplifier, and an input transformer. Other electronic circuitry, containing preamplifiers and equalization components, is fix-mounted on the chassis.

As previously described, a careful study was made at the start of this program of the relative merits of transistors and nuvistors. At that time, the best transistors had 6 to 10 db more noise than nuvistors, and 6 to 20 db less overload margin than nuvistors. This made the choice of nuvistors rather obvious, since they had a dynamic range 12 to 30 db greater than transistors. Nuvistors were also advantageous from the point of view of serviceability, since they have closely controlled performance parameters (usually falling within ±10%) compared to transistors —where the beta frequently varies 5 to 1 within a single designation.

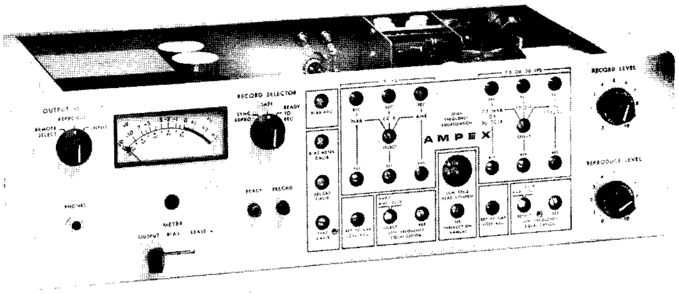

FIG. 2. Electronics for one channel of the MR-70. "Line Amp." module is shown in place, "Record Amp." and "Bias and Erase Amp." modules removed.

Since major improvements have been made in the art of designing and manufacturing solid state devices, periodic re-evaluations of the situation were made; the latest (for the purpose of this paper) was conducted a few weeks prior to the October, 1963 AES Convention. A number of Fairchild 2N2484 silicon planar transistors (known for their low noise and leakage) were compared to a number of RCA 7895 nuvistors. The data taken indicated that the noise spectra of the two groups of devices were approximately equal in the audio band. The transistors were better than the nuvistors when tested with very low source impedances, worse with higher source impedances. The 2N2484 had a clipping level 20 db lower than the nuvistor when each was operated under minimum noise conditions. (It is possible to increase the collector voltage on the 2N2484 to 60 v, so the maximum output before clipping can be only 5 to 6 db below that of the nuvistor, but transistor noise is higher under these conditions.) This re-evaluation indicated that our basic decision, made at the start of the program, is still valid.

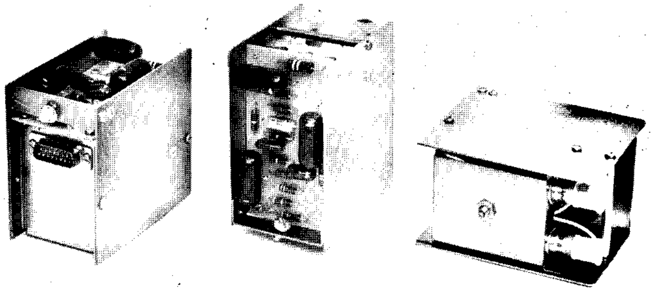

FIG. 3. Plug-in modules for the MR-70.

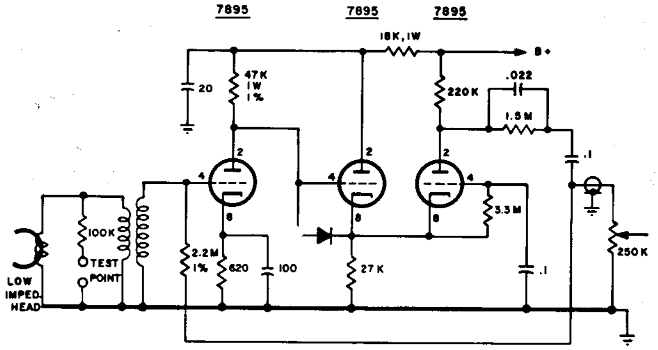

FIG. 4. Reproducer preamplifier circuit diagram.

To achieve the performance level goals, many novel circuits had to be devised. The input circuit is a particularly important example.[1] The circuit, shown in Fig. 4, uses three direct-coupled nuvistors with a feedback loop returned to the grid of the first stage. This circuit makes maximum use of the inherently high input impedance and low equivalent input noise resistance of the nuvistor. Due to this configuration, the noise at 40 cps is reduced by 20 db.

The curves in Fig. 5 show the spectral noise density of an Ampex 300-2 (dashed curve), a modern solid state recorder (dotted curve), and the Ampex MR-70 (lower solid curve). As shown by these curves, the noise of the MR-70 reproducing circuit is substantially below that of the other two reproducers, and is well below the biased tape noise of a typical low noise tape (upper solid curve). The combi-

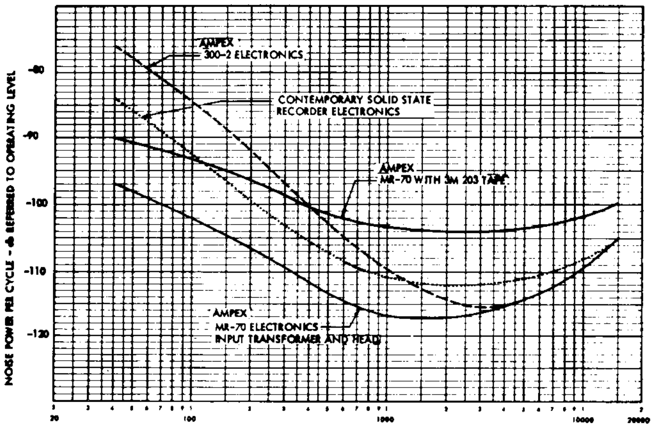

FIG. 5. Spectral noise density: equipment noise (reproducing head and amplifier only) for the Ampex 300-2, for the contemporary solid-state professional recorder, and for the new MR-70; system noise for the MR-70 with 3M No. 203 low-noise tape. All for two track, ¼-in. tape at 15 ips, with NAB equalization.

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[1] E. P. Skov, "Noise Limitations in Tape Reproducers," *J. Audio Eng. Soc.* **12**, 280 (Oct., 1964).

nation of the low noise reproducing circuit and the symmetrical 150 kc bias current produces recordings that are limited in noise only by the tape characteristics. A full-track 15 ips NAB recording on 3M No. 203 tape produces a weighted noise 73 db below the 3% third harmonic distortion level. The equipment noise (i.e., tape stopped) is another 9 db lower, providing a generous margin for any future improvements in tape.

The unweighted noise in a 20 cps to 15 kc band is 70 db below the same reference level, and the unweighted equipment noise another 8 db lower than this. Thus this low noise electronic circuitry used with the low noise tape will better the present typical unweighted noise specification by 10 db.

The current state of the art in electronics permits the design of amplifiers with very low distortion. To measure the reproducing amplifier distortion of the MR-70, for instance, the input signal from a low-distortion oscillator was fed into the reproduction section Test Point (shown in Fig. 1), and the output from the Line Out terminals was measured with a voltmeter and analyzed with a wave analyzer.[2] A similar test method was used with two other recorders. Figure 6 shows the resulting distortion curves. Since the saturated tape output is approximately 14 db above operating level, these curves indicate that the MR-70 electronics will not exceed 1% distortion even when the tape is saturated.

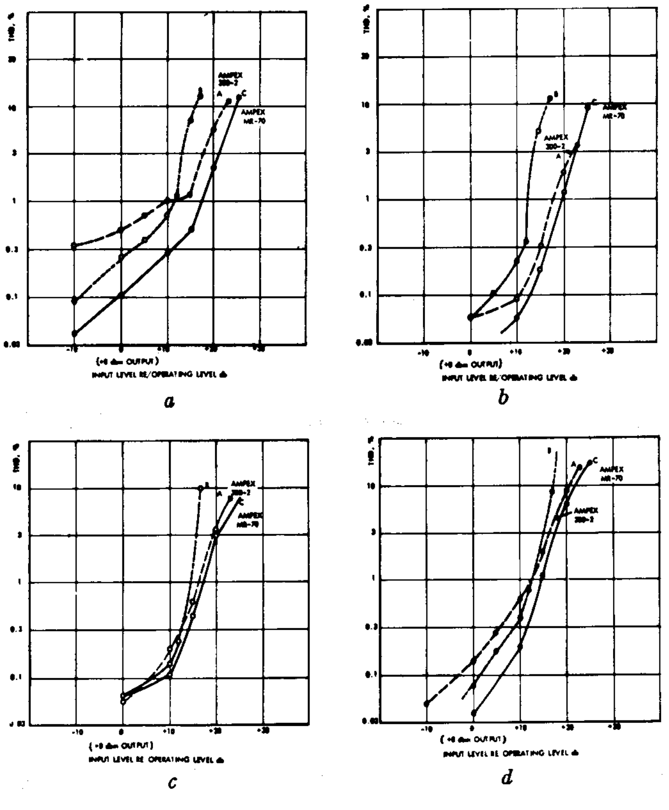

FIG. 6. Total harmonic distortion *vs* input level for three reproducing amplifiers at four frequencies. *a*. Distortion at 40 cps; *b*. Distortion at 500 cps; *c*. Distortion at 5 kc; *d*. Distortion at 15 kc. *Curves A* are for Ampex 300-2; *Curves B*, contemporary solid-state professional recorder; *Curves C*, Ampex MR-70. All data for NAB 15 ips post-emphasis; rms sum of harmonic components shown.

Probably the most strongly voiced requirement in the area of distortion was the need for a greater overload margin in recording amplifiers. To gather evidence on this feature, distortion *vs* level was measured for the recording systems corresponding to the reproducing systems measured before. The input signal was fed into Line In, and the output signal was taken from the recording section Test Point (as shown for the MR-70 in Fig. 1, and similarly for the other recorders). As evidenced by the curves of Fig. 7, a very substantial reduction in recording amplifier distortion was accomplished.

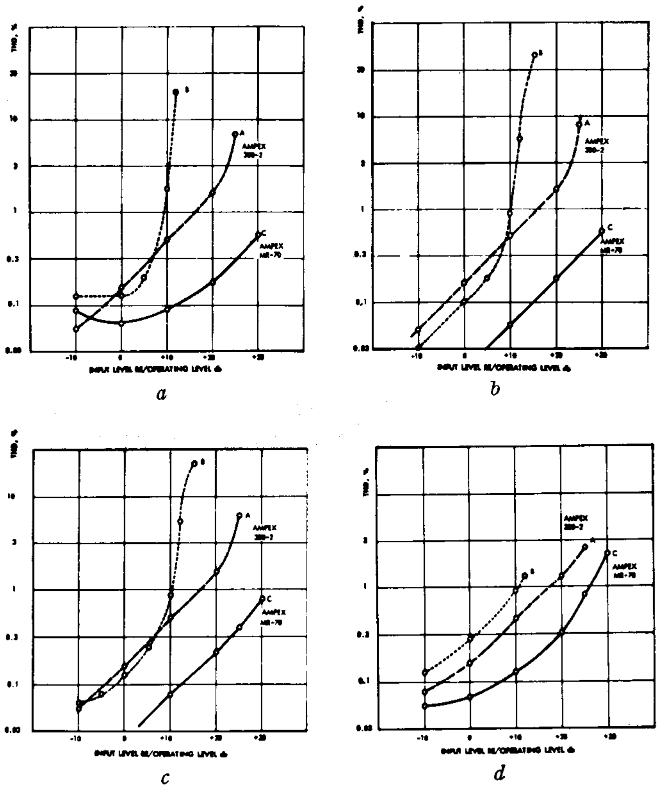

FIG. 7. Total harmonic distortion *vs* input level for three recording amplifiers, at four frequencies. *a*. Distortion at 40 cps; *b*. Distortion at 500 cps; *c*. Distortion at 5 kc; *d*. Distortion at 15 kc. *Curves A*, Ampex 300-2; *Curves B*, contemporary solid-state professional recorder; *Curves C*, Ampex MR-70. All data for 15 ips, pre-emphasis set for reproduction with NAB post-emphasis; rms sum of harmonic components shown.

The dynamic range of a recording may be limited by the tape or by the electronics. This is discussed in further detail in a companion paper,[3] but it may be stated in a general way that present low noise tape required improvements in the electronics to fully realize its potential. The MR-70 electronics has sufficient recording power and sufficiently low reproducing system noise so that it would not

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[2] Test equipment was as follows: Low distortion oscillator, Burr Brown 9147 B (second harmonic, –80 db; third harmonic, –76 db); Voltmeter, Hewlett Packard Model 400L; Wave analyzer, Hewlett Packard Model 302 A.

[3] R. Z. Langevin, "Dynamic Range Limitations in Tape Recording," *J. Audio Eng. Soc.* 12, 294 (Oct., 1964).

limit the potentialities of a tape with an additional 10 db improvement in dynamic range.

HEAD ASSEMBLY

In order to fulfill the desired requirements, a new head assembly had to be designed. A major change from previous Ampex head assemblies is the use of fixed azimuth and zenith on the heads. This means that the mounting surface for each head is machined to be perpendicular to the gap line with extremely close tolerances for both azimuth and zenith, and no adjustment is required. This machining operation is performed on a precision jig borer, with high-power optics used to align the cutter to 90° in relation to the gap line.

A new "common-pole" ferrite erase head was also designed. This increases the erase efficiency and (since it is parallel-fed) helps retain a higher Q in a tank circuit. This, in turn, tends to reduce bias distortion.

TAPE TRANSPORT

In analyzing the requirements for better tape transport performance, six areas of needed improvement were isolated. They were: *1.* Improvement in tape guidance, *2.* Reduction of flutter, *3.* Reduction of frequency modulation noise ("scrape flutter"), *4.* Improvement in starting time, *5.* Improvement in speed stability, and, *6.* Improvement in azimuth stability.

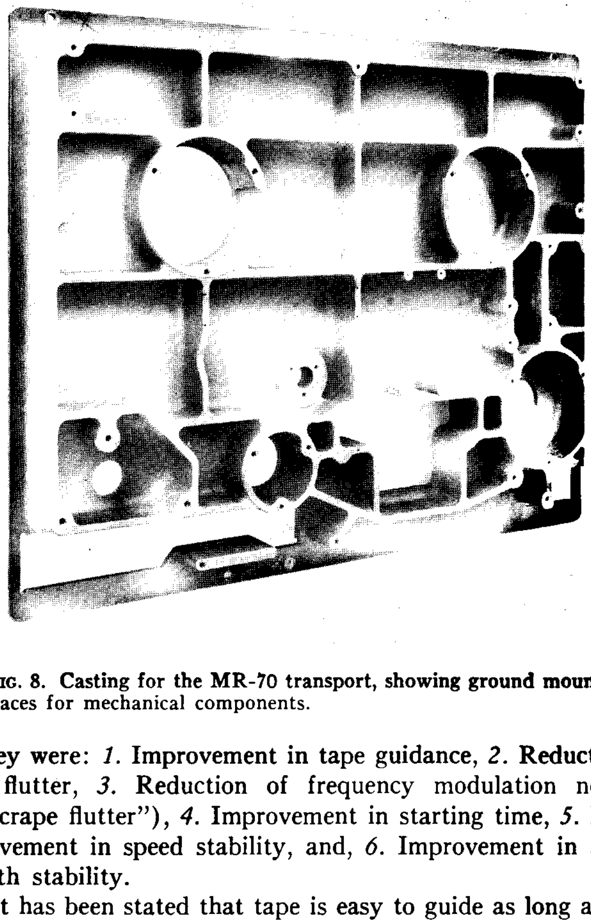

FIG. 8. Casting for the MR-70 transport, showing ground mounting surfaces for mechanical components.

It has been stated that tape is easy to guide as long as it is allowed to follow the path it wants to take. This means that the axes of all rotating elements (e.g., reel spindles, reel idler, capstan and capstan idler) must be parallel to each other and perpendicular to the tape path. This is difficult to achieve unless a very stable and rigid frame is used to mount the various mechanical components. In the MR-70, a new large casting was designed (Fig. 8), with top and bottom surfaces ground flat and parallel within 3 mil over the entire surface. A long tape guide was provided as the tape approaches and rides on the reel idler, since the tape has the greatest guiding strength here where it is wrapped around the reel idler. This design has considerably improved the high frequency amplitude stability of the equipment.

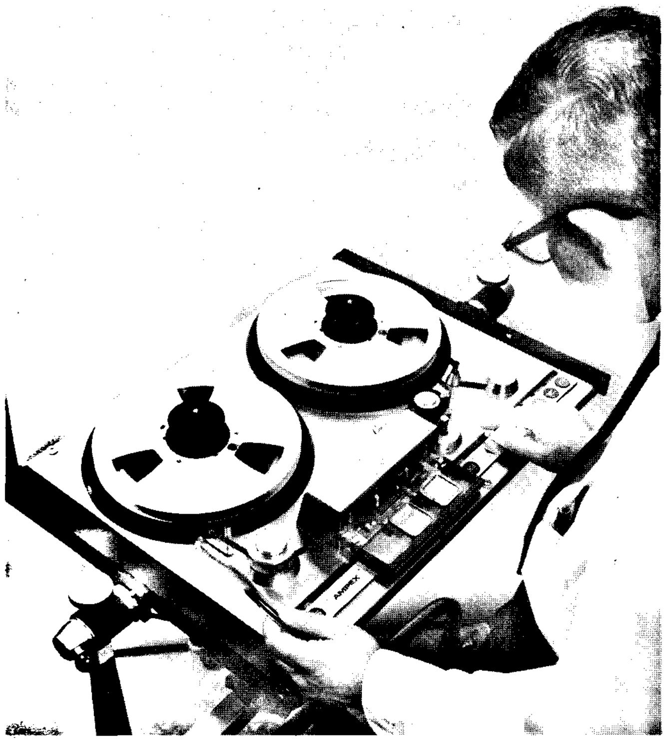

FIG. 9. Editing with the MR-70 transport.

While many of the proven principles of the Ampex Model 300 tape transport have been retained in the MR-70 design, the actual design details have been greatly modified to cope with the increased demands for performance and reliability. Added flexibility was requested for the control system, and a new system was developed which provides continuously variable winding speeds, automatic tape and head-gate lifters, and several editing features (see Fig. 9).

Ease of loading and editing is of paramount importance; we felt that a closed T-loop drive would complicate these functions, and so did not seriously consider this type of drive. The required flutter performance was attainable with a single capstan drive utilizing an indirect drive system and a viscous damper.[4] The MR-70 is capable of total rms flutter performance around 0.03%, including all components from 0.5 to 250 cps.

Frequency modulation noise is a function of the roughness of tape, guide, and head surfaces, and the geometry of the tape transport. Quantitative discussions of this phenome-

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[4] J. G. McKnight, "Mechanical Damping in Tape Transports," *J. Audio Eng. Soc.* 12, 140 (April, 1964).

non, and means for its reduction are given by Werner,[5] and by Belger and Heidorn.[6] (A largely qualitative introduction in English is given by von Behren and Youngquist.[7]) An idler in the head assembly of the MR-70 essentially eliminates the FM noise (scrape flutter), as is shown by Fig. 10.

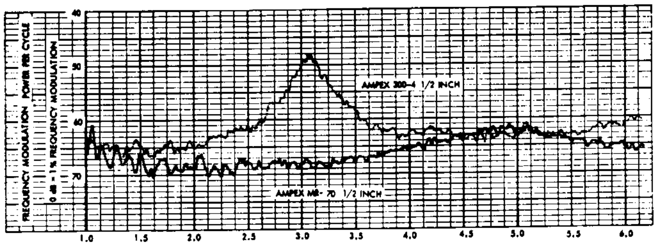

FIG. 10. Spectral density of frequency modulation (scrape flutter): *dotted line,* Ampex 300-4; *solid line,* Ampex MR-70, with an idler in the head assembly. The 23 kc carrier frequency is recorded, reproduced through Ampex FR-100 "fm reproducing amplifier" as demodulator, and analyzed with GR 1900-A Wave Analyzer.

For most present-day tape transports, several seconds are required for the transport to achieve stable tape motion. In the MR-70, an additional torque motor accelerates the reel idler to the correct speed, and a damped flywheel suppresses any oscillations. This permits the MR-70 to achieve stable tape motion in approximately ½ second.

A constant tension supply system was found necessary to maintain speed and azimuth stability. The position of the tape as it leaves the supply reel is sensed by the arm of a mechano-electric transducer which in turn controls the voltage to the supply motor so as to produce constant tape tension at any tape-pack diameter.

CONCLUSION

We feel that the majority of the requirements of the recording industry have been met in the actual performance of the MR-70. We may have inadvertently overlooked some requirements, and undoubtedly some solutions will not be universally approved. We believe, however, that this recorder will improve the art of sound recording and reproduction, and hope that it will stimulate further developments in techniques, tapes, and equipment.

ACKNOWLEDGEMENTS

The authors wish to thank the many members of the recording industry for their suggestions, and the entire Ampex Corporation Audio Engineering Department for work in designing the recorder and helping with the collection of data for this paper.

Our particular thanks to Edward Conti, R. Z. Langevin, John G. McKnight, William Pottberg, George Rehklau, and Erling Skov. Without their efforts neither this paper nor the MR-70 would exist.

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[5] P. H. Werner, "Die mechanischen Eigenschaften verschiedener Magnettonbaender und ihr Einfluss auf die Aufnahmequalitaet," ("The Mechanical Properties of Various Magnetic-Recording Tapes and Their Influence on Recording Quality") *Technische Mitteilungen, Schweizerische Post Telegraphen und Telephonverwaltung* 30, 173 (1952). In German and French.

[6] E. Belger and G. Heidorn, "Ueber die Laengsschwingungen von Magnettonbaendern" ("Longitudinal Vibrations in Magnetic Tapes"), *Rundfunktechnische Mitteilungen* 3, 51 (1959). In German.

[7] R. A. von Behren and R. J. Youngquist, "Frequency-Modulation Noise in Magnetic Recording," *J. Audio Eng. Soc.* 3, 26 (1955).

Noise Limitations in Tape Reproducers*

ERLING P. SKOV

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Thermal noise, shot noise, 1/*f* noise, induced grid noise and Barkhausen noise are described and quantitative examples are given. Expressions for thermal noise in mixed resistive and reactive circuits are derived. The relationship between head thermal noise and amplifier shot noise is discussed and measured data are given. The question of tubes *vs* transistors is touched upon with respect to noise and overload capabilities. The relationship between the noise index of a resistor and the crossover frequency between thermal and 1/*f* noise is calculated and displayed as a family of curves, and examples are given. Signal-to-noise measurement data from the Ampex MR-70 master recorder are given along with a brief description of the basic design philosophy with respect to noise reduction.

INTRODUCTION

MAGNETIC recording has been performed since 1899, when Valdemar Poulsen filed patent application in the United States for his telegraphone. His original process has seen considerable improvements through the years: W. L. Carlson and G. W. Carpenter (ac-biasing, 1927), Kurt Stille and Blattner (steel tape), Pfleumer (coated paper and plastic tapes), C. N. Hickman of the Bell Telephone Laboratories (vicalloy tape, 1937), are just some of the contributors.[1]

Improvements have continued, and in attempting to obtain the best signal-to-noise ratio in the widest frequency band, the limitations in overload distortion and noise have been investigated more closely, resulting in practical improvements. Lower-noise tubes and transistors have come of age, and the trend in tape development has been towards lower noise, better saturation characteristics at short wavelengths, smoother surfaces for less spacing loss, more homogeneous tapes for fewer dropouts, and better wear characteristics, to mention only some of the advances already obtained. The equipment designer will have to keep pace with tape improvements by providing lower noise head-amplifier designs; better yet, he should keep at least 10 db ahead for possible tape improvements.

Although many noise sources are treated in textbooks, this paper will attempt to pool them all together and to point out some of the basic signal-to-noise limitations in the specific field of audio recording.

NOISE MEASUREMENT TECHNIQUE

Noise was measured as a function of frequency, in small bands of constant width; this measurement may be called the "noise frequency response."[2]

The noise frequency response was measured with a HP Model 302A wave-analyzer with a 6 cps constant bandwidth. The speed of the meter on this analyzer is much too fast for noise measurements in such a narrow band; its fluctuations made readings very difficult. The averaging time should be several times the reciprocal of the bandwidth in cps:[3]

$$t = 1/6\ \text{cps} \times 3 = 1/2\ \text{sec.}$$

The addition of a capacitor across the meter movement achieves this result. The meter resistance is 100 ohm; therefore,

$$C = t/R = \tfrac{1}{2}/R = 5000\ \mu\text{F}.$$

The value 6000 μF was used in all noise measurements in this paper. A more correct method would have been to use an averaging thermocouple meter which would read e^2 instead of e_{av}, the error, however, is not serious and of course does not appear in relative measurements. The narrow bandwidth of the analyzer makes it possible to measure fairly accurate noise frequency responses of devices with sharply varying frequency responses such as high-Q circuits.

TYPES OF NOISE

Five different types of noise will be considered, along with the devices in which they occur. Among amplifying devices only tubes and transistors will be considered,[4] since

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* To be presented October 13, 1964 at the Sixteenth Annual Fall Convention of the Audio Engineering Society, New York.

they constitute the majority of amplifiers in use. Hum and microphonic effects will not be mentioned since they do not present absolute physical limitations.

The five basic noise sources and the devices in which they are of practical importance are as follows:

1. Thermal noise:	Resistors, transistors, tape heads, transformers
2. Shot noise:	Tubes, transistors
3. $1/f$ noise:	Resistors, tubes, transistors
4. Induced grid noise:	Tubes
5. Barkhausen noise:	Tape heads, transformers

Thermal Noise

Thermal or Johnson noise is a measure of the kinetic energy of the conduction electrons. This noise, if measured with a constant bandwidth wave analyzer, will exhibit a flat frequency response (white noise). This type of noise is found in resistors, in the base area of transistors,[5] and in tape playback heads and transformers, where the combined copper resistance and effective resistance from eddy current losses will produce thermal noise. (Thermal noise in vacuum tubes appears to be negligible at audio frequencies.) Wire-wound resistors have thermal noise only; that is, noise will not increase if dc current is sent through them, unless *1.* the dissipated power increases the temperature of the resistor or *2.* it has faulty contacts. In low-level input circuits, however, the heat rise is negligible.

Example 1: How many db will the thermal noise increase in a wire-wound resistor if its temperature is raised 30°C above room temperature?

Noise voltage, $e_n = \surd(4kTRB)$, where $k =$ Boltzmann's constant $(1.374 \times 10^{-23}$ Joules/°K), $T =$ absolute temperature in °K, $R =$ resistance in ohm, $B =$ bandwidth in cps.

Noise voltage increase for 27°C room temperature (300°K):

$$20 \log \surd(4kT_1RB/4kT_2RB) = 20 \log \surd[(300+30)/300] = 0.4 \text{ db}.$$

Film resistors exhibit thermal noise and often also some semiconductor ($1/f$) noise. Films of thicknesses less than the mean free path of the conduction electrons will show higher resistivity than the bulk material,[6] due to the scattering of the electrons when they meet the surfaces of the film.

The base resistance of transistors will only contribute to noise in transistors with low shot noise.

Thermal noise is present in any device with a resistive component, when the temperature is sufficiently above 0°K. Normal amplifier circuits operate near or above room temperature so their resistive components do produce thermal noise. The reactive circuit components do not produce noise but serve only as energy storage devices. They will modify the frequency distribution of the thermal noise produced by the resistive components.

The noise frequency response of five basic resistive-reactive combinations will now be investigated: 1) Two resistors in parallel; 2) A resistor and a capacitor in parallel; 3) A resistor and an inductor in parallel; 4) A resistor and a loss free parallel resonant circuit in parallel; 5) A resistor and a lossy parallel resonant circuit in parallel.

Sample Calculations

1. Derive the noise frequency response of two resistors in parallel (see Fig. 1):

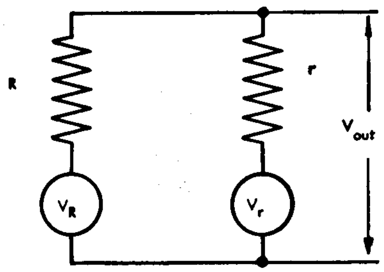

FIG. 1. Circuit for calculating the noise voltage from two resistors in parallel.

V_r contribution to V_{out}: $[(R/r+R)]V_r$

V_R contribution to V_{out}: $[(r/r+R)]V_R$

Total noise: $\surd(\{[V_r\cdot(R/r+R)]^2+[V_R\cdot(r/r+R)]^2\})$

Since $V_r^2 = 4kTBr$ and $V_R^2 = 4kTBR$;

Total noise:

$$\surd\{[4kTBr(R/r+R)^2+4kTBR(r/r+R)^2]\} = \surd(4kTB) \times \surd(rR/r+R)$$

which is the thermal noise of a resistor composed of r and R in parallel.

2. Derive the noise frequency response of a resistor and a capacitor in parallel (see Fig. 2):

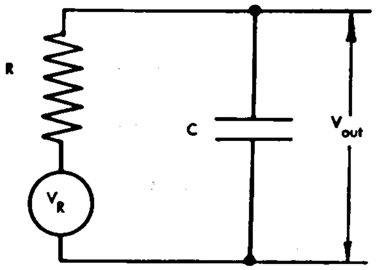

FIG. 2. Circuit for calculating the noise voltage from a resistor and a capacitor in parallel.

V_R contribution to V_{out}:

$$V_R \frac{1/j\omega C}{R+(1/j\omega C)} = V_R \frac{1}{j\omega CR+1}.$$

Since $V_R^2 = 4kTBR$,

$$V_{out} = \surd(4kTBR)/(j\omega CR+1).$$

If $j\omega CR << 1$: $V_{out} = \surd(4kTB) \times \surd(R)$ (low frequencies)

If $j\omega CR >> 1$: $V_{out} = \surd(4kTB)/[j\omega C \times \surd(R)]$ (high frequencies)

At low frequencies where the capacitor has no effect, the noise is pure thermal noise from R; therefore, the noise voltage increases 3 db with a doubling of R $(V_{out} \propto \surd(R))$.

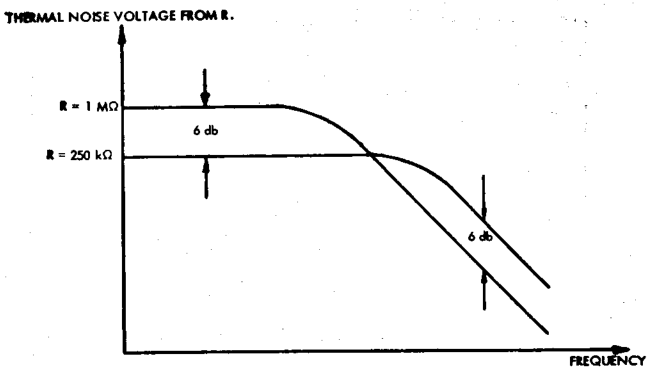

FIG. 3. Noise voltage frequency response of a resistor R and a capacitor in parallel, with R as parameter and $C = 300$ pF.

At high frequencies the noise voltage decreases 6 db/oct with frequency ($V_{out} \propto 1/j\omega C$), and 3 db with a doubling of R ($V_{out} \propto 1/\surd(\mathrm{R})$).

Example 2: Plot the noise curves for $C = 300$ pF in parallel with $R = 1$ Megohm and $R = 250$ kilohm respectively (see Fig. 3). This shows that greater resistance produces more noise at low frequencies, but less noise at high frequencies.

3. Derive the noise frequency response of a resistor and an inductor in parallel (see Fig. 4):

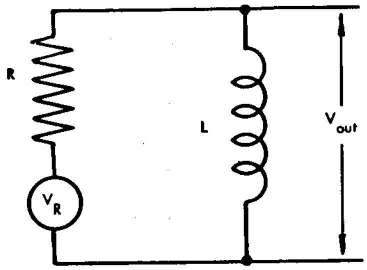

FIG. 4. Circuit for calculating the noise voltage from a resistor and an inductor in parallel.

V_R contribution to V_{out}:

$$V_R \frac{j\omega L}{j\omega L + R} = V_R \frac{1}{1 + (R/j\omega L)}.$$

Since $V_R = \surd(4kTBR)$,

$$V_{out} = \surd(4kTBR)/[1 + (R/j\omega L)].$$

If $R << j\omega L$: $V_{out} = \surd(4kTB) \times \surd(R)$

(high frequencies)

If $R >> j\omega L$: $V_{out} = [\surd(4kTBR)]\, j\omega L/R$

$= j\omega L\, \surd(4kTB)/\surd(\mathrm{R})$

(low frequencies)

At high frequencies where the inductor has no effect, the noise is pure thermal noise from R; therefore, the noise voltage increases 3 db with a doubling of R ($V_{out} \propto \surd(R)$. At low frequencies the noise voltage increases 6 db/oct with frequency ($V_{out} \propto j\omega L$) and decreases 3 db with a doubling of R ($V_{out} \propto 1/\surd(R)$.

Example 3: Plot the noise curves for $L = 100$ H in parallel with 1 Megohm and 250 kilohm respectively (see Fig. 5). This shows that the greater resistance produces more noise at high frequencies, but less noise at low frequencies.

4. Derive the noise frequency response of a resistor and a loss-free parallel resonant circuit in parallel (Fig. 6):

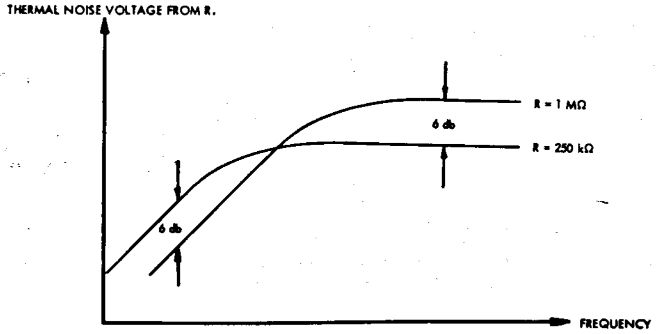

FIG. 5. Noise voltage frequency response of a resistor R and an inductor in parallel, with R as parameter and $L = 100$ H.

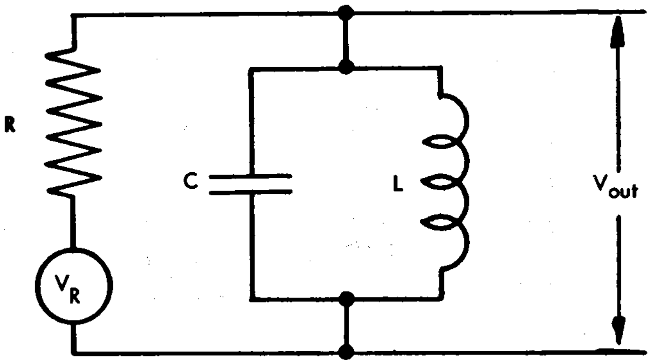

FIG. 6. Circuit for calculating the noise voltage from a resistor and a loss-free parallel resonant circuit in parallel.

V_R contribution to $V_{out} =$

$$= V_R \frac{[j\omega L(1/j\omega C)]/[j\omega L + (1/j\omega C)]}{R + \{[j\omega L(1/j\omega C)]/[j\omega L + (1/j\omega C)]\}}$$

$$= V_R \left\{ 1 \middle/ \frac{R[1-(\omega/\omega_0)^2]}{j\omega L} + 1 \right\}$$

where $\omega_0{}^2 = 1/LC$.

At resonance ($\omega = \omega_0$):

$$V_{out} = V_R = \surd(4kTB) \times \surd(R).$$

Below resonance ($\omega << \omega_0$):

$$V_{out} = V_R\{1/[(R/j\omega L) + 1]\}$$

which is identical to *3.* above.

Above resonance ($\omega >> \omega_0$):

$$V_{out} = V_R[1/(j\omega CR + 1)]$$

which is identical to *2.* above.

At resonance where the parallel circuit has no effect (infinite impedance) the noise is pure thermal noise from R. At low frequencies (below resonance) the noise voltage increases 6 db/oct with frequency ($V_{out} \propto j\omega L$) and decreases 3 db with a doubling of R ($V_{out} \propto 1/\surd(R)$. At high frequencies (above resonance) the noise voltage decreases 6 db/oct with frequency ($V_{out} \propto 1/j\omega C$) and 3 db with a doubling of R ($V_{out} \propto 1/\sqrt{(R)}$.

Example 4: Plot the noise curves for $L = 100\,\mathrm{H}$, $C = 300\,\mathrm{pF}$, ($f_0 = 900$ cps) and $R = 1$ Megohm and 250 kilohm, respectively. This may be done by any one of three approaches:

a) Exactly, by direct calculation from the equation

$$V_{out} = V_R \frac{1}{\{R[1-(\omega/\omega_0)^2]/j\omega L\}+1}$$

b) Approximately, by first locating the voltage at resonance $[(V_{out} \text{ at } \omega_0) = \sqrt{(4kTB)} \times \sqrt{(R)}]$; this is the voltage due to R alone, since the L-C circuit is assumed to have an infinite impedance at resonance. The 6 db/octave asymptotes may then be located, using the knowledge that the resonant Q may be determined from the equation $Q = R/\omega_0 L$, and that the asymptotes meet at the point where $Q = 1$. In this example, when $R = 1$ Megohm, $Q = 1.77$; this ratio is 5 db. Therefore the intersection of the asymptotes must be 5 db below the resonant impedance, as is shown in Fig. 7. Likewise, when $R = 250$ kilohm, $Q = 0.44$; this ratio is –7 db. Therefore this intersection of asymptotes must be 7 db above the resonant impedance, as also shown.

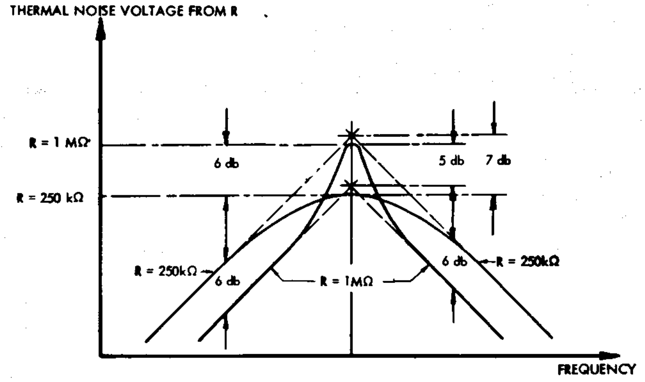

FIG. 7. Noise voltage frequency response of a resistor R and a loss-free parallel resonant circuit in parallel, with R as parameter, $L = 100\,\mathrm{H}$ and $C = 300\,\mathrm{pF}$.

c) Approximately, by locating the voltage at resonance $[(V_{out} \text{ at } \omega_0) = \sqrt{(4kTB)} \times \sqrt{(R)}]$, and by using the curves of *Examples 2* and *3* at the frequencies well above and well below resonance to determine the asymptotes of response.

5. Derive the noise frequency response of a resistor and a lossy parallel resonant circuit in parallel (see Fig. 8):

V_R contribution to V_{out} =

$$= V_R \frac{[(j\omega L+r)(1/j\omega C)]/[(j\omega L+r)+(1/j\omega C)]}{R+[(j\omega L+r)(1/j\omega C)]/[(j\omega L+r)+(1/j\omega C)]}$$

$$= V_R \frac{j\omega L+r}{R[1-(\omega/\omega_0)^2]+j\omega CRr+j\omega L+r}.$$

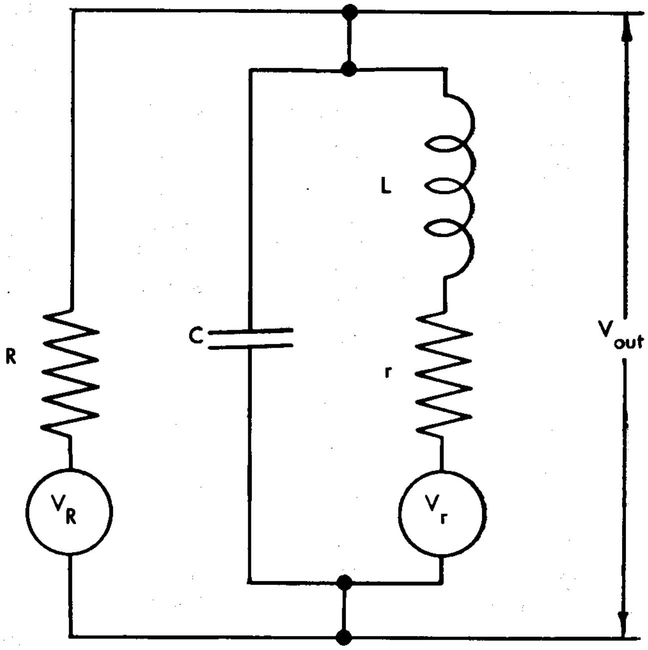

FIG. 8. Circuit for calculating the noise voltage from a resistor and a lossy parallel resonant circuit in parallel.

V_r contribution to V_{out} =

$$= V_r \frac{[R(1/j\omega C)]/[R+(1/j\omega C)]}{r+j\omega L+\{R(1/j\omega C)/[R+(1/j\omega C)]\}}$$

$$= V_r \frac{R}{R[1-(\omega/\omega_0)^2]+j\omega CRr+j\omega L+r}.$$

Total noise voltage V_{out} =

$$= \sqrt{\frac{V_R{}^2(j\omega L+r)^2+V_r{}^2R^2}{\{R[1-(\omega/\omega_0)^2]+j\omega CRr+j\omega L+r\}^2}}$$

$$= \sqrt{(4kTB)} \times \frac{\sqrt{[R(j\omega L+r)^2+rR^2]}}{R[1-(\omega/\omega_0)^2]+j\omega CRr+j\omega L+r}.$$

At very low frequencies ($\omega \sim 0$),

$$V_{out} = \sqrt{(4kTB)}\,[\sqrt{(Rr^2+rR^2)}]/(R+r)$$
$$= \sqrt{(4kTB)}\,\sqrt{[Rr/(R+r)]},$$

which is the thermal noise of r and R in parallel.

At resonance ($\omega = \omega_0$),

$$V_{out} = \sqrt{(4kTB)} \frac{\sqrt{[R(j\omega_0 L+r)^2+rR^2]}}{j\omega_0 CRr+j\omega_0 L+r}.$$

Example 5: Find V_{out} at resonance when $R = 1$ Megohm ($f_0 = 900$ cps), $L = 100\,\mathrm{H}$, $C = 300\,\mathrm{pF}$ and $r = 10$ kilohm. Plot the noise curves for $R = 1$ Megohm and 250 kilohm respectively (see Fig. 9).

V_{out} =

$$= \frac{\sqrt{\{4kTB[10^6(j2\pi\times 900\times 100+10^4)^2+10^4\times 10^{12}]\}}}{j2\pi 900\times 300\times 10^{-12}\times 10^6\times 10^4+j2\pi 900\times 100+10^4}$$

or $V_{out} =$

$$= \sqrt{(4kTB)} \times \sqrt{(0.886 \text{ Megohm} + j\,0.016 \text{ ohm})}$$

The impedance at resonance is thus reduced from 1 Megohm to 0.886 Megohm because r prevents the resonant circuit impedance from reaching infinity.

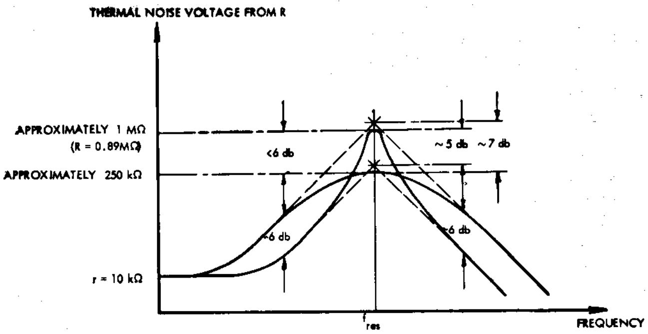

FIG. 9. Noise voltage frequency response of a resistor R and a lossy parallel resonant circuit in parallel, with R as parameter, $L = 100$ H, $C = 300$ pF and $r = 10$ kilohm.

Thermal Noise in a Tape Reproducing Head

The circuit configuration in Section *5.* in the Thermal Noise section (Fig. 8) above is typical of a tape reproducing head. The dc resistance of the copper winding is constant $= r$; the parallel resistance R due to eddy current losses in the laminations varies with frequency as a function of the mechanical, electrical and magnetic properties of the laminations. A lamination can be considered a "shorted" turn around the flux lines through it, (Fig. 10), the turn having a finite resistance similar to a shorted turn on a regular transformer. The resistance of this single turn, R', times n^2, where n is the total number of turns on the head, is the transformed parallel resistance $R = n^2R'$ of one lamination as measured across the head winding. Since eddy currents will resist the penetration of magnetic fields into the central parts of a lamination, the induction decreases from the surface towards the center when a magnetic field is applied from the outside. If the magnetic field is developed inside the lamination by the thermal noise in the conducting lamination material, the outer layers will both produce thermal noise and act as shields for the magnetic field produced by the thermal noise near the center of the lamination, so that R cannot be calculated directly from the cross-sectional area of the lamination. For the complete head all the laminations must be considered and the noise voltages added in rms fashion, since the laminations are independent, uncorrelated and non-concentric noise sources.

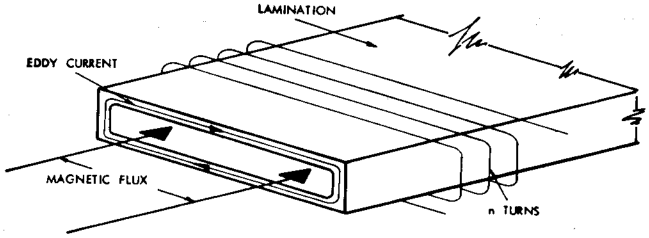

FIG. 10. Cross-sectional view of a lamination, indicating the paths of the eddy currents and the magnetic flux.

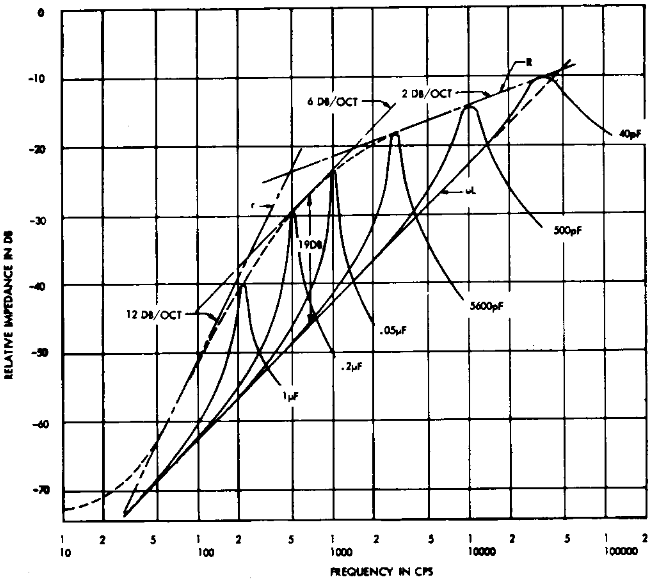

FIG. 11. Measured relative impedance of a 0.5 H playback head ($R = 130$ kilohm at 57 kc and $r = 100$ ohm) with capacitance as parameter.

The permeability μ, the resistivity ρ, and the lamination thickness d of the laminations, and the frequency f will influence the effective R. Figure 11 shows a measurement of the impedance *vs* frequency for a sample reproducing head. The head is resonated with different low-loss capacitors. Note especially that R (the effective parallel resistance, which is of course also the impedance at resonance) *vs* frequency appears to have a slope of 2 db/oct. Two db/oct variation of R may also be expressed as 1 db/oct

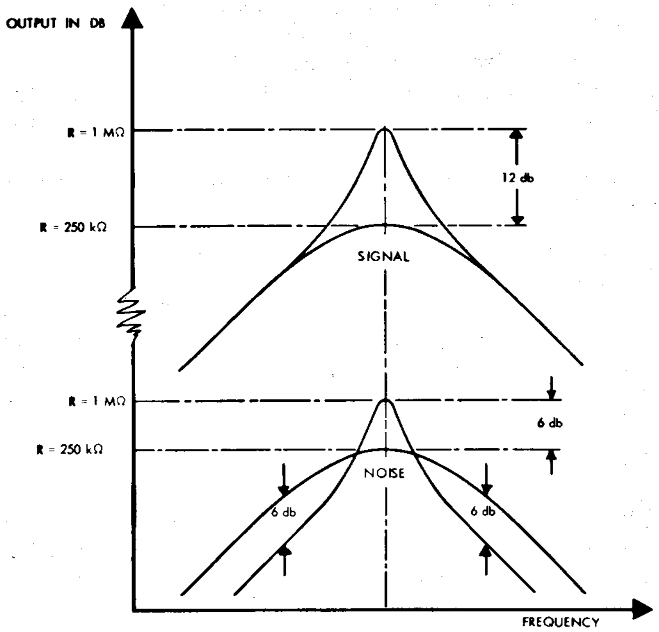

FIG. 12. Signal and noise frequency responses of a tape reproduce head with eddy current losses (R) as parameter.

variation of $\surd(R)$; this represents a signal-to-noise ratio increase with frequency of 1 db/oct due to the fact that the signal voltage is proportional to R and the noise voltage to $\surd(R)$. The signal-to-noise ratio is, therefore, proportional to $R/\surd(R) = \surd(R)$. Figure 12 displays two signal and noise curves with R as parameter which illustrates this relationship. Referring again to Fig. 11, at 450 cps the curve of R intersects a 12 db/oct curve (ω^2) which represents the copper loss due to r. Since Q can be expressed in two ways: $Q = \omega L/r$ or $R/\omega L$, the series resistance r can be transformed to a parallel resistance R using the relationship $R = \omega^2 L^2/r$ where ω^2 represents the 12 db/oct slope. $Q = 1$ occurs at 57 kc where the R curve crosses the ωL curve. Maximum Q is seen to occur in the 500-1000 cps range and is 19 db or 8.9.

According to Section *4.* of the Thermal Noise section (a resistor and a loss-free parallel resonant circuit in parallel, the noise voltage *vs* frequency away from resonance has 6 db/oct slopes. The curve representing the real (noise-producing) part of the head impedance must then have 12 db/oct slope because of the relationship between noise voltage and the value of the noise-generating resistance ($e^2 = 4kTBR$). In the Appendix is given the mathematical demonstration of the 12 db/oct slopes, and Figure 13 shows ωL and the real part of the head impedance *vs* frequency as measured on a Dranetz Complex Impedance-Admittance meter, using the same head as used in Fig. 11. (The slopes are not exactly 12 db/oct because of the finite dc winding resistance.) Figure 13 also shows $Q = 1$ ($\omega_0 L = R$) near 57 kc.

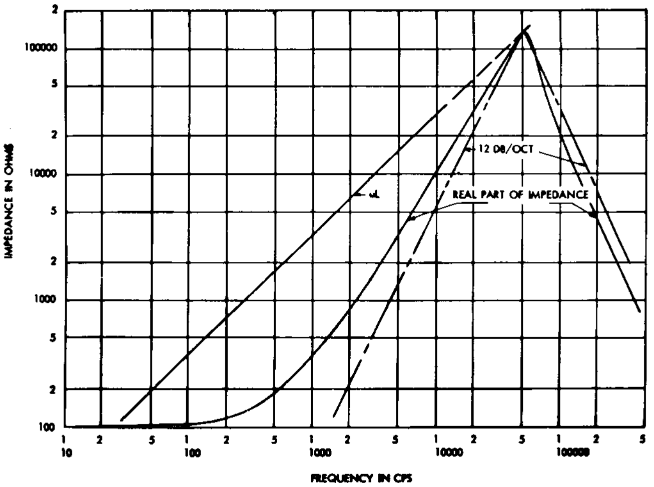

FIG. 13. Measured reactive (ωL) and real impedance of the same head used for the measurements of Fig. 11.

The inductance of tape reproducing heads for audio is often in the 0.5 to 2 H range, which has been considered a good compromise between high overall output and high resonant frequency. Most amplifier inputs look like a capacitance and a resistance in parallel. By varying these, the frequency and Q of the head resonance can be adjusted to compensate for reproducing head gap losses.

If higher inductance heads are desirable for higher overall voltage output, it becomes increasingly more difficult to keep the head resonance above the audible range. Figure 14 is a graph of the relative head impedances of three heads with the following inductances: 1.5 H, 15 H and 150 H. All three heads were made from similar laminations and were connected to the grid of a Nuvistor triode. The $1/\omega C$ portions of the resonance curves are seen to be almost identical, indicating that the cable and tube input capacitances in this case are much higher than the distributed capacitance in the head windings. (The 150 H head had 27000 turns of No. 50 wire and had an impedance at resonance of 6.7 Megohm. The dc resistance of the winding was 13.8 kilohm.)

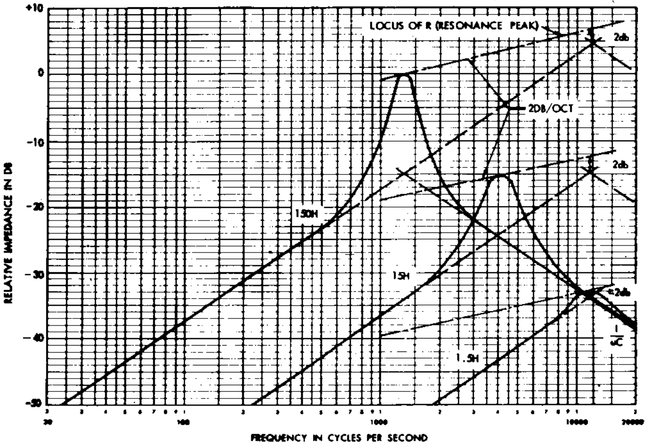

FIG. 14. Measured relative impedance of three different heads made on identical cores but with the inductances 1.5 H, 15 H and 150 H respectively.

Multitrack heads in this inductance range are not easily manufactured since the limited winding space will demand the use of extremely delicate wire. A more practical solution is the use of a low-impedance head in conjunction with step-up transformers. The inherent added noise from a transformer should preferably be kept an order of magnitude lower than the head noise.

By drawing 2 db/oct slope lines through the resonance peaks in Fig. 14, it is evident that all three heads have similar Q's at any given frequency (say 12 kc), indicating identical iron losses. These identical iron losses will produce different absolute values of R due to the difference in the number of turns n on the heads. Since R is proportional to n^2 and the number of turns on the three heads has the ratio $1:\surd(10):10$, the respective R's must have the ratio 1:10:100 or 0 db:20 db:40 db as verified in Fig. 14. Since a large R means a high signal-to-noise ratio (see Fig. 12), it is also evident that the amplifier input resistance shall preferably be an order of magnitude higher than R, and R_{eq} (the equivalent noise resistance in series with the amplifier input) as low as possible. For the 150 H head, R_{eq} should preferably be below the dc resistance of the head ($r = 13.8$ kilohm); this will make the thermal head noise the only significant noise contributor up to the frequency where head noise equals amplifier noise. (This frequency is normally above the audible range.)

Figure 14 shows that when the capacitance across the head winding is reduced and the resonance frequency increases, R increases at a 2 db/oct rate. The signal-to-noise ratio, therefore, increases at a 1 db/oct (3.3 db/decade) rate as shown before. Every reduction by a factor of two of the capacitance will then increase the signal-to-noise ratio by 0.5 db. Optimum signal-to-noise ratio is therefore obtained for the following conditions: 1. Minimum head capacitance; 2. Maximum resonant impedance R of the head; 3. Minimum dc resistance r of the head winding; 4. Minimum equivalent input noise resistance R_{eq} of the amplifier; 5. Maximum input resistance of the amplifier.

Shot Noise

Shot noise is found in tubes and transistors. As the dc current through a tube or a transistor is carried by particles with finite velocities (electrons in tubes and electrons and holes in transistors), rather than by a continuous medium, a small ac noise current will be superimposed upon the dc plate and collector current. In the case of a tube with noninteracting electron transits, the noise current at frequencies low compared with the reciprocal of electron transit time is:[7]

$$I_N = \Gamma\sqrt{(2iedf)} \quad \text{or}$$
$$I_{N,rms} = \Gamma \times 1.78 \times 10^{-5}\sqrt{(idf)},$$

where Γ is a smoothing factor due to space charge, e is the electron charge 1.59×10^{-19} coulomb, I_N is noise current in μA, i is the dc current in mA and df the bandwidth in cps. The formula points out that for constant bandwidth, lower noise will result from lower plate and collector current and lower Γ (more space charge). The latter can be achieved by lowering the plate voltage at constant cathode temperature.

Shot noise *current* divided by the transconductance of a tube will give the equivalent noise *voltage* at the grid, and since the energy per cps of bandwidth is constant, this noise is similar to thermal noise and can be expressed as the thermal noise of an "equivalent" noise resistor at room temperature, thereby eliminating the bandwidth specification.

At 27°C room temperature and 1000°K cathode temperature, the equivalent noise resistor as given by different authors averages:[8]

$$R_{eq} \sim 3/g_m \quad (R_{eq} \text{ in ohms and } g_m \text{ in mhos}).$$

The lowest equivalent noise resistance is thus obtained with a tube operated at the highest possible transconductance, and a low noise tube is therefore a tube with high transconductance at low plate current.

Since the noise current in a transistor can be expressed as $I_N = \sqrt{(2iedf)}$, it is also called "shot noise." This type of noise is produced when the dc current passes the emitter and collector junctions. Today's transistors, especially the silicon types, are quite competitive with tubes when it comes to thermal, shot and $1/f$ noise, and the price seems to come down at a fairly rapid rate too. Betas of 100 to 1000 are common, making it possible to obtain high input impedance $[\beta \times (R_E + r_e)]$ and high voltage amplification simultaneously; therefore, the coupling capacitors between stages can be reliable inexpensive nonelectrolytic (e.g., paper or mylar) types.

The following example will illustrate typical shot noise frequency distributions for a Nuvistor tube and a silicon transistor.

Example 6: **Measure the noise voltage of a tube** (Nuvistor 7895) and transistor (Fairchild S 3568, a version of 2N 2484) *vs* frequency with the following source impedances: $R_s = 10$ ohm, $R_s = 1$ kilohm and $R_s = 51$ kilohm. The results are shown in Fig. 15. $R_s = 10$ ohm can be considered a short circuit; since the $R_s = 1$ kilohm curves at mid-frequencies are approximately 3 db above the $R_s = 10$ ohm curves, R_{eq} must also be approximately 1 kilohm for both tube and transistor at these frequencies. Even the low frequency ($1/f$) noise is comparable within a couple of db. The tube appears to be superior at low frequencies and the transistor at high frequencies for $R_s = 51$ kilohm, but the difference is small.

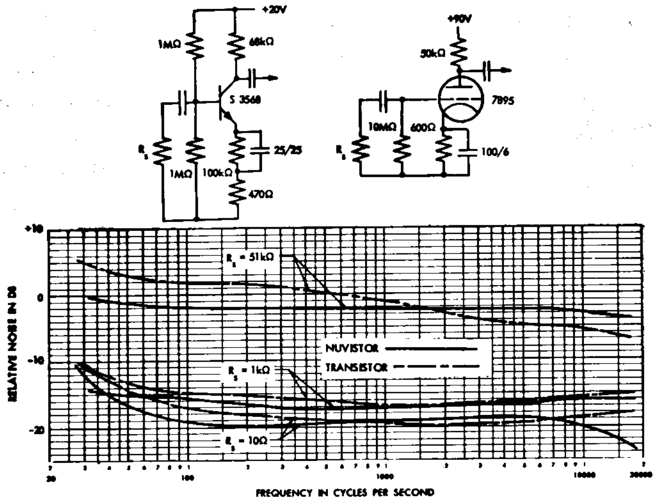

FIG. 15. Measured tube and transistor noise (and noise from the source resistance) *vs* frequency with source resistance as parameter.

In order to find the maximum signal-to-noise ratio, the output levels which produced 10% second harmonic distortion were also measured: they were 2.2 V_{rms} for the transistor and 33 V_{rms} for the Nuvistor. This considerable difference (24 db) in maximum signal-to-noise ratio is, however, significant only in certain circuit configurations which have to accept a wide range of input signals, such as microphone preamplifiers, or where a passive attenuation is found between stages (volume control, filters, etc.). In a tape-reproducer preamplifier with post-equalization provided by negative feedback, no such attenuation is necessary and tube and transistor amplifiers perform identically.

Since many silicon transistors have a maximum collector to emitter voltage, $V_{CE} = 60$ V (some even 150 V) it is possible to get maximum "undistorted" output levels comparable to those of tubes, if low noise is not the primary design goal.

A semiconductor device which in many respects performs like a tube is the field effect transistor (FET). Its input impedance is inherently in the Megohm range, but the noise factor is low at high source resistances only.

Example 7: Calculate the R_{eq} for a FET with a 1.5 db noise figure when $R_8 = 200$ kilohm.

Source resistance noise: $e_s^2 = 4kTBR_s$
Equivalent noise resistance noise: $e_R^2 = 4kTBR_{eq}$
Total noise: $e_{tot}^2 = e_s^2 + e_R^2 = 4kTB\,(R_s + R_{eq})$
Noise factor $(F) = e_{tot}^2/e_s^2 = 1 + R_{eq}/R_s$
Noise figure $(NF) = 10 \log F = 1.5$ db
$F = 1{,}414 = 1 + R_{eq}/R_s$ or
$R_{eq} = 100$ kilohm (compared to 1 kilohm for a tube or regular transistor).

1/*f* Noise

The term "1/*f* noise" means that the rms noise power in a band of constant absolute width is inversely proportional to frequency. The noise voltage is then proportional to $\sqrt{(1/f)}$, i.e., it has a 3 db/oct slope on this basis. Such 1/*f* noise is found in resistors, tubes and transistors. There are two different types of 1/*f* noise: one is "flicker" noise from the fluctuation of thermionic emission from oxide coated cathode in tubes, which is reduced by space charge; the other may be termed "semiconductor noise" since it can be interpreted[9] as fluctuations of conductivity in a semiconductor (fluctuations in the number of carriers in the conduction band). "Excess noise" (over thermal) and "current noise" are other expressions for 1/*f* noise. This noise is found in a variety of materials:[10] carbon granules, carbon filaments, graphite powder, pyrolytic carbon films, germanium filaments, thin metal films, lead sulphide films, all types of rectifying barriers in germanium, silicon and metallic oxides, single crystal cuprous oxide and the interface layer (barium orthosilicate) in tubes between the nickel cathode tube and its oxide coating.

1/*f* Noise in Tubes

It is especially important to have the orthosilicate interface layer in tubes under control, as it may change with time. A tube with low initial 1/*f* noise can grow a considerable "interface" layer with time, unless the purity of the nickel is controlled when the tube is manufactured. The interface layer is produced by the impurity (silicon) in the nickel cathode which, by diffusion, combines with the oxide coating and forms orthosilicate. Since this interface layer exhibits a parallel connection of a capacitance and a resistance in series with the cathode lead, it is important that it be eliminated if the tube is to be used at rf frequencies. Thus, this type of improvement in tubes for rf and computer use will also benefit their use at audio frequencies. Tubes like 6SN7 and 12AU7 have been tested for interface 1/*f* noise, and a 40 db increase in equivalent noise at 30 cps was measured after several hundreds hours of use.[11] It was distinguished from flicker 1/*f* noise by its closer dependence upon the square of the plate current, whereas flicker noise is proportional initially to the square of the emission current. A test of aged tubes at different cathode temperatures gave the following results:[11]

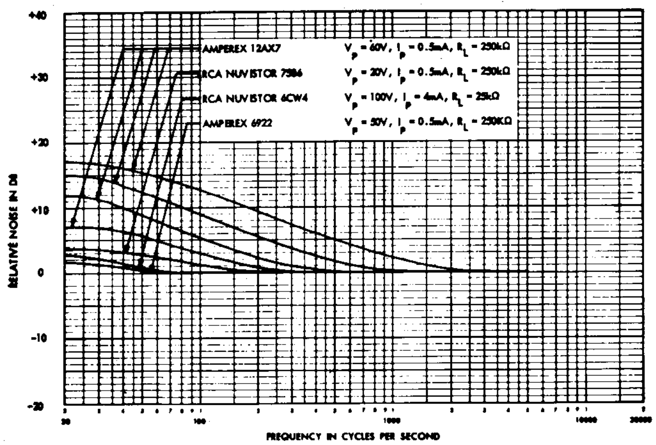

FIG. 16. Noise frequency response of different tubes showing different amounts of 1/*f* noise.

Filament voltage	6 V	16 V	18 V
Interface resistance	5000 ohm	40 ohm	0.5 ohm
Noise resistance (30 cps)	7×10^{10} ohm	6.3×10^{6} ohm	No value given

Another illustration of the difference between tube types with respect to 1/*f* noise is displayed in Fig. 16 which shows that approximately 16 db lower noise can be obtained at 30 cps with a tube with controlled interface, such as 6922 or 7586, as compared to a standard audio tube such as 12AX7. Rf tubes with controlled interface are not necessarily good audio tubes, however. Some are very microphonic at frequencies which are difficult to damp, and since rf tubes have relatively small mechanical structures in order to keep the inter-electrode capacitance down the cathode temperature must be rather high for a given maximum plate current, with reduced life as a result. The RCA Nuvistors have controlled interface and low cathode temperature; because of the small size, the microphonics are at a high frequency (3-5 kc) which is fairly easy to damp; so far, the Nuvistors have been found to be an excellent type of low noise tube.

1/*f* Noise in Transistors and Resistors

Bell[12] defines metallic conductors and semiconductors as follows: "In a metallic conductor the number of conduction electrons remains constant and therefore temperature variation of conductivity is a function only of their mobility. In a semiconductor the charge carriers (electrons and holes) are not permanently in the conduction band, but are raised to it from their ground states by thermal or other excitation."

More practically measurable properties of semiconductors are negative temperature coefficient and 1/*f* noise.

The 1/*f* noise in transistors is due[13] to leakage current and contaminated or imperfect semiconductor surfaces and can be reduced by proper manufacturing until the crossover frequency with shot noise is below 100 cps even at low collector currents. (See Fig. 15.)

The $1/f$ noise in resistors will occur when current passes through them. The National Bureau of Standards has recommended as a noise index that the unit μV noise per V dc in a frequency decade in db be used as a measure of $1/f$ noise in a resistor.

A resistor has a 0 db noise index when 1 μV noise is measured in a frequency decade with 1 V dc applied across it. The resistor value is immaterial. If 2 V dc is applied, 2 μV is produced in a decade if the noise index is 0 db. A noise index of -6 db results in ½ μV noise in a decade for 1 V dc applied voltage.

Different resistor types have different noise indexes. Resistor noise measurements using the Quan-Tech Laboratories, Inc., Model 315 Resistor Noise Test Set gave the following results: -30 to -45 db for good metal film resistors and $+3$ to -30 db for carbon film and composition resistors. All three types showed highest noise index for the highest value resistors. The low value resistors are made of a material with properties close to those of the bulk material, whereas the high value resistors, for example film resistors, rely on the resistance of extremely thin conducting layers with properties several orders of magnitude different from those of the bulk material. One brand of metal film resistors had the following distribution:

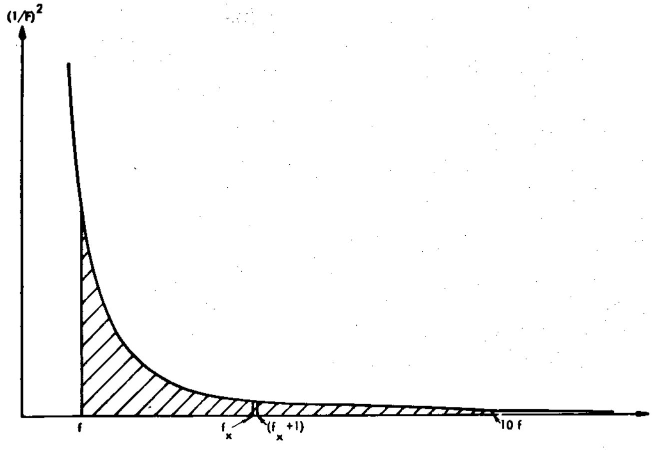

FIG. 17. The curve shows the square of $1/f$ noise voltage *vs* frequency, using linear scales. The crosshatched area represents the total $(1/f \text{ noise})^2$ in a decade. The area from f_x cps to f_{x+1} cps represents the $(1/f \text{ noise})^2$ in a one cps wide band.

47 kilohm	-45 db
100 kilohm	-33 db
825 kilohm	-30 db

Ten resistors of each value were measured.

Relationship Between Noise Index of a Resistor and the Crossover Frequency Between Thermal Noise and 1/f Noise

In this section will be calculated the total $1/f$ noise in a decade produced by a resistor R, the average 1 cps bandwidth noise in a decade and the frequency at which it occurs. The frequency f_x where it equals the 1 cps bandwidth thermal noise will be found and an expression for the relationship between resistance value R, crossover frequency f_x, noise index and applied dc voltage will be derived.

Figure 17 shows the square of noise voltage *vs* frequency of $1/f$ noise using linear scales. Since noise must be added rms, $(e_{noise})^2$ in a decade (f to $10f$) is:

$$\begin{aligned}\int_f^{10f} C\times 1/f\times df &= C(\ln 10f-\ln f)\\ &= C\times\ln(10f/f)\\ &= C\ln 10.\end{aligned}$$

The average $(\textit{noise voltage})^2$ per cps is thus:

$$C\ln 10/(10f-f)=C\ln 10/9f,$$

where f is the lowest frequency in the decade measured, and C is a constant.

The $(\textit{noise voltage})^2$ in a one cps frequency band from f_x to (f_x+1) cps is

$$C\int_{f_x}^{f_x+1}(1/f)df=C\times\ln[(f_x+1)/f_x].$$

To find the frequency f_x at which this one cps bandwidth $(\textit{noise voltage})^2$ equals the average $(\textit{noise voltage})^2$ in a decade:

$$(C\ln 10)/9f=C\ln[(f_x+1)/f_x]$$

or, multiplying with the factor $\log N/\ln N = 1/2.3026$ and eliminating C:

$$\log 10/9f=1/9f=\log[(f_x+1)/f_x]=\log[1+(1/f_x)]. \quad (1)$$

If the audio range only is considered, f will vary between 20 cps and 2 kc (20 kc will be the highest frequency in the decade). In order to solve this equation with known accuracy in the audio frequency spectrum, the two limiting frequencies $f=20$ cps and 2 kc are inserted in Eq (1):

$$\begin{aligned}&\log(1+1/f_x)=1/9f=1/180=0.00555\\ &1+1/f_x=1.012874 \text{ or } f_x=77.68 \text{ cps}\\ &\qquad\qquad =77.68/20=3.88f\\ &\log(1+1/f_x)=1/9f=1/18000=0.0000555\\ &1+1/f_x=1.0001278 \text{ or } f_x=7825 \text{ cps or}\\ &\qquad\qquad 7825/2000=3.91f.\end{aligned}$$

The difference is seen to be less than 1%.

The average one cps bandwidth $1/f$ $(\textit{noise voltage})^2$ is:

$$\begin{aligned}(C\ln 10)/9f&=(C\ln 10)/(9f_x/3.9)\\ &=(C/f_x)(2.3026\times 3.9/9)\\ &=.998\times(C/f_x)\sim C/f_x.\end{aligned}$$

The average one cps bandwidth thermal $(\textit{noise voltage})^2$ is $4kTR$. The average one cps bandwidth thermal and $1/f$ $(\textit{noise voltage})^2$ are equal when

$$4kTR=C/f_x \text{ or } Rf_x=C/4kT.$$

The value of C is found from

$$e_1/f_{\text{decade}}^2=C\ln 10=(1\ \mu\text{V}/\text{Vdc})^2 \qquad \text{(0 db noise index)}$$

$$C=(1/10^{12}\ln 10)[V^2/(\mathrm{Vdc})^2]$$

$$Rf_x=10^{23}/(4\times 1.374\times 300\times 10^{12}\times \ln 10) \qquad (2)$$

$$=26.4\times 10^6. \qquad \text{(0 db noise index, 1 V dc)}$$

The thermal noise voltage across resistors from 1 ohm to 100 Megohm is calculated from $e=\sqrt{(4kTBR)}$. The bandwidth is chosen to be 1 cps and the voltage is expressed in db below 1 V. Since thermal noise has a flat frequency response, the resistor thermal noises are plotted as horizontal lines, as shown in Fig. 18. $R=1$ Megohm inserted in resistor to be —178 db. The —40 db noise index, 1 V dc curve crosses —178 db at 26.4 cps, so the dc voltage must be kept below 1 V.

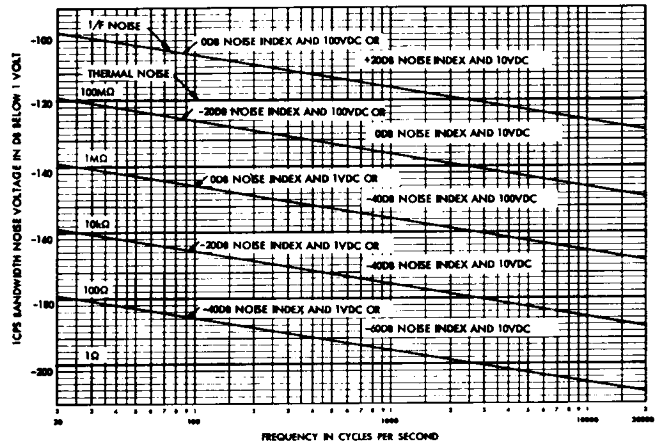

FIG. 18. **Relationship** between thermal noise and 1/*f* noise in a resistor at audio frequency, with noise index and applied dc voltage as parameters.

Eq. (2) yields a crossover frequency f_x between thermal and 1/*f* noise of 26.4 cps. The 1/*f* frequency response for 0 db noise index and 1 V applied dc is then obtained by drawing a 3 db/oct sloping line through the crossover point. The 1/*f* noise curves for different noise indexes and applied dc voltages are obtained by drawing lines parallel to the 3 db/oct sloping line. If 10 times the dc voltage is applied, the noise voltage increases 10 times or 20 db; if the noise index is 20 db lower the noise curve will move down 20 db. The data given by Stansbury[14] also prove the validity of Fig. 18. The 1/*f* *noise*, unlike thermal noise, is independent of resistor value. This is due to the fact that the noise is proportoinal to the dc *current* through a resistor (current noise).

The following two examples may illustrate the magnitude of 1/*f* noise in resistors:

Example 8: What is the noise voltage increase near 30 cps of a 1 W 10 kilohm carbon film resistor with a noise index of 0 db when 1 W dc power is dissipated in it?

$V=\sqrt{(RW)}=\sqrt{(10^4\times 1)}=100$ V. Figure 18 shows a 1/*f* noise at 30 cps for 0 db noise index and 100 V dc of —101 db. The thermal noise of a 10 kilohm resistor is —158 db (Fig. 18) so the increase is 57 db!

Example 9: How much dc can be applied across a 100 ohm metal film resistor with a —40 db noise index if 1/*f* noise shall be negligible in the audible frequency range (30-15000 cps)?

The 1/*f* noise is negligible when it is below the thermal noise. Figure 18 shows the thermal noise of a 100 ohm resistor to be —178 db. The —40 db noise index, 1 V dc curve crosses —178 db at 26.4 cps, so the dc voltage must be kept below 1 V.

The curves for thermal noise and 1/*f* *noise* in Fig. 18 are plotted for 1 cps bandwidth for reasons of simplicity. Other bandwidths, for example, $B=10$ cps, 5 cps or 1/10 cps, could just as well be used; in this case the relationship between thermal and 1/*f* noise would not change, but the whole system of curves would move up $20\log\sqrt{(B)}$ db.

When a pure ac signal is applied across a non-wirewound resistor, 1/*f* noise will be produced proportionally to the amplitude of the signal and a process similar to modulation noise can be observed.

Induced Grid Noise

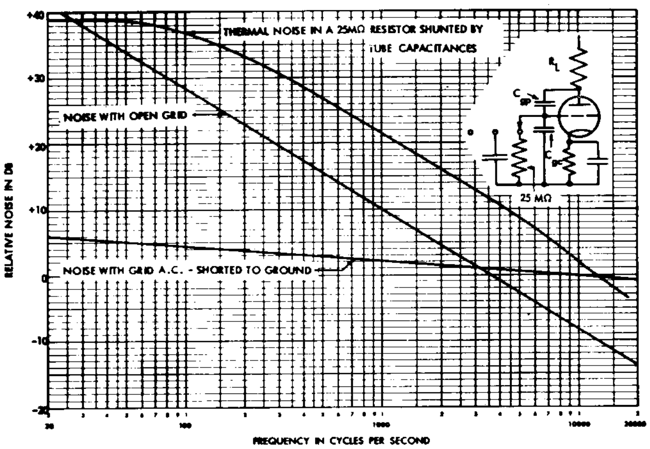

FIG. 19. **Measured** noise frequency responses for the determination of induced grid noise.

Induced grid noise is found in tubes. When the electrons emitted from the cathode pass through the grid wires on their way to the plate, they create small current pulses as they approach and move away from the grid wires. If the grid is shorted to ground, no grid noise voltage will develop. If there is a finite impedance between grid and ground, the noise voltage will be the noise current times this impedance. Maximum noise is obtained with the highest impedance, which means that a tube with an open grid would exhibit maximum induced grid noise; in this case, the noise is attenuated only by the tube capacitances.

Since the same electrons which pass the grid will eventually hit the plate, a certain correlation between induced grid noise and shot noise seems to exist. At rf frequencies the electron transit time becomes significant. Van der Ziel and others have found a noise minimum at these frequencies by detuning a resonant circuit slightly so as to obtain a certain phase relationship between the grid and the plate. At audio frequencies, however, little use can be made of this transit time phenomenon. The induced grid noise is approximately equal to the thermal noise arising from the grid-cathode conductance at a temperature 1¼ times that of the cathode, according to Bakker[15] and others, and can therefore, at audio frequencies, be considered as having the same noise spectrum as the thermal noise in a resistor at room temperature.

Figure 19 shows a test circuit used to determine induced grid noise. The thermal noise frequency response is modified by the interelectrode capacitances. C_{gp} is the "cold" plate-grid capacitance, and C_{gc} is the "cold" grid-cathode capacitance. If C_{gc} were infinite (shunted by a large external capacitor) the only noise would be shot noise (and $1/f$ noise) and the tube would have full amplification. If C_{gc} is not large compared with C_{gp} feedback will result from plate to grid. The amount of feedback is $1-\beta A$, where

$$\beta=\frac{(1/j\omega C_{gc})}{(1/j\omega C_{gc})+(1/j\omega C_{gp})}=\frac{C_{gp}}{C_{gp}+C_{gc}}.$$

Since the input impedance is capacitive, $C_{gc}+(1+A)C_{gp}$, it is possible to measure the equivalent noise resistance of the induced grid noise by measuring the noise frequency response.

An RCA Nuvistor 7586 was connected as shown in Fig. 19. Plate voltage was 30 V and the grid could be switched between a 25 Megohm resistor to ground, ground directly and open circuit. The cathode potential was monitored to ensure constant dc operating conditions in all three cases.

Figure 19 shows the noise frequency response with open grid, with a 25 Megohm source resistor and with the grid shorted to ground. The open grid and 25 Megohm curves exhibit 6 db/oct slopes with a 12 db level difference. Using the formulas derived in Section 2 of the Thermal Noise section, we find that there is a 24 db difference (16 times) in noise resistance; in other words, the equivalent noise resistance, R_{eq} due to induced grid noise is approximately 16×25 Megohm $= 400$ Megohm. It is also interesting to note that due to the shorting out of the negative feedback from the tube capacitance, the noise at 20 kc *increased* 11 db when the floating grid was shorted to ground.

Barkhausen Noise

Magnetic materials with crystalline structure, whether iron, nickel, cobalt or others, consist of small units called "domains." Each domain is a small magnet, magnetized to saturation in one direction. The domains are separated by walls which provide the shift in direction of magnetization. If the material is "unmagnetized," the sizes of the differently oriented domains are distributed evenly and form closed magnetic flux paths inside the material. If an external field is applied, the following will happen, according to D. A. Bell:[16] *1.* The domains oriented approximately parallel to the outside magnetizing field will grow in size, moving the walls outwards at the expense of the domains oriented in other directions. *2.* This growth is fast and reversible until it meets a discontinuity, an impurity in the material or an irregularity in the crystalline structure. The wall movement will stop at this place until enough energy is present to overcome the disturbance. The resulting sudden jump in magnetization when the disturbance is overcome, damped only by eddy currents in the material, constitutes the Barkhausen noise. Once a Barkhausen jump has occurred the situation is irreversible for small signal levels at this particular place in the crystal. This hysteresis effect is negligible at very low signal levels where there is complete reversibility. *3.* All the domains are now oriented in the same direction, but not quite in the direction of the applied field, because they are held back by the crystalline structure (anisotropy). The last domain movement possible is that of bringing their orientation in exact accordance with the applied field. No further Barkhausen jumps are possible, and the process is reversible and noise-free.

Example 10. What is the order of magnitude of induction in a tape reproducing head?

A full track ¼-in. tape recorded to saturation may supply a flux of 600-800 milliMaxwell. With a head efficiency of 75%, about 500 mMx will go through the major volume of the laminations. If the cross-sectional area of the pole pieces is 0.2 cm^2, the induction is 0.5 Maxwells/0.2 cm^2 = 2.5 Gauss. Small local volumes near the pole tips will have a higher flux line concentration, but the major volume, exhibiting the major number of possible Barkhausen jumps, will have a flux concentration of approximately 2.5 Gauss. The saturation induction of a good head material is 6000 to 7000 Gauss, depending upon lamination thickness. Since even the induction from a saturated tape is very small compared to this head material saturation induction, the Barkhausen noise is very low, even at tape saturation. At tape noise levels it will be practically nonexistent.

NOISE IN A REPRODUCING SYSTEM

The different noise sources mentioned in the previous sections were considered in the design of the head and preamplifier of the Ampex MR-70 master recorder. Nuvistors were preferred over transistors because of their inherently high input impedance and low equivalent input noise resistance, and also because of their high reliability and low $1/f$ noise. The MR-70 plays ¼-in., ½-in. and 1-in. tapes, and a multitude of head combinations can be used. The heads are all low-impedance, approximately 17 mH, and a step-up transformer is used in the amplifier. The eddy current losses and copper losses in the head depend

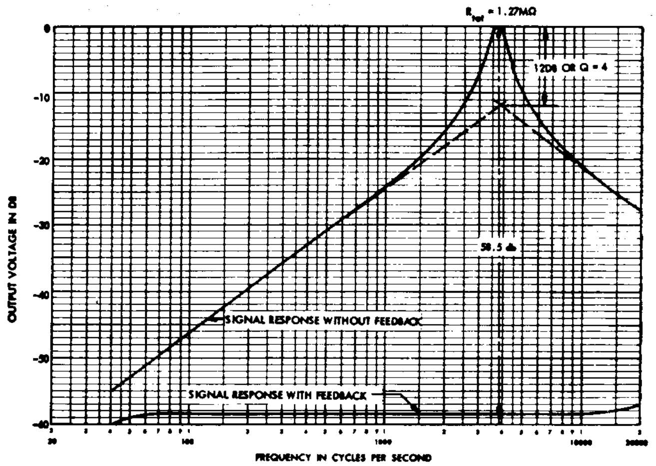

FIG. 20. Signal output voltage *vs* frequency with and without negative feedback for the MR-70 playback head.

upon the number and type of laminations, the number of turns on the head and the winding space.

The following data for a ¼-in. full track head will give a quantitative picture of the noise contributions in the MR-70 playback system:

Head: $L = 17$ mH, dc resistance $r = 1$ ohm, equivalent parallel resistance $R = 1500$ ohm when resonated at 3.7 kc with a low loss capacitor.

Transformer: $L_{sec} = 80$ H, $r_{pri} = 0.75$ ohm, $r_{sec} = 850$ ohm $R_{sec} = 22$ Megohm at 2 kc; turns ratio 1:30.

The total dc resistance r_{tot} of the head and transformer combination, referred to the secondary, is then $30^2 \times (1 + 0.75) + 850 = 2425$ ohm. The total parallel resistance R_{tot} of the head and transformer combination, referred to the secondary, is $30^2 \times 1500 = 1.35$ Megohm in parallel with 22 Megohm = 1.27 Megohm.

The difference in noise voltage between r_{tot} and R_{tot} should therefore be $\sqrt{[(1.27 \times 10^6)/2425]} = 22.9$ or 27 db, which is confirmed by measurement (see Fig. 21). The total inductance of the head and transformer combination referred to the secondary is: $30^2 \times 17\text{ mH} = 15.3$ H in parallel with 80 H = 12.9 H. The Q at head resonance is

$$R_{tot}/\omega_0 L = (1.27 \times 10^6)/(2\pi \times 3700 \times 12.9) = 4.2,$$

which is confirmed in Fig. 20. The equivalent input noise resistance R_{eq} is 800 to 1000 ohm as shown by the dotted line in Fig. 21, which falls below the head noise curve. The measured signal curve (Fig. 20) and noise curve (Fig. 21) are seen to be similar in shape except at low frequencies; this results in a constant signal-to-noise ratio at mid- and high-frequencies but a reduced one at low-frequencies. The low-frequency deterioration is due to the dc resistance r_{tot} or the equivalent input noise resistance R_{eq}, whichever is larger. Figure 21 shows the dc resistance r_{tot} to be the largest in the case of the MR-70, so that the amplifier noise is masked by thermal head noise from dc to 20 kc (assuming amplifier $1/f$ noise never rears its ugly head).

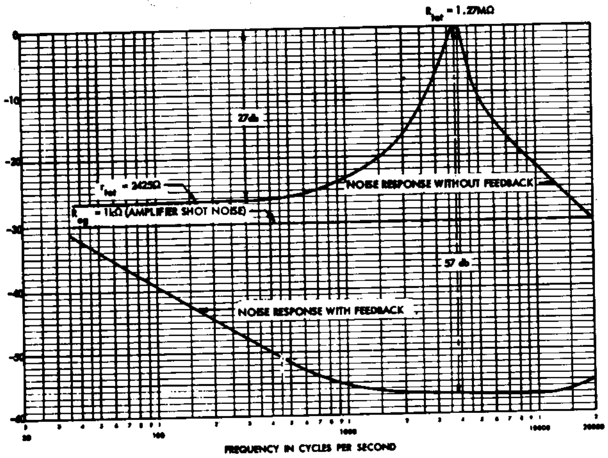

FIG. 21. Noise output voltage *vs* frequency with and without negative feedback for the MR-70 playback head.

The curves in Fig. 20 are measured with constant flux in the head. This is not a practical condition because, at economical tape speeds (e.g., 7½ and 15 ips), the tape flux is constant at low- and mid-frequencies only, and decreases at high frequencies. This flux loss at low speeds is compensated for by increasing the high frequency amplification in the recording as well as the reproducing amplifiers, thereby introducing a deterioration of both the signal handling capabilities and the noise level. A high tape speed (30 ips) is available in the MR-70 to minimize this deterioration.

When the flux in the head is constant, its time derivative ($d\phi/dt$) will rise 6 db/ oct with frequency (see Fig. 22).

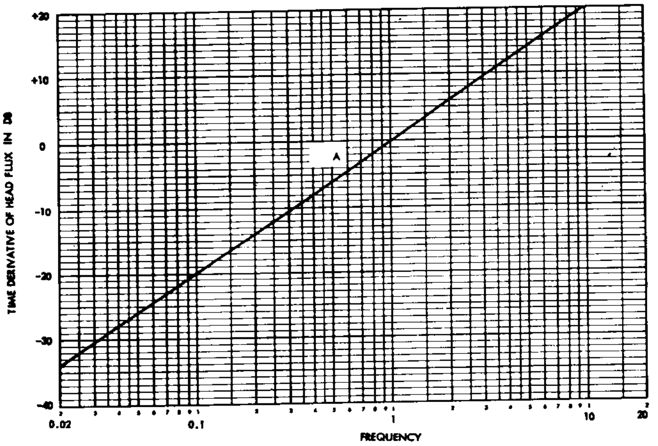

FIG. 22. Time derivative ($d\phi/dt$) of the (constant) head flux *vs* frequency.

The head circuit shown in Fig. 6, which actually constitutes a low-pass filter (its response is shown in Fig. 23), will modify this 6 db/oct curve. Figure 24 shows the resulting voltage output from the head with the parallel resistance R as parameter.

With low values of R it is possible to obtain a flat output voltage frequency response in a desired bandwidth for constant flux in the head. A low value of R, however, will drastically reduce the signal-to-noise ratio, as described above in the Thermal Noise Section and displayed in Fig. 12. By using negative feedback, however, a "noise free"

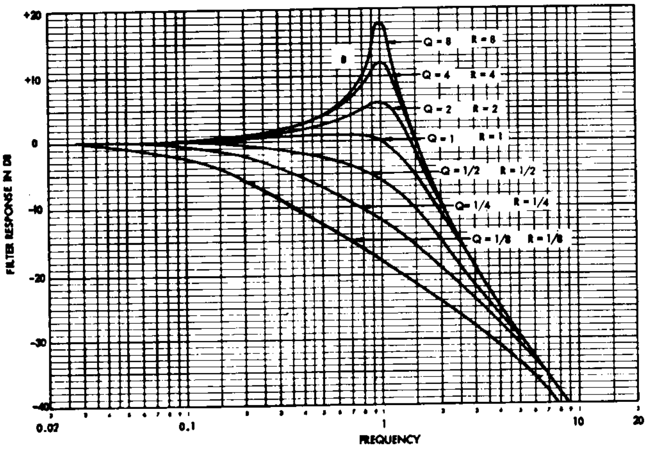

FIG. 23. Filter response of the electrical head circuit shown in Fig. 6 with R as parameter.

resistance can be generated to provide the necessary damping without adding significantly to the noise.

A three-stage amplifier is used in the MR-70 (grounded cathode, grounded plate, grounded grid) with low phase shift. Total amplification is 62 db. A 2.7 Megohm feedback resistor is connected from the plate of the last stage directly to the effective head (secondary of input transformer), resulting in 58.5 db damping at the head resonance frequency. Because the head resonance curve has long 6 db/oct slopes with constant phase shift and the amplifier has wideband response, there is no danger of instability.

Negative feedback is also used in the disc recording field where moving coil cutting heads, mechanically resonating near the middle of the audio frequency band, are equalized to a flat velocity frequency response.

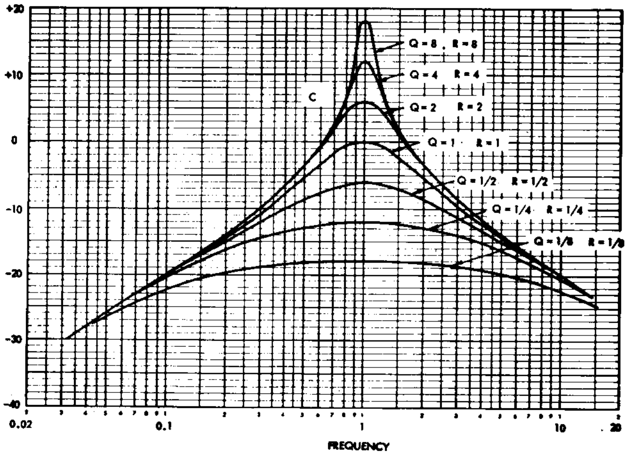

FIG. 24. Resulting head output voltage, with R as parameter, for the head circuit shown in Fig. 6.

Because of the resulting flat voltage frequency response from the head, there is no danger of excessive CCIF intermodulation distortion[17] which results if a passive equalization follows a source of distortion (tube or transistor).

REFERENCES

1. S. J. Begun, *Magnetic Recording,* (Rinehart Books, Inc., New York, 1951), Chapter 1, pp. 4, 7, 10.

2. When the noise power is divided by the bandwidth, the result is approximately equal to the "spectral noise density." Spectral noise density is defined as the noise per cps of bandwidth, as measured in an infinitesimally small bandwidth ΔB; that is, *spectral noise density* = (*noise within* ΔB)/ΔB. The term "spectral noise density" is not used here for two reasons: *1.* Most of the data was taken in 6 cps wide bands, and not reduced to the "per cps" form; and 2, the term "noise frequency response" seems more descriptive for the purposes of this paper.

3. F. E. Terman and T. M. Pettit, *Electronic Measurements,* (McGraw-Hill Book Co., Inc., New York, 1952), Second Edition, Sec. 812, pp. 354-355.

4. This excludes parametric amplifiers and magnetic amplifiers, for instance.

5. E. G. Nielsen, "Behavior of Noise Figure in Junction Transistors," *Proc. IRE,* 957 (1957).

6. D. A. Bell, *Electrical Noise,* (D. Van Nostrand Co., Inc., New York, 1960), Chapter 11, p. 247.

7. *Ibid.*, Chapter 5, pp. 96, 130.

8. *Ibid.*, Chapter 8, p. 151.

9. *Ibid.*, Chapter 10, p. 210.

10. *Ibid.*, Chapter 10, p. 212.

11. *Ibid.*, Chapter 8, pp. 163-164.

Since the feedback is applied directly to the head, the feedback resistor will load the head slightly and add thermal noise. With a 2.7 Megohm feedback resistor and a head resonant impedance of 1.27 Megohm, the resulting impedance is 860 kilohm. This will reduce the signal by 3 db and the noise by 1.5 db, and result in a net reduction of the signal-to-noise ratio of 1.5 db.

Figures 20 and 21 confirm this. If a regular resistor had been used to achieve the 58.5 db damping, the signal-to-noise would have been reduced by $58.5/2 = 29.3$ db!

ACKNOWLEDGMENT

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APPENDIX

Noise-Producing Real Part of Head Impedance *vs* Frequency

The impedance of the circuit in Fig. 6 is given by

$$Z = \frac{\{[j\omega L(1/j\omega C)]/[j\omega L + (1/j\omega C)]\}R}{\{[j\omega L(1/j\omega C)]/[j\omega L + (1/j\omega C)]\} + R}$$

$$= \frac{j\omega LR}{j\omega L + R[1-(\omega/\omega_0)^2]}$$

$$= \frac{j\omega LR\{R[1-(\omega/\omega_0)^2] - j\omega L\}}{R^2[1-(\omega/\omega_0)^2]^2 + \omega^2 L^2}$$

where $\omega_0{}^2 = 1/LC.$

The real part of Z is:

$$a = \omega^2 L^2 R/\{R^2[1-(\omega/\omega_0)^2]^2 + \omega^2 L^2\}.$$

At resonance ($\omega = \omega_0$): $a = R.$

Above resonance ($\omega >> \omega_0$):

$$a = \frac{\omega^2 L^2 R}{R^2(\omega/\omega_0)^4 + \omega^2 L^2} \sim \frac{L^2 R}{(\omega^2 R^2/\omega_0{}^4) + L^2} \sim \frac{L^2 \omega_0{}^4}{R\omega^2}$$

for $\omega^2 R^2/\omega_0{}^4 >> L^2,$

which represents a 12 db/oct decreasing slope.

12. *Ibid.*, Chapter 8, p. 151.

13. L. Blaser, "The Design of Low Noise, High Input Impedance Amplifiers," *The Solid State Journal,* 21 (July, 1961).

14. A. Stansbury, "Measuring Resistor Current Noise," *Electronic Equipment Engineering,* 17 (June, 1961).

15. C. J. Bakker, "Fluctuations and Electron Inertia," *Physica* 8, 23 (1941).

16. D. A. Bell, *op. cit.*, Chapter 13, p. 279.

17. If two high frequency tones, for example 5000 cps and 5050 cps, of similar amplitude are played back simultaneously through an amplifier stage, a 50 cps difference tone will be produced. If the difference tone amounts to, say 0.2% of the 5000 cps tone, it will then be boosted approximately 30 db by the (50 μsec) passive playback equalization to $31.6 \times 0.2 \sim 6\%$, which is a considerable increase in distortion. Program material of dominant percussive content will tend to sound muddy under these conditions because of the boosted low-frequency distortion products.

The values $R = 130$ kilohm, $f_0 = 50$ kc, $f = 200$ kc and $L = 0.5$ H from Fig. 13 yield

$$\omega^2 R^2/\omega_0^4 \sim 2.7 >> L^2 = 0.25.$$

Below resonance ($\omega << \omega_0$):

$$a = \frac{\omega^2 L^2 R}{R^2 + \omega^2 L^2} \sim \frac{1}{(R/\omega^2 L^2) + (1/R)} \sim \frac{\omega^2 L^2}{R}$$

for

$$R/\omega^2 L^2 >> 1/R,$$

which represents a 12 db/oct increasing slope.

The values $R = 130$ kilohm, $f = 10$ kc and $L = 0.5$ H from Fig. 13 yield:

$$R/\omega^2 L^2 = 1.3 \times 10^{-4} >> 1/R = 7.7 \times 10^{-6}.$$

THE AUTHOR

Erling P. Skov was graduated from the Technical University of Denmark in 1951, receiving his M.S. degree in electrical engineering. From 1951-1956 he was a research and development engineer at Linnet and Laursen, Inc., in Copenhagen. In 1956 he joined the engineering staff of Fairchild Recording Equipment Corporation in New York, where he worked on stereo disk cutting systems, stereo playback cartridges, turntables and associated amplifiers and filters. He became engineer-in-charge of professional products engineering at Fairchild in July 1960, but joined Ampex Corporation in December of the same year. He designed the Ampex FM Multiplex adapter and has mostly worked on the design of low noise tape playback systems and the problems in recording on magnetic sheets. He became staff engineer in 1964 and is currently serving as project engineer on the Ampex spot announcer.

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Mechanical Damping in Tape Transports[*]

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Most of the dynamic elements of a tape transport store energy rather than dissipate it; therefore disturbances due to mechanical imperfections will result in the undesirable condition of oscillations of speed which die out slowly. The various physical mechanisms usable for dissipating (damping) this energy are shown along with a quantitative analysis (using electrical analogs) of their effects.

INTRODUCTION

The length of tape and the rotating elements of a tape transport comprise a mechanical network whose object is to transport the tape from a supply reel, past recording and reproducing transducers (heads), and onto a take-up reel. It is necessary for most applications that the tape speed at the heads be as constant as possible. However, mechanical imperfections of all rotating elements and changes in the power line voltage generate forces and velocities which cause variations of the tape speed. The amount of speed change at the heads due to these forces and velocities depends on the transmission characteristics of this mechanical network of tape lengths and rotating elements.

The energy losses of the mechanical elements are usually small. Because of this, transient disturbances (e.g., tape splices, tape rubbing on a reel, line voltage shifts, etc.) will cause undamped oscillations or "ringing" at the natural resonant frequencies of the network. Additionally, the velocity generated by periodic disturbances (e.g., a "cogging" reel motor, or an eccentric capstan or reel idler, etc.) will be amplified whenever the disturbance frequency falls near a resonant frequency of the mechanical network.

The usual design procedure is to reduce the periodic disturbances by means of increased mechanical accuracy; there are, of course, certain practical limitations to this approach. One can also attempt to place the resonant frequencies so that they do not coincide with the rotational speeds which generate the periodic disturbances. This procedure is valuable but of limited usefulness since the range of possible disturbing frequencies is usually broad, especially for a multi-speed transport. This is illustrated in Fig. 1, using the Ampex Model MR-70 as an example; the rotational speeds and the "cogging" and torque pulses from the reel motors are shown. (Other effects, such as harmonics of rotational speeds and ball-bearing ball speeds may exist, but are not shown.) The rates shown completely cover the

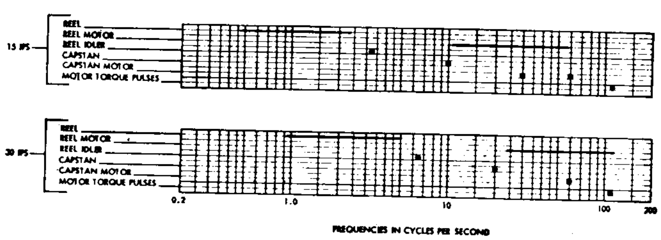

Fig. 1. Frequencies of potential periodic disturbances in the Ampex Model MR-70 tape transport.

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[*] Presented October 17, 1963 at the Fifteenth Annual Fall Convention of the Audio Engineering Society, New York.

range from ¼ to 120 cps. Also, the transient disturbances cannot usually be eliminated, and they will excite any resonant condition.

Reduction of the amplitudes of these resonances is possible by providing energy-dissipating elements, called mechanical responsiveness, and corresponding to electrical resistance. Note that electrical dissipators (resistors) are the simplest and most common electrical element; low-loss (high Q) elements are difficult to design. On the other hand, mechanical elements are commonly high-Q, and a dissipator (responsiveness) is comparatively difficult to design. This paper is concerned with the responsiveness of the elements of the transport and the practical application of additional responsiveness elements.

ANALYSIS OF THE TAPE TRANSPORT

Let us first consider a tape transport; for convenience of analysis, we will simplify as much as possible. Fig. 2a shows a simplified tape transport—supply reel (*1*), tape (*2*), reel idler (turn-around roller) (*3*), tape (*4*), transducer (head) (*5*), tape (*6*), capstan motor system (*7*), tape (*8*), and take-up reel (*9*). This is a network of torsional masses (*1,3,7,9*) a torsional compliance (*7*), and several translational compliances (*2,4,6,8*). In Fig. 2b these have been converted into an equivalent all-translational system and shown as a mechanical schematic drawing, using mechanical symbols; this schematic can be drawn from the physical system by inspection. (Vibrations of the elements—e.g., the head—may also occur, but are ignored in this simplification. Also, in this drawing the damping associated with each element is not shown; it will be considered later.) Figure 2c shows the analogous electrical schematic, drawn by inspection using the mobility analogue. Mass becomes capacitance, compliance becomes inductance, responsiveness (the damping property) becomes resistance, force becomes current, and velocity becomes voltage. The principles of analogy are discussed in detail in the literature.[1]

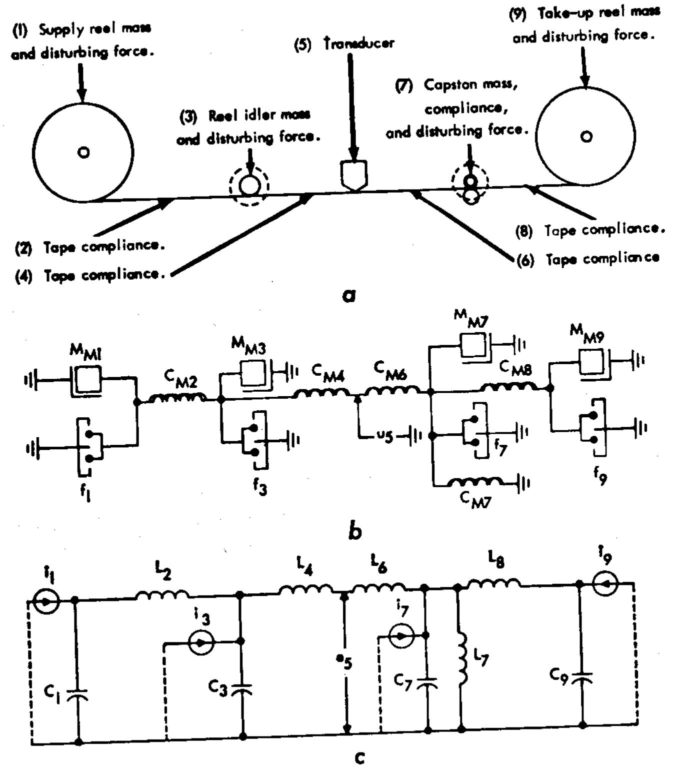

FIG. 2. Representation of a simplified tape transport. *a*. Physical system. *b*. Mechanical schematic diagram. *c*. Analogous electrical schematic diagram.

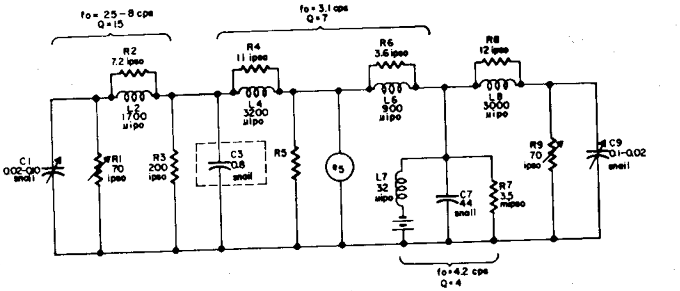

FIG. 3. Electrical analog of an indirect-drive tape transport with ¼ in. wide tape. When the viscous damper of Fig. 15 is used in place of *C3*, this analog represents the Ampex Model MR-70 tape transport.

Quantitative study of any system requires measurement of the mechanical values. Metric units are most easily handled in the analogs, and are recommended. We have used the English inch-ounce-second units and, since our suppliers, machinists, etc., are not familiar with metric units, have invented our own set of unit names since a system of units does not exist for mechanical quantities. We have basic units for force, the ounce; for distance, the inch; and for time, the second. The English foot-pound-second system has a mass unit called a "slug," and we have named the inch-ounce mass unit (dimensionally, oz-sec²/in.) a "snail." (The dictionary says the "slug" is "closely related to the snail," if there *must* be a justification.) It is fairly common to express velocity not as "inches per second," but by its initials, "ips." We have likewise named the compliance unit "ipo" (inches per ounce), and the responsiveness unit "ipso" (inches per second-ounce).

Figure 3 shows a simplified schematic of a tape transport, using a mixture of electrical symbols and the mechanical component values. The greatest mass is that of the capstan, 44 snails (*C7*); the reel idler is much smaller (0.8 snail, *C3*), and the reel still smaller (0.02 to 0.10 snail, *C1*). The drive motor is very stiff (low compliance) (32 μipo, *L7*), and the major compliances are those of the tape itself (1000 to 3000 μipo each, *L2, 4, 6, 8*). Most of the responsiveness elements are too large to have appreciable damping at the frequencies of concern—*C1/L2* with *R1* and *R2* has a Q of approximately 15 at its resonant frequency of 20 cps; *C3/(L4 + L6)* with *R3, 4, 5, 6* has a Q of approximately 7 at its resonant frequency of 3.1 cps; and *C7/L7* with *R7* (which is an "added damping"; to be discussed

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[1] See especially L. L. Beranek, *Acoustics* (McGraw-Hill Book Co., New York, 1st ed., 1954), Chapter 3, Part VI.

below) has a Q of 4 at its resonant frequency of 4.2 cps. Note that the tape speed *per se* does not enter at all into this network.

MEASUREMENTS WITH AN ANALOG ELECTRICAL NETWORK

The mechanical values may easily be converted to the analogous electrical values, as described by Wolf,[2] and the response of this electrical network may then be measured. Such an analog study calls attention to the fact that many variables enter into the performance of a tape transport system. Thus, 1. Disturbances may come from the reeling systems, the reel idler, the heads, and the capstan system; therefore the response from disturbances at each of these inputs should be considered. 2. The speed should be constant at both the recording and the reproducing heads; but the response is different at each head position, requiring separate measurements for the effect at each head. 3. Each element may have disturbances which are of an essentially constant-force nature (e.g., a dragging break or bearing) or of a constant-velocity nature (e.g., an eccentric pulley); responses to velocity and to force inputs are generally different, so that two separate measurements are needed. 4. Many of the elements are variable—e.g., the equivalent mass of the reel of tape changes as the tape is reeled off; the tape compliance changes with different tape thicknesses, widths, and Young's moduli; the compliance of synchronous motors changes with the line voltage; the compliance of a multiple speed synchronous motor is different for each speed.

Therefore the measurements shown in the remainder of this paper represent but a few of many measurements.

RESPONSE OF A TRANSPORT ANALOG WITH LITTLE DAMPING

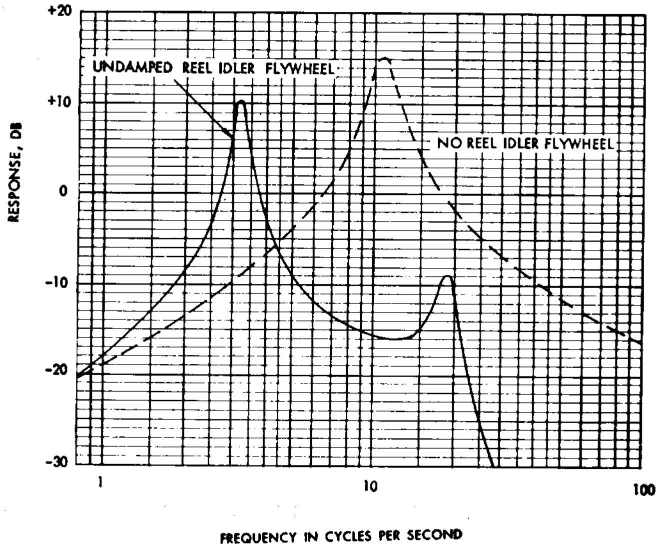

FIG. 4. Frequency response of the analog of the Ampex MR-70 tape transport with ¼ in. wide tape. Velocity at the recording head due to a force at the supply reel; with no reel idler flywheel, and with an undamped reel idler flywheel.

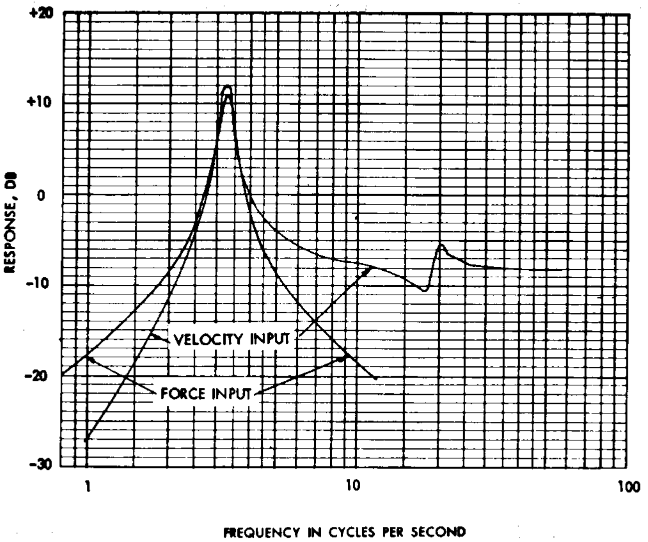

FIG. 5. Frequency response of the analog of the Ampex MR-70 tape transport with ¼ in. wide tape. Velocity at the recording head due to a force and due to velocity at the reel idler, with an undamped idler flywheel.

Let us consider the frequency response of the velocity at the recording head, when a constant force *vs* frequency is applied at the supply reel—e.g., "cogging" of the reel motor. With no reel idler flywheel (Fig. 4, dashed curve) a large

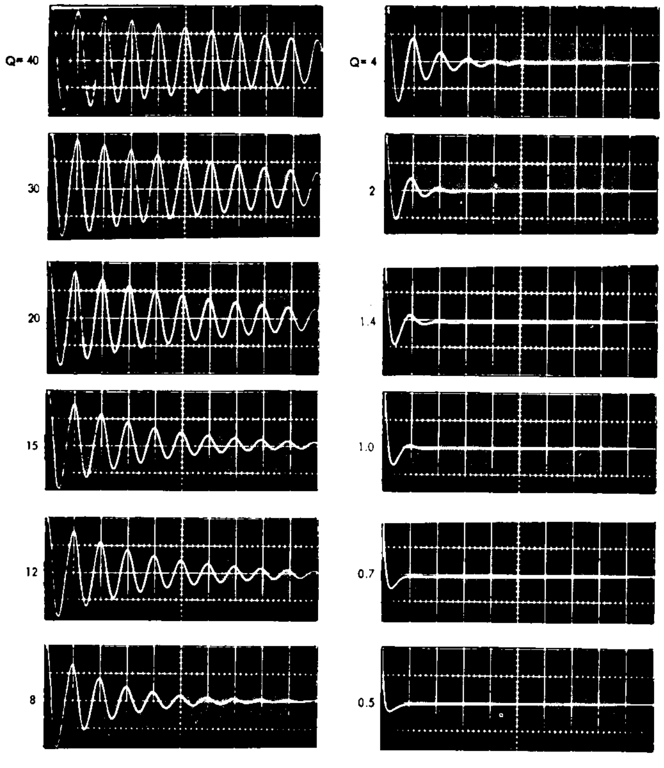

FIG. 6. Oscillograms of damped waves produced by a step voltage change across a resonant circuit, for different values of Q.

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[2] W. Wolf, "An Investigation of Speed Variations in a Magnetic Tape Recorder, with the Aid of Electro-Mechanical Analogies," *J. Audio Eng. Soc.* 119 (1962).

resonant peak occurs at 12 cps as the tape compliance resonates with the equivalent mass of the supply reel. (This resonant frequency varies with the amount of tape on the reel.) Unfortunately the reel motor "slot rate" frequency passes through this range, and will cause high speed variations (i.e., flutter).

Addition of a flywheel (Fig. 4, solid curve) attenuates the response at this frequency, but produces a new resonance around 3 cps. This new peak falls near the rotational frequency of the reel idler itself. A still larger flywheel would further lower the resonant frequency so that it would coincide with the reel rotation frequency.

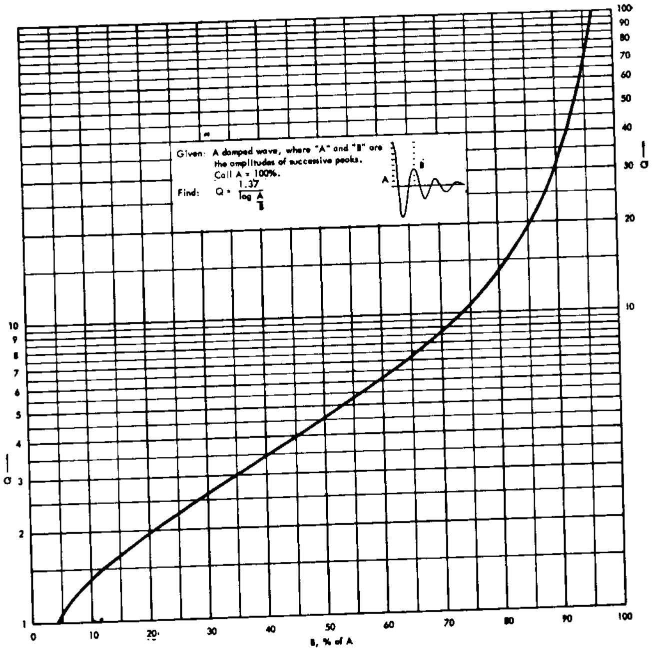

FIG. 7. Q of a damped resonant system as a function of the amplitudes of successive peaks of a damped wave.

Figure 5 shows the response to force and velocity inputs at the reel idler, when an undamped reel idler flywheel is used. The resonance is seen in this analog to amplify the effect of an eccentricity of the reel idler pulley, or a dragging brake or bearing, by four times. (The Q of the actual mechanical system is really 7—our electrical circuit had too much resistance.) When the reel idler rotation frequency coincides with the resonant frequency—as it does with ¼ in. tape at 15 ips—the slightest imperfection of the reel idler will cause large amounts of speed variation (flutter) at 3.2 cps. Moving the frequency is of little help, as some *other* disturbance frequency will then be amplified.

A transient disturbance will cause a damped wave of speed variations to occur. Figure 6 shows examples for different values of Q. A Q of 7, as mentioned above, will damp out to 10% in 5 or 6 cycles, or about 2 seconds for a 3 cps resonant frequency—a considerable time.

This transient disturbance provides a convenient method to measure Q: the damped waveform is recorded on an oscillograph, the ratio of successive peak amplitudes is measured, and Q found from the graph of Fig. 7.

ADDITION OF RESPONSIVENESS

Three-Element Filters

For analytical simplification let us discuss a very simple filter configuration consisting of one L and one C, with addition of an R, as shown in Fig. 8. (Many practical tape transport mechanical networks actually approach this simple case.) Our mechanical systems so far have been comparatively undamped, as shown by the left figure. A responsiveness element can be added in one of four locations in such a network. Each position, however, has some greater or lesser defect:

1. Positions with dc loss

a) R in series with L: a dc drop in voltage occurs when current is transmitted through this filter. Mechanically this means that the output speed is less than the input speed. This is satisfactory if the load is constant, or if some average-speed variation is tolerable.

b) R in parallel with C: a steady state power loss occurs here. Mechanically, this means that for adequate damping of a typical small high speed synchronous motor, the steady load would amount to ten times the available horsepower output of the motor!

2. Positions without dc loss

R in parallel with L, and R in series with C: the transmission of the filter at high frequencies approaches that of an R-C or R-L circuit, impairing the filtering action.

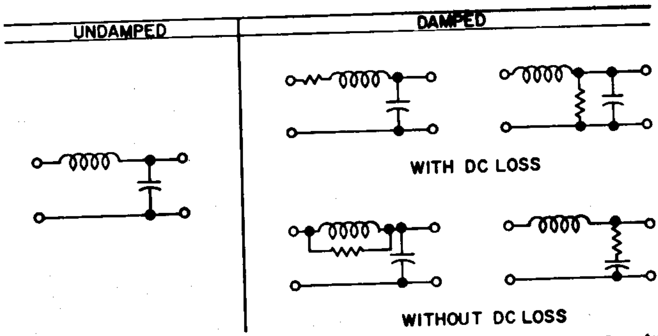

FIG. 8. Simple low-pass filter configurations, without and with damping.

Figure 9 shows two examples of filters with added responsiveness (these have been used in motion picture sound systems[3]). The effectiveness of these systems is difficult to determine, as no values are given. Figure 9a is for a shunt to ground by a "grease pad," and Fig. 9b is for a responsiveness across a series compliance, using oil driven through a small hole.

Figure 10 shows addition of responsiveness in a tape transport: the usual direct-drive capstan system using a hysteresis synchronous motor (for example the Ampex Model 351) has an equivalent mass of 22 snails, a resonant frequency of 3.5 cps for 15 ips tape speed, and a Q of 10.

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[3] W. J. Albersheim and D. MacKenzie, "Analysis of Sound-Film Drives," *J. Soc. Motion Picture Engrs.* 37, 452 (1941). The drawings of Figs. 9a and b, and 13a and b are taken from this reference.

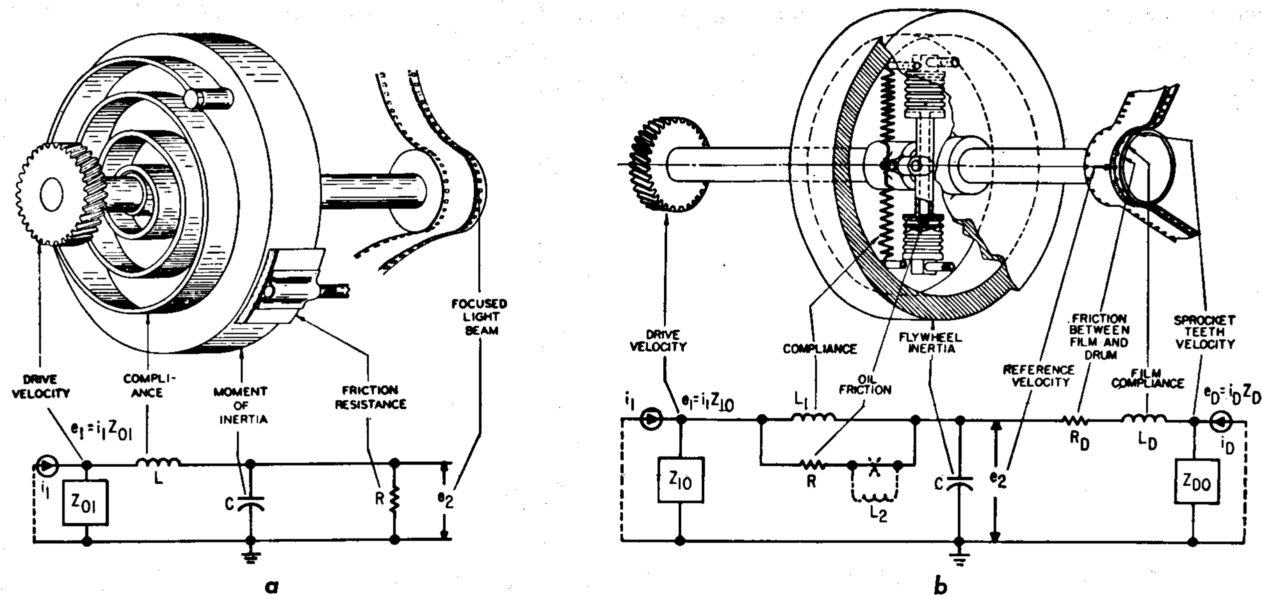

FIG. 9. Examples of mechanical low-pass filters with added responsiveness. *a*. Added shunt responsiveness to ground. *b*. Added responsiveness across a series compliance.

An indirect drive as used on the Ampex Models 300 and MR-70 uses a synchronous motor to drive the capstan flywheel through a rubber tire friction drive. The added series responsiveness damps the system to a Q of 3, with resonant frequency at 6 cps, and an equivalent mass of 50 snails. Although the series responsiveness precludes true synchronism, one must remember that the tape is always coupled to the capstan through a friction coupling, so that true synchronism is never possible in a tape recorder without servo control from the tape. In the MR-70, the hold-back tension is held constant through the reel, so that average-speed variations are minimized. The reduction in Q is of sufficient advantage to outweigh the disadvantage of slightly decreased speed regulation.

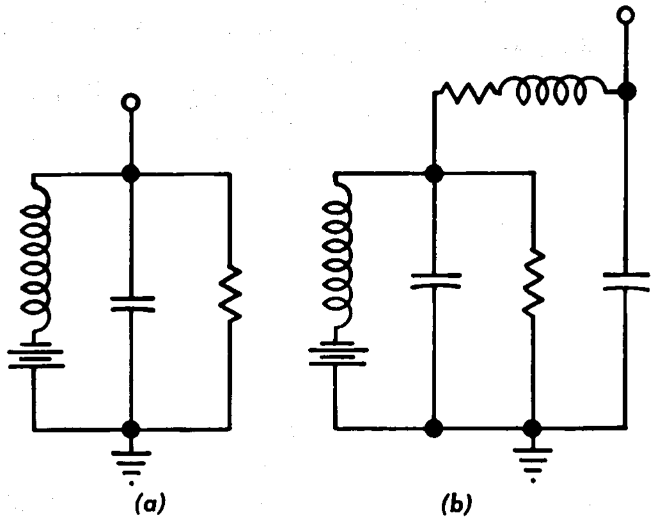

FIG. 10. Analogs of capstan drive systems employing hysteresis in synchronous motors. *a*. A direct-drive system, Q = 10 equivalent mass = 22 snails, resonant frequency = 3.5 cps). *b*. An indirect drive system, Q = 3 (equivalent mass = 50 snails, resonant frequency = 6 cps).

FOUR-ELEMENT FILTERS

Cook,[4] and Wente and Müller,[5] and Davis[6] describe four-element filters combining the good features of the response of a damped network, and the loss-free character of an undamped network. The circuit is shown in Fig. 11: an undamped capacitor (mass) *C1*, is shunted by a damped capacitor (mass) *C2*. (The Davis system places a damped

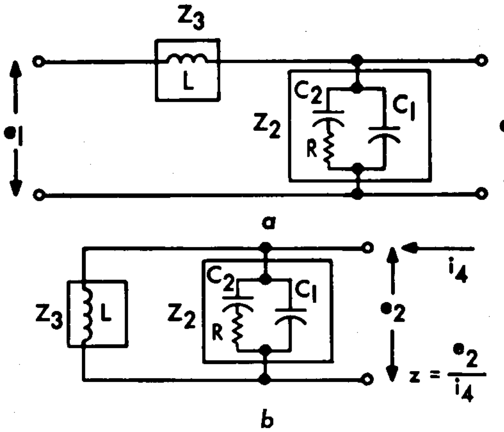

FIG. 11. Circuit of a four element low-pass filter. *a*. Circuit for response, e_2/e_1. *b*. Circuit for impedance, z/z_o.

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[4] E. D. Cook, "The Technical Aspects of the High Fidelity Reproducer," *J. Soc. Motion Picture Engrs.* **25**, 289 (1935).

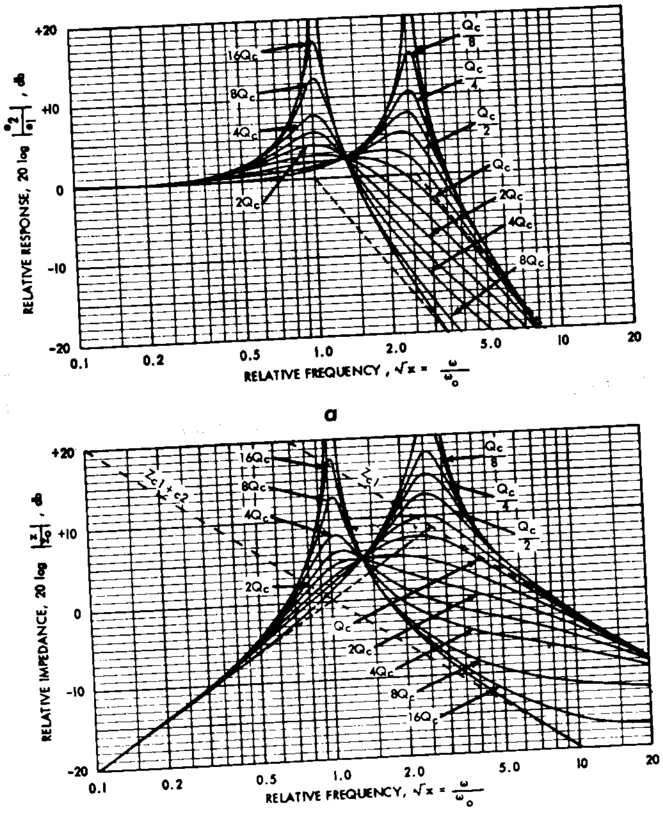
[5] E. C. Wente & A. H. Müller, "Internally Damped Rollers," *J. Soc. Motion Picture Engrs.* **37**, 406 (1941).

[6] C. C. Davis, "An Improved Film Drive Filter Mechanism," *J. Soc. Motion Picture Engrs.* **46**, 454 (1946). The drawing of Fig. 13c is taken from this reference.

inductor in series with the main inductor; the principle and results are similar.) Responses are shown in Fig. 12: a very low value of R gives a high Q at the frequency of resonance for $L + (C1$ in parallel with $C2)$; a very high value of R gives a high Q at the frequency of resonance for $L + C1$ only. At the correct intermediate value of R, the response is damped to a minimum.

Figure 13 shows three practical examples of four-element filters used in motion picture sound systems: (a) a damped inertia (viscous damper) with a solid flywheel mounted on bearings, inside an oil-filled shell; (b) a damped inertia (viscous damper) wherein the liquid in the shell is at once the inertia and, by the use of vanes, the responsiveness; and (c) a damped compliance, with a "dash-pot" on an idler system which deflects the film.

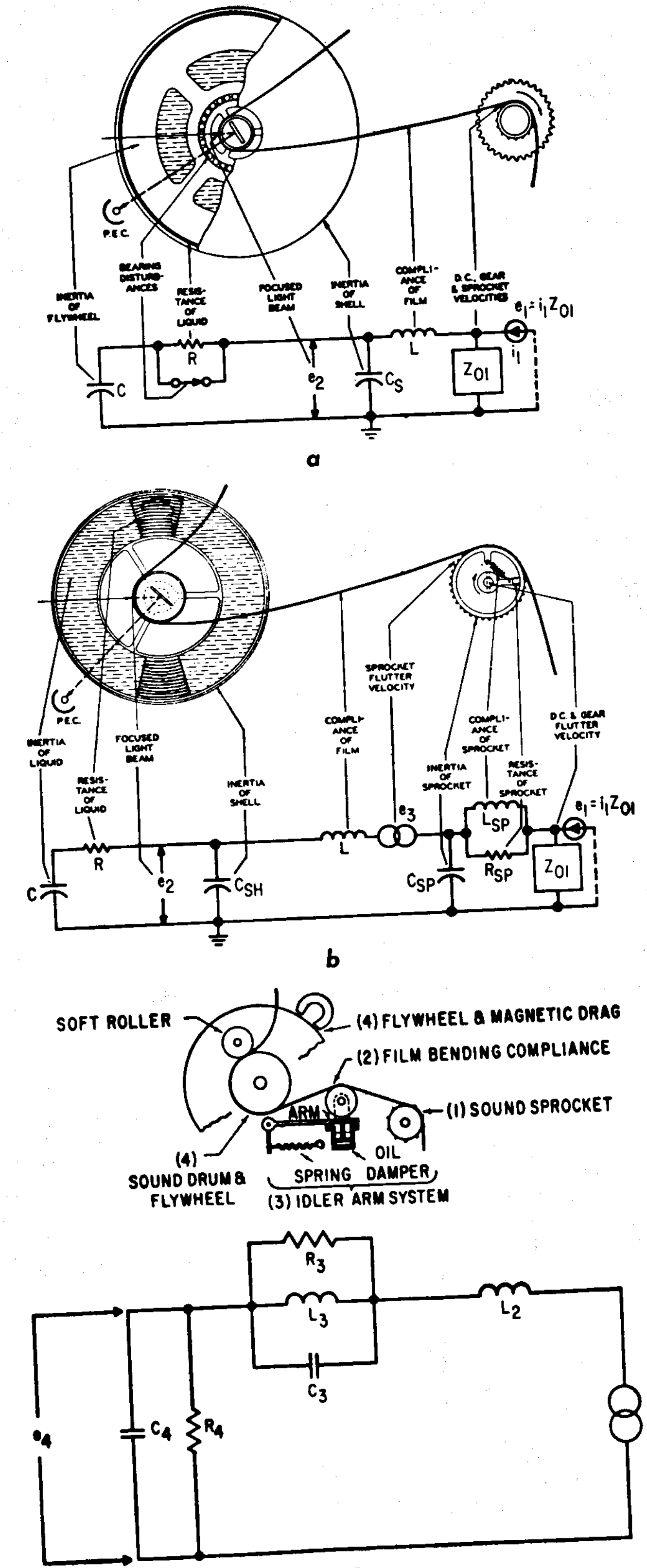
The design of such a four-element filter is described by Wente and Müller,[5] and by McKnight.[7,8] To see the effect of using a viscous-damper on the reel idler, consider Fig. 14 which shows the circuit and values of the viscous damper used in the MR-70 to replace the undamped flywheel ($C3$ of Fig. 3) used in the Models 300 and 351. Figure 15 compares responses measured with an electrical analog for the undamped reel idler flywheel, and for the viscous



a

b

Fig. 12. Response and impedance of a four element low-pass filter as in Fig. 11. $C_2 = 6C_1$. *a*. Response *vs* frequency. *b*. Impedance *vs* frequency.



a

b

c

Fig. 13. Examples of mechanical four element low-pass filters. *a*. Damped inertia (viscous damper) with a solid flywheel. *b*. Damped inertia (viscous damper) with a liquid flywheel, *c*. Damped compliance.

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[7] J. G. McKnight, "The Response of Simple Mechanical Networks," (unpublished).

[8] J. G. McKnight, "Design of a Viscous Damper," (unpublished).

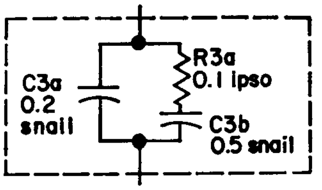


FIG. 14. Viscous damper to replace the reel idler flywheel (*C3*) in the transport analog of Fig. 3.

damper. Resultant velocity at the record head is shown for a force at the supply reel, a force at the reel idler, and a velocity at the reel idler. (In this case, the total inertia of the viscous damper equals that of the solid flywheel.)

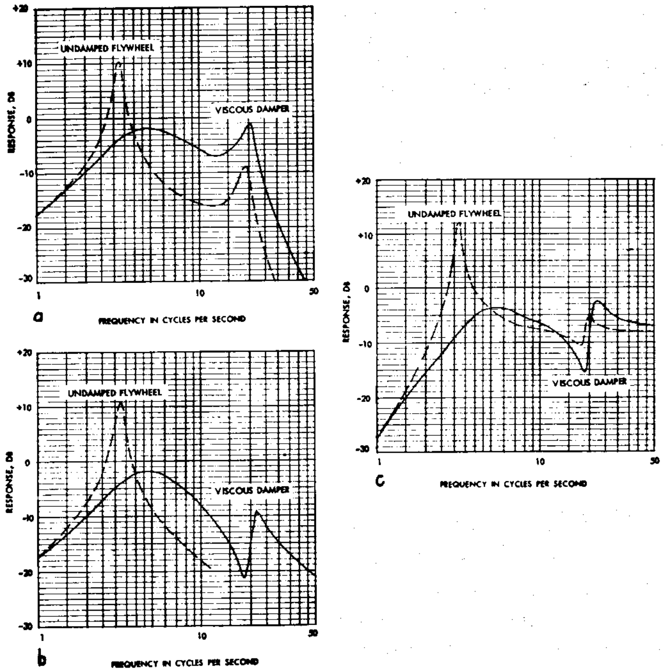


Fig. 15. Frequency response of the analog of Ampex MR-70 tape transport with ¼ inch wide tape, with an undamped flywheel, and with a viscous damper on the reel idler. Velocity at the recording head due to: *a*. a force at the supply reel. *b*. a force at the reel idler. *c*. a velocity at the reel idler.

The response at resonance for all three disturbances is seen to be decreased 12 to 16 db (i.e., one-fourth to one-sixth) when the viscous damper is used. The response at higher frequencies is slightly impaired, but still adequate.

The measured effect on an Ampex MR-70 transport with ½ in. tape is shown in Fig. 16. In this measurement the condition with the undamped reel idler flywheel (Fig. 16a) had a Q of 4.5. When the viscous damper was substituted (Fig. 16b) the Q fell to 1.7. The improvement in case of an impulsive disturbance at the supply reel is obvious.

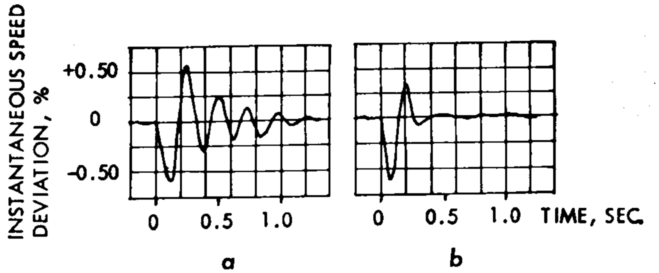
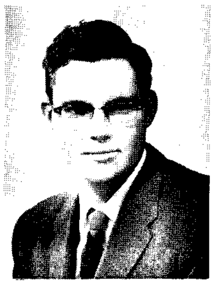


FIG. 16. Measured speed deviation *vs* time for an impulse at the supply reel, MR-70 tape transport with ½ inch wide tape. *a*. Undamped flywheel on reel idler, Q = 4.5. *b*. Viscous damper on reel idler, Q = 1.7.

CONCLUSIONS

We have shown that the mechanical system of a tape transport consists mainly of elements with low dissipation, and that the high Q resonances resulting will almost certainly coincide with one of the many possible disturbances, resulting in large speed variations (flutter) at one frequency or another. A change of mass or compliance may avoid coincidence of resonances with the worst of the disturbances, but a better solution is to add a responsiveness (damping) element to reduce the response at resonance. An example of a practical damper is shown which provides a response reduction of four to six times.

THE AUTHOR



John G. McKnight was born in Seattle in 1931. He received a B.S. in electrical engineering from Stanford University in 1952.

He has been with Ampex Corporation since 1953, with the exception of the years 1945-56 when he was assigned to the engineering staff of the U. S. Armed Forces Radio Service in New York. The Ampex Corporation appointed him manager of the advanced audio section of the Professional Audio Division in 1959, and staff engineer in the Ampex Audio Division in 1961.

Mr. McKnight's work has been in research and engineering on the dynamics of tape transports as well as on magnetic recording, especially as it concerns the recording of music. He is an amateur musician, and has presented and published papers on energy distribution in music, noise considerations and measurement in magnetic recording, equalization in magnetic recording, stereophonic recording, and transport speed variations (flutter).

He is a governor of the Audio Engineering Society and a member of the Recording and Reproducing Standards Committee of the IEEE and the Magnetic Tape Subcommittee of the NAB Recording and Reproducing Standards Committee, a Fellow of AES and a Senior Member of IEEE.

Dynamic Range Limitations in Tape Recording *

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Limitations are imposed on the dynamic range of a tape recording by the type of tape, track width and tape speed. Dynamic range will decrease as tape speed decreases, and as track width decreases; the style of tape will also control dynamic range. Equipment can limit the maximum signal by adding amplifier distortion, by incorrect bias, or by introducing even-order distortion from magnetized heads or unsymmetrical bias. Misalignment of track height between recording and reproducing heads can degrade the dynamic range. The common noise contributors are: dc noise from magnetized heads or unsymmetrical bias; noise caused by too low a bias frequency; modulation noise (both AM and FM), and reproducing head and amplifier noise. It is, however, perfectly feasible to build a recorder that poses no practical limitation to dynamic range. Even with low-noise tapes and narrow track widths, the noise is limited by the tape itself.

FOR the purpose of this paper, dynamic range will be defined as the ratio between the maximum output signal and the noise, expressed in db. It is an experimentally verified fact that audio recording is often carried to the saturation flux of the tape; therefore, the "maximum output" signal will be taken as tape saturation. It is impossible to express the dynamic range of a tape recorder with one number because both tape saturation and tape noise vary with frequency. We therefore plot dynamic range as a function of frequency, using noise density (i.e., noise power per cps of bandwidth) in place of wideband noise. (Noise power per cps is measured with a narrowband wave analyzer of known bandwidth, then converted to power in a one cps bandwidth by subtracting from the noise reading the quantity 10 log bandwidth [in cps] of the analyzer.)

The plot of dynamic range *vs* frequency is a useful tool, but even it falls short of completely specifying the system performance. It does not indicate the signal-handling capabilities of the tape system adequately, since it gives no information about the amount or type of distortion that occurs between the saturation and the noise. Our current signal-to-noise specification gives the wideband noise below the 3% harmonic distortion level; this specification too fails to indicate the signal handling capabilities, and tells nothing about the ceiling or saturation of the system.

Signal handling varies widely between tapes. As an example, 3M No. 111 tape has a "3% distortion-to-saturation" ratio of 8½ db while 3M No. 120 high output tape has a ratio of 6½ db. If operating level is established with an Ampex Standard Tape, 3M No. 111 tape will have approximately 0.8% third harmonic distortion at operating level, while No. 120 will have 0.25%. System nonlinearity will also vary widely with frequency because of short wavelength saturation and because of the presence of recording pre-emphasis at the low- and high-frequency ends of the spectrum. This distortion will also vary with the type of tape being used.

A reasonable description of system performance can be obtained by plotting saturation, response and noise density as a function of frequency. Figures 1, 2, and 3 show ex-

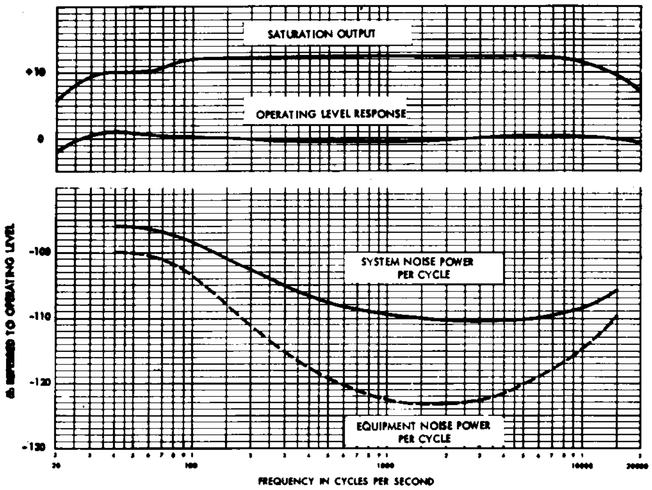

FIG. 1. Plot of 15 ips saturation, response, and noise density as a function of frequency. (Full track, 3M 203 tape. Data taken on an Ampex MR-70 tape recorder using NAB equalization, 150 kc bias, and a 234 mil wide reproduce head with 180 μin. gap.)

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* Presented October 17, 1963 at the Fifteenth Annual Fall Convention of the Audio Engineering Society, New York.

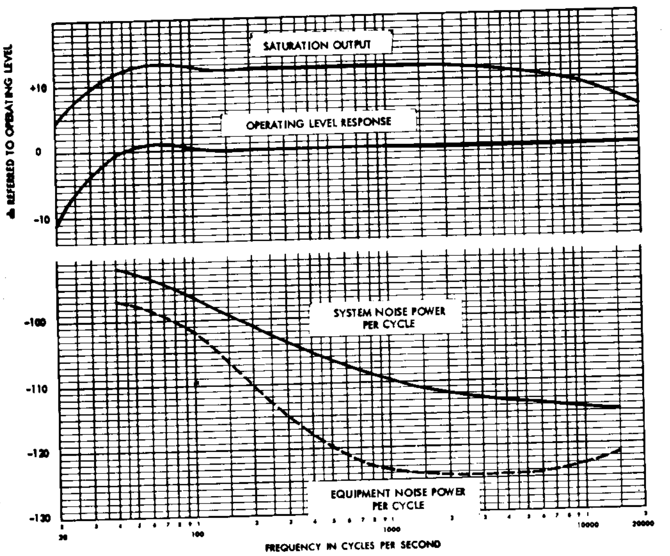

FIG. 2. Plot of 30 ips saturation, response, and noise density as a function of frequency. (Full track, 3 M 203 tape. Data taken on an Ampex MR-70 tape recorder using 17.5 μsec post-emphasis and no low-frequency pre-emphasis, 150 kc bias, and a 234 mil wide reproduce head with 180 μin. gap.)

amples of three such curves, for 7½, 15, and 30 ips. In this discussion on the limitations of dynamic range, it will first be assumed that the tape recorder has no effect on the performance;[1] under this condition, the dynamic range will vary only with track width, tape speed, and the type of tape. Then, the practical limitations that may be posed by the equipment will be discussed. Response should be run at a sufficiently low level to avoid compression. A rough indication of how signal-handling capabilities vary with frequency can be obtained by noting how close the response curve is to saturation at different frequencies. Using a mid-frequency as a reference, if another frequency is 5 db closer to saturation, one could guess the nonlinearities to be 10 db worse than at the mid-frequency, since third harmonic distortion rises with the square of the input signal.[2] Figure 4 is a summary of the data on the previous figures, showing the dynamic range for the three speeds.

Figure 5 shows how the dynamic range varies with track width; the signal is proportional to the track width w and the noise to $\sqrt{w}$. Therefore, the signal-to-noise ratio is proportional to $\sqrt{w}$. Figure 6 shows the variation of dynamic range with type of tape for a regular tape, a high output tape and a low noise tape.

Now, what practical limitations can the recorder pose on the dynamic range? On the saturation end, the most obvious limitation is imposed by the overload characteristics of the recording and reproducing amplifiers. With input

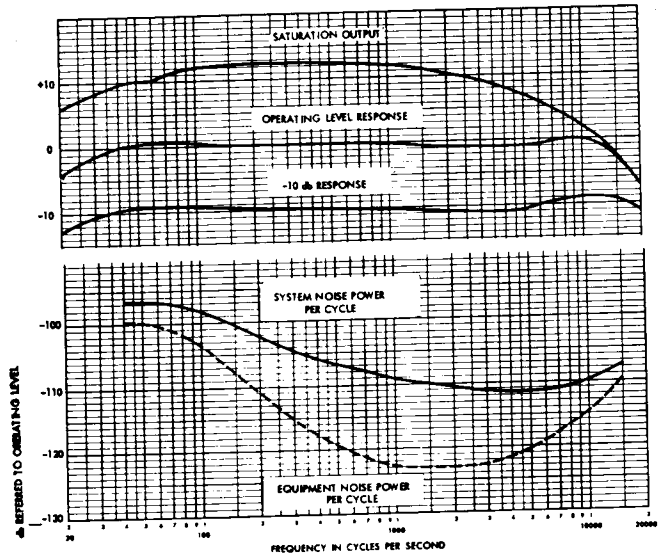

FIG. 3. Plot of 7½ ips saturation, response, and noise density as a function of frequency. (Full track, 3M 203 tape. Data taken on an Ampex MR-70 tape recorder using NAB equalization, 150 kc bias, and a 234 mil wide reproduce head with 180 μin. gap.)

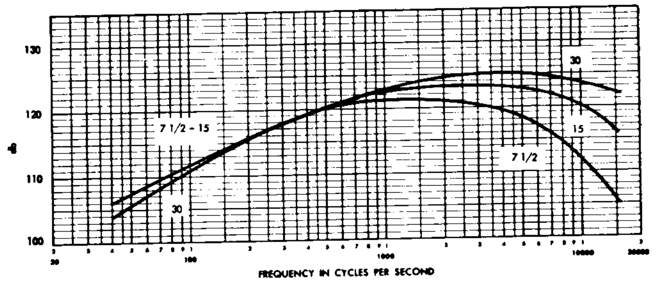

FIG. 4. Dynamic range as a function of frequency for 7½, 15, and 30 ips. (Full track, 3M 203 tape.)

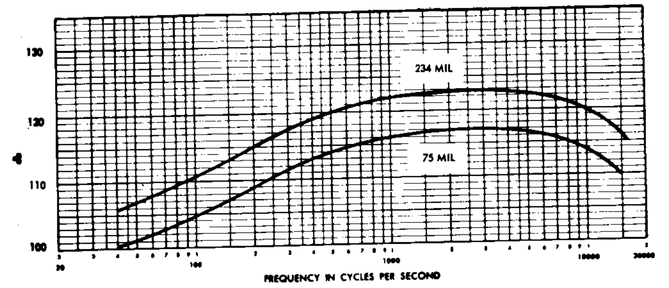

FIG. 5. Dynamic range as a function of frequency for track widths of 75 and 234 mil. (15 ips, 3M 203 tape.)

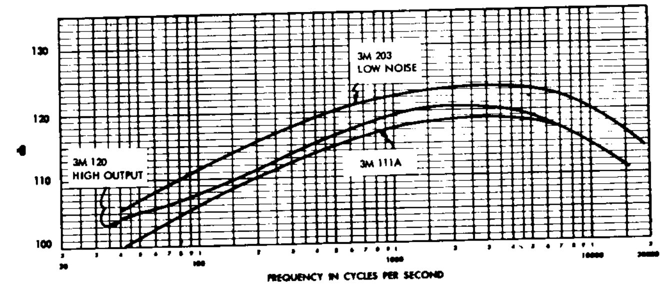

FIG. 6. Dynamic range as a function of frequency for three styles of tape. (15 ips, full track.)

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[1] Figures 1, 2, and 3 also show the equipment noise. As long as the equipment noise is 6 db below the tape noise, it will only raise the system noise by 1 db. It can be seen that from 100 cps to 10,000 cps the dynamic range is controlled by the tape.

[2] R. Z. Langevin, "Intermodulation Distortion in Tape Recording," *J. Audio Eng. Soc.* **11**, 270 (1963). See especially Fig. 1 of this reference.

signals 20 db above the "operating level" (vu meter zero), the tape has still not reached complete saturation. Music peaks have been measured 26 db above a vu meter zero indication. An amplifier that clips can also generate higher-order distortion products which can be most annoying; this clipping is especially likely when operating in the "red" of the vu meter. Crosstalk in recording and in reproducing, print-through, and insufficient erasure are also obvious offenders.

The magnetic recording process is inherently symmetrical, which means that even-order distortion will not be present. Even-order distortion can be introduced, however, either through even-order distortion in the recording or reproducing amplifiers, or by a dc flux in the recording head due to a magnetized head or asymmetrical bias waveform.[3]

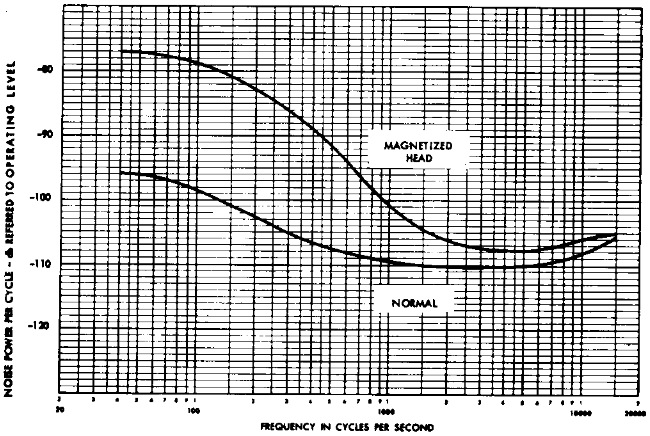

Fig. 7. Noise density as a function of frequency for a magnetized head and a normal head. (15 ips, full track, 3M 203 tape.)

A practical limitation to dynamic range which is not obvious is misalignment between the heights of the recording and the reproducing heads. This can be a particular problem with narrow track-widths. It occurs because the "operating level" of the reproducer is normally set with a full-track standard alignment tape. The recording level is then set to produce the same output as the standard tape. If the track from the recording head misses the reproducing head, it is necessary to record at a higher flux level to obtain the same output from the reproducing head that the standard tape gave. This higher flux level being recorded has a greater distortion than the nominal 1%, thereby reducing the distance to the saturation level. The reproducer still "sees" essentially the same tape noise.

If the bias wavelength is too long (i.e., too low a frequency for a given speed), the noise will increase. At 30 ips, for instance, an increase of bias frequency from 100 kc to 150 kc causes a noticeable reduction of noise.

If the bias amplitude is less than that necessary for maximum recording sensitivity (peak bias) for the tape in use, the distortion increases. This has been discussed in detail in the literature.[4-7]

The system including tape recorder/tape/reproducer may amplitude- and/or frequency-modulate a signal passed through it; the modulation sidebands are often audible, and constitute modulation noise.[8] The modulation noise will usually set the practical limits to the dynamic range *in the presence of a signal.*

AM noise is caused by poor tape-to-head contact and irregularities in the tape. "DC noise" is a form of AM noise which occurs when unsymmetrical bias waveform or magnetized heads or tape guides cause a dc magnetization to be present all of the time.[9] This dc is amplitude modulated, and the audible result consists of pops and crackles plus a general increase in hiss level. Figure 7 shows how a slightly magnetized head can increase the noise. (This magnetization was insufficient to cause erasure of a 1 mil wavelength signal—15 kc at 15 ips—yet the noise increased dramatically.) When an ac signal is recorded, it is similarly amplitude-modulated.[10] The effect is audible mainly with low-frequency signals; it also has the sound of pops and

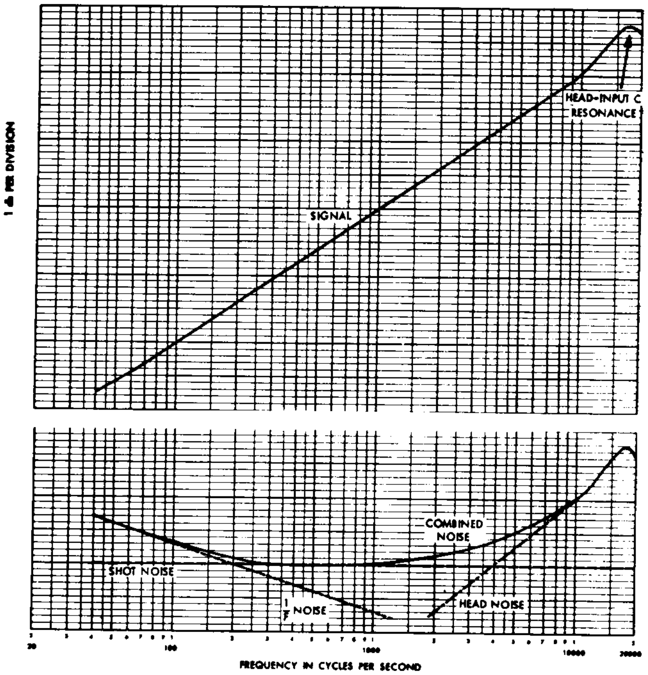

Fig. 8. Signal and noise as a function of frequency from a conventional reproducing head-preamplifier system. (Hypothetical case. All equalization removed.)

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[3] *Ibid.* See especially Fig. 2 of this reference.

[4] G. L. Dimmick and S. W. Johnson, "Optimum High-Frequency Bias in Magnetic Recording," *J. Soc. Motion Picture Engrs.* 51, 489 (1948). See Figs. 2, 3, and 4.

[5] R. Herr, B. F. Murphey, and W. W. Wetzel, "Some Distinctive Properties of Magnetic-Recording Media," *J. Soc. Motion Picture Engrs.* 52, 77 (1949). See Figs. 1 and 2.

[6] W. K. Westmijze, "Studies on Magnetic Recording," *Philips Res. Rep.* 8, 148 (1953). See Figs. 33 and 34.

[7] E. Belger and P. Scherer, "Investigations of More Recent Types of Magnetic Recording Tapes," *Rundfunktechnische Mitteilungen* 5, 193 (1961). (In German) See Figs. 4, 5, 6, and 9.

[8] P. Smaller, "An Experimental Investigation of the Noise in Magnetic Tape Recording which is a Function of the Tape Characteristics," *J. Audio Eng. Soc.* 7, 196 (1959).

[9] J. W. Gratian, "Noise in Magnetic Recording Systems as Influenced by the Characteristics of Bias and Erase Signals," *J. Acoust. Soc. Am.* 21, 74 (1949).

[10] D. F. Eldridge, "DC and Modulation Noise in Magnetic Tape," *Proc. Intermag. Conf., IEEE,* T-149 (April, 1963).

crackles. This noise is more noticeable at higher tape speeds.

Frequency modulation with a high modulating frequency (e.g., around 3 kc) is called "scrape flutter"; it is noticed when recording high frequencies, and results in a hiss riding along with the tone.[11]

The reproducing head and preamplifier system is the final offender. The preamplifier usually limits the dynamic range at the lower frequencies. The $1/f$ noise in tubes, resistors, and transistors normally predominates. However, with low-noise tapes it is possible to approach the shot noise in the 200 to 500 cps region. At the extreme high end, the eddy current loss in the reproducing head normally limits dynamic range. Both reproducing head and preamplifier can introduce hum (the power line frequency plus its harmonics) in reproduction.

Figure 8 shows the noise and signal present in a conventional reproducing head plus preamplifier combination with all equalization removed.[12] (This hypothetical case, based on a flat amplifier, illustrates how the different noise components contribute to dynamic range. In a practical reproducer, the reproducing preamplifier is not flat but falls at a 6 db per octave rate until 3180 cps, where it flattens out for the 50 microsecond post-emphasis. However, this equalization will affect both signal and noise in the same way, so that the signal-to-noise ratio will remain unchanged.)

It can be seen that in the high-frequency region, the head noise predominates and the signal-to-noise ratio is constant. As the mid-frequency region is approached, shot noise in the tube or transistor controls. This causes the signal-to-noise ratio to deteriorate at a 6 db per octave rate as the frequency decreases. At the low frequencies the $1/f$ noise of the amplifier takes over. It increases at a 3 db per octave rate which results in a signal-to-noise degradation of 9 db per octave.

In all of the other curves presented in this paper, we have used a new form of reproducing preamplifier developed by Erling Skov.[12] It completely eliminates the active first stage as a noise source. Thus, it eliminates the effect of shot noise and $1/f$ noise; the only noise remaining is that contributed by the head and the input transformer. An examination of Fig. 8 shows that the elimination of shot noise effects results in an improved signal-to-noise ratio in the mid- and low-frequency region.

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THE AUTHOR

Robert Z. Langevin was born in 1925, in California. He attended the University of Colorado and the University of Michigan, where he received a B.S. degree in electrical engineering in 1946. Through 1950 he worked for companies founded by his father: Langevin Manufacturing Corporation, and Carl Langevin, Incorporated. In 1951 he joined Ampex Corporation and became associated with the engineering of audio tape recorders. In 1960 he served with Vega Electronics Corporation as project engineer for their wireless microphone. He rejoined Ampex in 1961, where he is presently engaged as an electronic engineer in the Audio Division.

Mr. Langevin is a member of the Institute of Electrical and Electronics Engineers and a charter member of the Audio Engineering Society.

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[11] R. A. von Behren and R. J. Youngquist, "Frequency Modulation Noise in Magnetic Recording," *J. Audio Eng. Soc.* 3, 26 (1955).

[12] E. P. Skov, "Noise Limitations in Tape Reproducers," *J. Audio Eng. Soc.* **12**, 280 (Oct., 1964).