THE TECHNICAL ASPECTS OF THE ES-200 SYSTEM

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March 1, 1969

TABLE OF CONTENTS

LIST OF FIGURES

LIST OF TABLES

PREFACE

It is not my intention to completely describe the ES-200 System. Rather, I will attempt to describe those technical aspects of the system which were particularly my own contribution, and which, therefore, it might be difficult to otherwise understand. Portions of the Demodulator, for example, are uniquely Maung Gyi's and will not be discussed here. For information about the Reproduce Equalizers, Rich Smith should be consulted. For the many details about the harnesses, cabling, grounding, sheet metal, drawings, shieldings, etc., Don Burri is most knowledgable. The -20 volt slow-turn-on-and-off regulator was designed by Gordon Svendsen, the preamp power supply by Ron Wagner, the monitoring circuitry by Hartvig Melbye.

1. SYSTEM BLOCK DIAGRAM

A sketch of the block diagram of the ES-200 System is given in Figure 1 with assembly and schematic drawing numbers for reference.

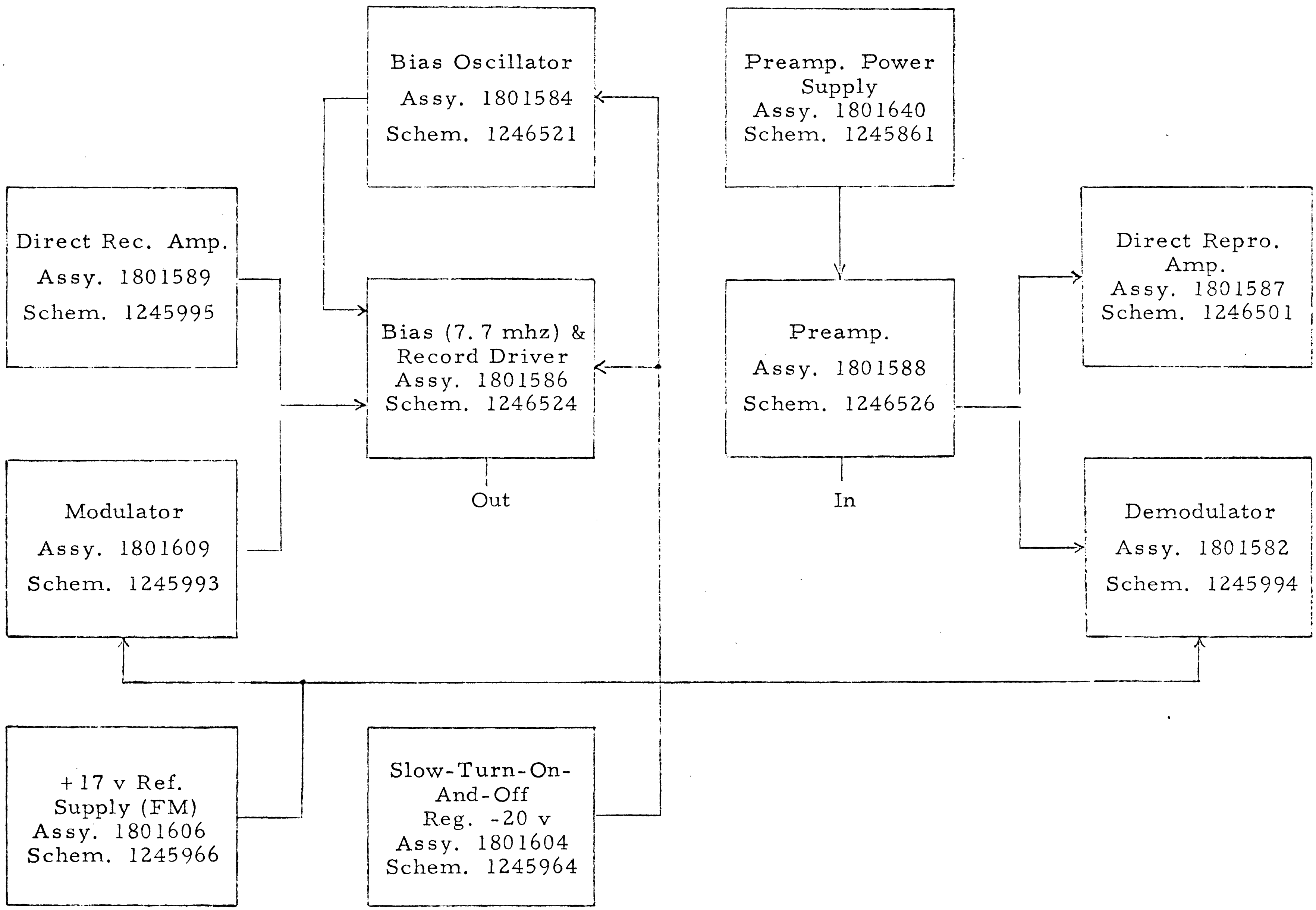

Figure 1. ES-200 System Block Diagram

2. RECORD SIGNAL PATH CONSIDERATIONS

Some general decisions were made at the outset. First, it is not convenient, for physical reasons, to put signal path adjustments in the record driver bay. There is not room and they are physically difficult to get to. Therefore, the record driver must accommodate all head variations and whatever overdrive is required (20 db over the 1% third level, in the ES-200 System). These variations must then be scaled back to the level adjustment where minimum and maximum voltage requirements may then be computed for the drive for this adjustment. Then, knowing the desired range of input levels, input amplifier and divider requirements are calculated to provide this required drive in all cases.

Since the input level range is 80:1, and since a 2:1 range in head drive requirements may be anticipated, it becomes necessary to use some jumpered input dividers. No amplifier could handle the range, and no pot will provide adequate resolutions for such a large variation in levels. A detailed discussion of how these requirements are worked out is given in the remainder of Section 2.

2.1 Record Amplifier

A quasi-block diagram of the signal path portion of the record system is given in Figure 2. A conventional approach to the problem (20 KΩ input, 20 db overdrive, 0.5 db response from 50 hz to 2.4 mhz) might have been to put the gain adjustment at the input and pick off the

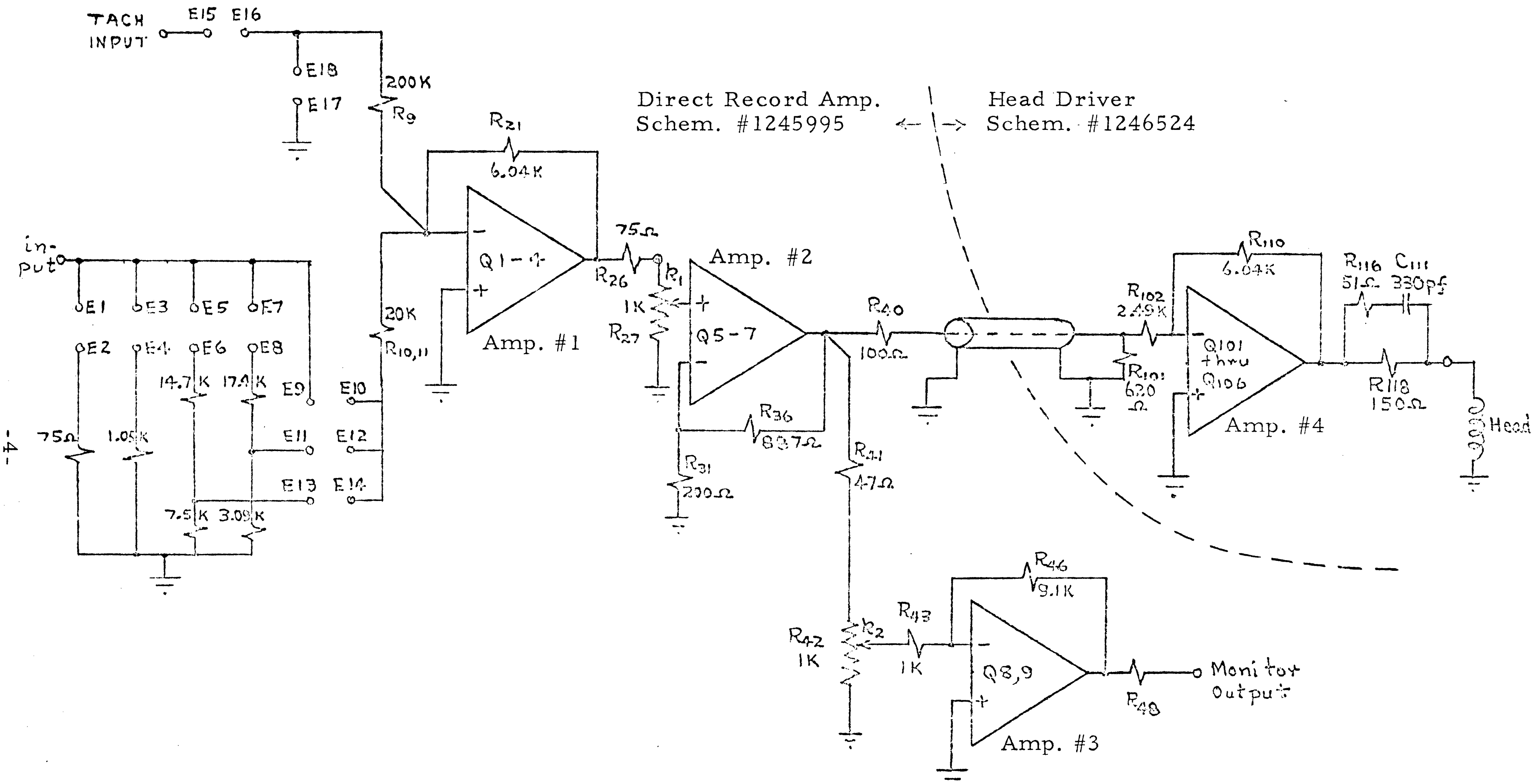

Figure 2. Signal Path Portion of Record System

signal from the adjustment with a series feedback, high input impedance amplifier. However, the stray capacitance of a 20 K pot alone precludes this kind of approach, without even considering the additional loading of the resistance and capacitance of an amplifier. The only possibility is to use a shunt feedback amplifier Q1 - Q4 as an impedance converter and drive the level adjustments R_{27} with it. The key to being able to do this is the shunt feedback triple which is the only amplifier that will provide the bandwidth, input impedance, low distortion, and output swing required to satisfy all the conditions. The design of this unusual amplifier will be discussed in Section 7.

Since the shunt feedback amplifier has a virtual ground at its input (base Q1), looking into the amplifier at E10, 12 or 14, we see the sum of R_{10} and R_{11}, or almost exactly 20 KΩ. Thus with only jumper E9-E10 in place, the input impedance is 20 K. The input resistor is split into R_{10} and R_{11} and bypassed between the two, to prevent the stray capacitance of a 20 K resistor from causing the amplifier's gain to rise at very high frequencies. The divider resistors are picked to give the desired attenuation and still look like 20 KΩ at the input. For example, with jumpers E5-E6 and E13-E14 in place,

$$14.7\ \text{K} + \left(7.5\ \text{K} \parallel 20\ \text{K}\right) = 20.15\ \text{K} \qquad \text{Eq. 1}$$

The dividers are chosen so as to provide the following input ranges:

0.250 vrms to	1.50 vrms	no divider
0.903 vrms to	5.48 vrms	1st divider
3.330 vrms to	20.00 vrms	2nd divider

Table 1. Input Ranges of Record Amplifier

The dividers should be selected such that the output range covered by amplifier #1 is always the same. Therefore:

$$\frac{1.50}{0.25} = \frac{5.480}{0.903} = \frac{20}{3.33} = 6 \qquad \text{Eq. 2}$$

Furthermore, so that the overlap of the three ranges will be the same in db, the ranges need to have a geometric relationship to each other. Therefore:

$$\frac{0.903}{0.250} = \frac{3.330}{0.903} = \frac{5.48}{1.50} = \frac{20}{5.48} = \sqrt{\frac{20}{1.5}} = \sqrt{\frac{3.33}{.25}} = 3.65 \qquad \text{Eq. 3}$$

Conveniently, two of the ranges include 1 vrms which is a common input level.

From the input (with jumper E9-E10 in place), to the current in the head, the following transfer equation applies at mid-band frequencies:

$$\frac{i_{head}}{e_{input}} =$$

$$\overbrace{\frac{R_{21}}{R_{10,11}}}^{\text{Amp. \#1}} \quad \frac{R_{27}}{R_{26}+R_{27}} \quad k_1 \quad \overbrace{\frac{R_{31}+R_{36}}{R_{31}}}^{\text{Amp. \#2}} \quad \frac{R_{101} \| R_{102}}{R_{40}+R_{101} \| R_{102}} \quad \overbrace{\frac{R_{110}}{R_{102}}}^{\text{Amp. \#4}} \quad \frac{1}{R_{118}} \qquad \text{Eq. 4}$$

$$\vdots \qquad \vdots \qquad \vdots \qquad \vdots \qquad \vdots \qquad \vdots \qquad \vdots$$

$$\frac{6.04K}{20K} \quad \frac{1K}{1.075K} \quad k_1 \quad \frac{200+887}{200} \quad \frac{0.62K \| 2.49K}{0.1K+0.62K \| 2.49K} \quad \frac{6.04K}{2.49K} \quad \frac{1}{150\Omega} \qquad \text{Eq. 5}$$

$$\vdots \qquad \vdots \qquad \vdots \qquad \vdots \qquad \vdots \qquad \vdots \qquad \vdots$$

$$0.302 \quad 0.93 \quad k_1 \quad 5.43 \quad .833 \quad 2.42 \quad \frac{1}{150\Omega} \qquad \text{Eq. 6}$$

The gain to the top of R_{27} is (the first two terms in Eq. 6)

$$0.302 \times 0.93 = 0.281 \qquad \text{Eq. 7}$$

Thus, for a maximum input (in any range) the voltage at this point is

$$1.5 \text{ vrms} \times 0.281 = 0.421 \text{ vrms} \qquad \text{Eq. 8}$$

With a 20 db overdrive, this becomes 4.21 vrms or 11.9 vptp. Clearly, we are approaching a maximum limit in output capability of a high frequency amplifier.

The requirements for low distortion place a limit on the maximum gain achievable in Amp. #2. Therefore, we are led quickly to the maximum achievable input voltage range for Amplifier #1 and thereby to the required number of input dividers to cover the 80:1 range in input voltage.

One arbitrary assumption is made about the system: the voltage on the cable going from the direct record amp. to the head driver shall be 0.5 vptp and the input impedance to the head driver card shall be 500 ohms. These are convenient values from the RFI standpoint, since both voltage and current are low. Another not-so-arbitrary assumption is made that record heads will vary 6 db or 2:1 in the record current required to achieve a 1% third recording. Then, geometrically centering this 2:1 spread about 0.5 vptp, we determine that head driver parameters will be chosen so as to guarantee 1% third recording with between 0.333 vptp and 0.667 vptp on the cable between the record amp and head driver. The gain from the wiper of pot R_{27} to the cable is, from Eq. 6

$$\text{Gain} = 5.43 \times 0.833 = 4.52 \qquad \text{Eq. 9}$$

Then, working backward from the cable, the spread of voltage required on the wiper of pot R_{27} to achieve 1% third recording is 0.0738 vptp to 0.1477 vptp. The minimum available at the top of the pot R_{27} is

$$0.25 \text{ vrms} \times 0.281 \times 2\sqrt{2} = 0.198 \text{ vptp} \qquad \text{Eq. 10}$$

and since the maximum required is 0.148 vptp, there is a 34% reserve on pot R_{27}. This is quite adequate to account for tolerances in gain determining resistors (all 1% metal film).

The 20 db overdrive is not so hard to meet in Amp. #2 because R_{27} will never be adjusted higher than 0.667 vptp on the cable; the highest output swing that Amp. #2 is required to handle is then

$$\frac{0.667}{0.833} \times 10 = 8 \text{ vptp} \qquad \text{Eq. 11}$$

Care must be taken to generate positive gain to both the head and the monitor outputs. Since the input Amplifier #1 is of the virtual ground, inverting type, and Amplifier #2 is positive gain, another inversion must be generated in both Amplifiers #3 and #4. Since the monitor output is not a critical one, a simple inverting pair is adequate. The feedback resistors are chosen so that with the smallest voltage at the output of Amplifier #2, a minimum of 3 volts ptp may be obtained at the monitor output (open circuit). See Hartvig Melbye for the requirements of this output. No attempt is made to prevent the monitor output amplifier from limiting on a 20 db overdrive.

2.2 Head Driver

Amplifier #4 in Figure 2 is, therefore, also a shunt feedback triple (see Section 7). Head current is generated by supplying a voltage to the left end of R_{118}, the assumption being that the head impedance is low enough compared to R_{118} that a current source is approximated. R_{116} and C_{111} must be chosen experimentally such that the desired pre-emphasis is obtained. Any other than an experimental approach is difficult because the inductance and Q of the head change with frequency (a non-linear system). R_{110} is determined by amplifier design considerations: if it gets larger, there is not enough feedback and the stray capacitance will begin to limit the amplifier's bandwidth. (Notice that R_{110} has the same value as R_{21} in Amp. #1.) Then R_{102} is selected to provide the desired voltage gain from input to amplifier output. This is determined by the maximum allowable output swing, 16 vptp, and the highest input voltage, 6.67 vptp. R_{101} is then selected to provide a 500 ohm input impedance.

Therefore, the voltage gain of Amplifier #4 may be increased by reducing R_{102} and increasing R_{101}. The limit occurs at the point where R_{102} becomes 500 ohms and R_{101} approaches infinity. Any increase in gain, however, reduces the number of db of overdrive available (assuming 1% third recording is still obtained with 0.333 to 0.667 vptp at the cable between the record amp. and the record driver).

2.3 Some Possible Variations to the Record Signal Path

A. Assume that, for some reason, the current required to produce 1% third recordings increased. The only possible way to accommodate this change without sacrificing the 20 db overdrive capability is to decrease R_{118} and reselect the pre-emphasis network. If, for example, the gain of Amplifier #4 is increased, it will limit before the 20 db is reached. The same argument applies to increasing gain at any earlier point in the system.

B. Assume that a customer wanted to be able to produce 1% third recordings with any voltage from 0.25 vrms to 5 vrms by simply adjusting the gain control without having to make any jumper changes. The only way this may be accomplished is by telling him that he must jumper E9-E10 (E1-E2, E3-E4 optional) and give up the 20 db overdrive capability. He may then do this without any circuit changes but with an overdrive limit of

$$\text{overdrive limit} = \frac{1.5 \text{ vrms} \times 10}{5 \text{ vrms}} = 3 = 9.5 \text{ db} \qquad \text{Eq. 12}$$

The reason for this is that the input Amplifier #1 will limit with an input greater than 15 vrms. (See Table 1.) (20 db greater than 1.5 vrms.)

2.4 A Summary of Design Philosophy

A. It is desirable to provide capability of 20 KΩ input, 50 hz to 2.4 mhz bandwidth ± 0.5 db, 20 db of overdrive simultaneously. Then there are no special constraints which must be enumerated in specifications and the amplifier becomes all-purpose.

B. The only way to achieve A. is to use the shunt feedback triple (see Section 7) as an impedance converter to drive the record level adjustment.

C. The way to accommodate an 80:1 range in input voltage is to use two attenuators and split the 80:1 range up into three 6:1 overlapping ranges. This permits reasonable amplifiers and reasonable record level pot resolution.

D. Care must be taken to provide positive gain to both the record head and the monitor output.

E. The best way to drive a record head is to generate a voltage source and drive the head through a resistor to provide a constant current.

F. The only way to achieve 16 vptp from a 20 volt power supply at 107 maptp to drive a head with low second order distortion (for FM) is the shunt feedback triple. (See Section 7.)

G. It is well to have an easy way to obtain more head current (change R_{118}) or more voltage gain in the head driver (change R_{101} and R_{102}).

H. To keep RFI low, it is well to operate the cable between the record amp. and the head driver at a low voltage level (0.5 vptp) and low current level (500 Ω load, 1 maptp).

I. The direct record signal path is unique -- as far as I know. Certainly it is unique to Ampex. It should be capitalized on in the advertising and sales efforts -- when customers get used to a record signal system which does everything, they will not settle for less.

3. BIAS GENERATION CONSIDERATIONS

3.1 Bias Frequency

According to IRIG 106-66, the bias frequency should be 3.2 times the highest signal frequency. In anticipation of a 2.4 mhz system (3.2 x 2.4 = 7.67), 7.7 mhz was chosen.

There is, however, another condition which must be considered. When FM is used, the carrier is approximately a square wave and so has a considerable third harmonic (1.296 mhz for a 432 khz carrier). The record process is very non-linear in an odd-order fashion; products such as $3f_1$, $3f_2$, $2f_1-f_2$, and f_2-2f_1, are generated in the magnetization on the tape. Early in the design of the I system, we picked 3.0 mhz as a bias frequency for use with a 432 khz carrier and noticed a bad spurious at 24 khz. In this case, f_1 was the third of the carrier (1.296 mhz) and f_2 was the bias (3 mhz). Then

$$f_2-2f_1 = 3 \text{ mhz} - 2 \times 1.296 \text{ mhz} = 408 \text{ khz} \qquad \text{Eq. 13}$$

And since

$$432 \text{ khz} - 408 \text{ khz} = 24 \text{ khz} \qquad \text{Eq. 14}$$

the 408 khz was demodulated as a 24 khz spurious. The bias frequency was raised to 3.4 mhz and the problem disappeared.

3.2 Bias Oscillator

A block diagram of the bias oscillator is given below in Figure 3. (See Schem. 1246521.) This design was originally done by Maung Gyi for the Mark II system. I am discussing it here because we did a great deal of work trying to improve it. Our biggest concern was to try to eliminate emitter followers (because they tend to oscillate) but we finally went back to the original design.

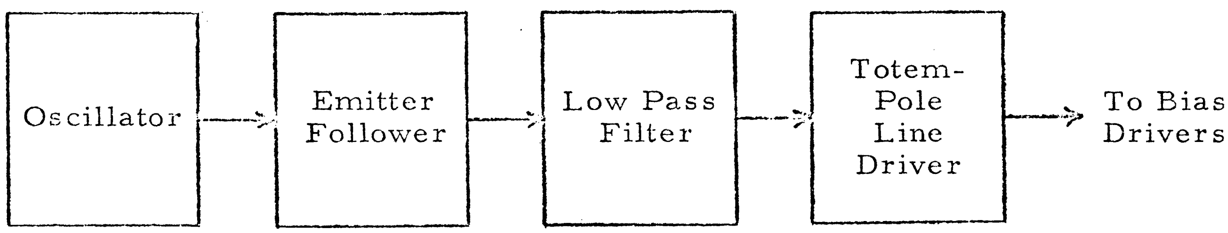

Figure 3. Bias Oscillator Block Diagram

The requirements of the bias oscillator are several. First, only one bias frequency may be used in any one recorder because the beat frequency which would occur between two nearly similar bias frequencies would appear everywhere. Second, it is desirable that the bias oscillator output driver have as nearly zero an output impedance as possible so that bias currents in other channels will not change if one or more bias drivers are removed. This precludes a feedback amplifier because without a series output resistor, the feedback loop would certainly oscillate with some combination of load.

The Totem Pole output stage seems to be a good solution. It has some feedback but the feedback factor is always less than one so the loop cannot oscillate. It is subject to emitter follower oscillations, however, so it is essential that the 100 Ω base resistor be used. The filter helps the output look sinusoidal; it is driven by an emitter follower, again with a base resistor.

The oscillator is a standard one. Some starting difficulty with this oscillator circuit has been reported in the past but we have used more feedback, resulting in a less linear, but more reliable oscillation.

A curious coincidence arose in the output driver. As the load current was increased, the voltage at the collector of the emitter follower output stage Q3 increased, thereby increasing the Miller multiplication of the collector capacitance of Q3. This modified the response of the filter in such a way as to increase the gain of the filter slightly. Thus, it is possible to see bias oscillators with zero output impedance. This is a second order effect, but it helps.

3.3 Bias Driver (See Schem. 1246524)

Several significant improvements have been made in the ES-200 Bias Driver over old designs. First, an emitter follower, Q107, is used to keep the load small for the bias oscillator. Next, to keep the oscillator's load sinusoidal a tuned circuit, L102, C114, is used in the driver (an idea contributed by Don Burri). In the FR-1400, 1800 driver the adjustment was done in the emitter of the class C driver. This means that an unreasonably large non-sinusoidal current flowed out to the adjustment pot and back to the transistor. This probably accounts for most of the high frequency RFI difficulties in the older versions. In the ES-200 the adjustment is done before the tuned circuit where the current is small and sinusoidal. Furthermore, the adjustment may be turned all the way to zero. Two well-heat-sinked transistors are used for the class C drivers (Q108, 109) and their biasing is done with the help of a diode CR103 which tracks the emitter-base voltage of Q108 and 109 with ambient changes in temperature. The result is a surprisingly stable output with temperature.

The argument for using class C drivers is that this is as close as one may come to pure switching without the stored-base-charge problems of saturated transistors. The key to successful class C operation is to return the high frequency currents removed by the filter (through L107 and C119) to their point of emanation from ground. It is absolutely imperative that this portion of the circuit be point grounded; that is, that each component's ground lead be brought separately to a single point.

One other feature of the ES-200 design is improved over the FR-1400. A torroid transformer was used rather than an air-core. This contains the flux which should eliminate coupling from one board to the next and the resulting necessity to adjust bias for each track several times.

One may wonder why C115 does not appear to resonate with L103 at the bias frequency. The capacitor was made smaller so that when added to the Miller collector capacitance of the drivers Q108 and 109, the total will resonate at the bias frequency. Even though in class C operation, this may be done since a large portion of the base drive current in Q108 and Q109 is Miller capacitance current and therefore roughly linear ("roughly," since C_{ob} varies inversely as the square root of collector voltage). This current flows even when the transistors are cut off. The Miller capacitance is perhaps best estimated by simply measuring the voltage gain from the bases to the common collectors of Q108 and Q109.

The rest of the bias arrangement is quite standard. See the FR-1400, for example, or the FR-1600 bias filtering and trapping.

4. THE PROBLEM OF MAGNETIZED RECORD HEADS

4.1 Slow-Turn-On-And-Off Power Supply

(See Schematic #1245964, Assembly or SP #1801604.)

This power supply, located in the main electronics tray power supply section, is controlled by the record relay. It supplies regulated power to the record and bias driver card and to the bias oscillator card. The record amplifier is always on, providing the normal loading to the customer and the normal monitor output. When the record relay is energized, therefore, this supply applies power to the record head driving electronics, and when it is not energized, there is no power -- the record head is completely at rest.

There is a difficulty which arises, however, if the voltage rises and falls too fast. If the rise/fall time is 0.1 sec., repeated operations of the record relay will cause the record head to become seriously magnetized. A rise/fall time of 0.3 sec. will result in slight magnetization. (A magnetized record head is recognized by an increase in second harmonic distortion from its normally desirable level of -50 db or more.) However, if the rise/fall time is 1 sec. or more, repeated operations of the record relay result in no appreciable change in second harmonic. The "slow-turn-on-and-off power supply" is specified to have a 10% to 90% rise and fall time of 1 sec. ± 25%.

It is believed that the all-metal record head is more susceptible to magnetization than the older composite heads. No explanation is available.

The reason for the record heads becoming magnetized is apparently that a sustained unipolar current flow above a certain magnitude will cause the record heads to magnetize. It was postulated that the offending current comes out of the output coupling capacitors in the record and bias drivers. Since the bias output capacitor is 51 pf (C125) and the record driver output capacitor is 47 μf (C112), it is apparent that the principle offender is C112. (See Head Driver Schematic #1246524.) It is possible that the bias driver could be operated on any normal power supply without magnetizing heads -- but it is simpler to have the whole head driver system operate from the same supply.

4.2 Unipolar Record Amplifier Inputs

There is one other cause of magnetized record heads. If the record amplifier input (with the recorder in the record mode) is pulsed with a unipolar input of sufficient amplitude, the record head will magnetize. Some data on this effect was taken. The pulse was generated by simply tapping the input lead to a known voltage one time. The results are given in Table 2.

	Amplitude of Unipolar Pulse	2nd Harm. Level
1% 3rd level	0.18 volts peak (sine wave)	-54 db
	0.36 volts peak (pulse)	-54 db
	0.72 volts peak (pulse)	-47 db
	1.44 volts peak (pulse)	-31 db

Table 2. Record Head Magnetization From Unipolar Pulses

Apparently, a pulse peak 6 db over the peak amplitude of that sine wave which yields 1% third recordings, will not magnetize the head. However, a pulse 12 db over will; a pulse 24 db causes serious magnetization.

5. PREAMP DESIGN

During the initial phases of the ES-200 design, it was believed, since the noise figure of the FR-1400 preamp at 1.5 mhz was typically about 1 db, that the preamp could not be significantly improved in noise performance. Later, however, it was found that it was contributing significant noise in the low frequency range. And so, a current mode preamp was tried, resulting in the same 1 db noise figure at the high end but a 12 db or more improvement in low end noise. However, since we believed that there was no inherent reason that current mode should be any different from voltage mode (except for eliminating high end resonance problems), we attempted to design a voltage mode preamp with the same front end as that used in the current mode preamp. This led to a very unique design and a patentable circuit (I. D. 1770) whose front end was identical to that of the current mode preamp and whose noise performance was also identical to that of the current mode preamp.

The result was that we had a new voltage mode design which was electrically and physically interchangeable with the old design but which had a 15 db or so improvement in low end noise. It makes some improvement in wideband noise in the H system but a greater improvement in the I system where the low end is 50 hz.

5.1 Version A (Modified FR-1400)

(See Schematic #1246526, Assy. or SP #1801588.)

Since time was limited and the FR-1400 preamp was believed to have excellent noise characteristics, it was decided to modify the FR-1400 preamp for use on the ES-200. The front end was kept the same but the gain of the amplifier was reduced, providing more feedback and therefore an amplifier with less distortion and wider bandwidth (flat within .5 db from 50 hz to 2.4 mhz). Then to bring the gain of the overall preamp back up to 40 db, another amplifier was added which drives the long lines to the electronics bays at the highest output level which will result from the worst case combination of tape speed, 20 db record overdrive, and high reproduce head output. Curves of distortion versus output level for several critical frequencies were measured and plotted (see D. Burri). Then families of head curves (both H and I) and curves of output versus record current overdrive were studied to find the most likely prospects for worst case distortion points. The fact that in regions where the head curves fall off, the equalizer boosts preamp distortion products was taken into account. In spite of the fact that in some cases the preamp output reaches 10 vptp, the worst case equalized distortion products were still about 50 db down, as they should be.

The first stage of the preamp is an emitter follower, Q1. To prevent emitter follower oscillations, a ferrite bead is used with the base lead passing through it twice (L1). This element has no effective impedance at frequencies in the passband of the recorder but it has a rising loss (resistance) at high frequencies where an emitter follower might oscillate. Thus it replaces the classical 100 ohm base resistor at high frequencies, but adds no noise in the passband.

The next three stages, Q2, 3, 4, form a standard series feedback amplifier where the gain is R6 plus R8 divided by R6. Bead L2 performs the same function for Q2 that L1 does for Q1.

Q5, 6, 7, 8, form another series feedback amplifier with a class A-B output stage Q7, 8, which helps keep power supply drain low. R28 and L5 in combination with the capacitance of the cable and the 1 K load resistance in the reproduce amp. form a low pass filter which is flat to about 3 mhz and cuts off thereafter. After much grounding and shielding work this was found to be the only effective means of stopping system oscillations which involve all 16 preamps and the cable. This appears to be a reliable cure, the only one we could find. It has a disadvantage, however, since the cable is part of the filter, the length of the cable must remain constant. This prohibits modifications to the machine which involve a cable length change without an accompanying change in the value of L5.

The series feedback amplifier will not be discussed in detail because it is so commonly used and understood.

5.2 Version B (Current Mode)

The current mode preamp upon which tests were made is given in Figure 4. It is a shunt feedback triple of the same basic type discussed in detail in Section 7. The special factor in the design was to make the high frequency feedback resistor R_{f1} enough larger than the R_p of the lead that the Norton equivalent noise current generator of R_{f1} is small

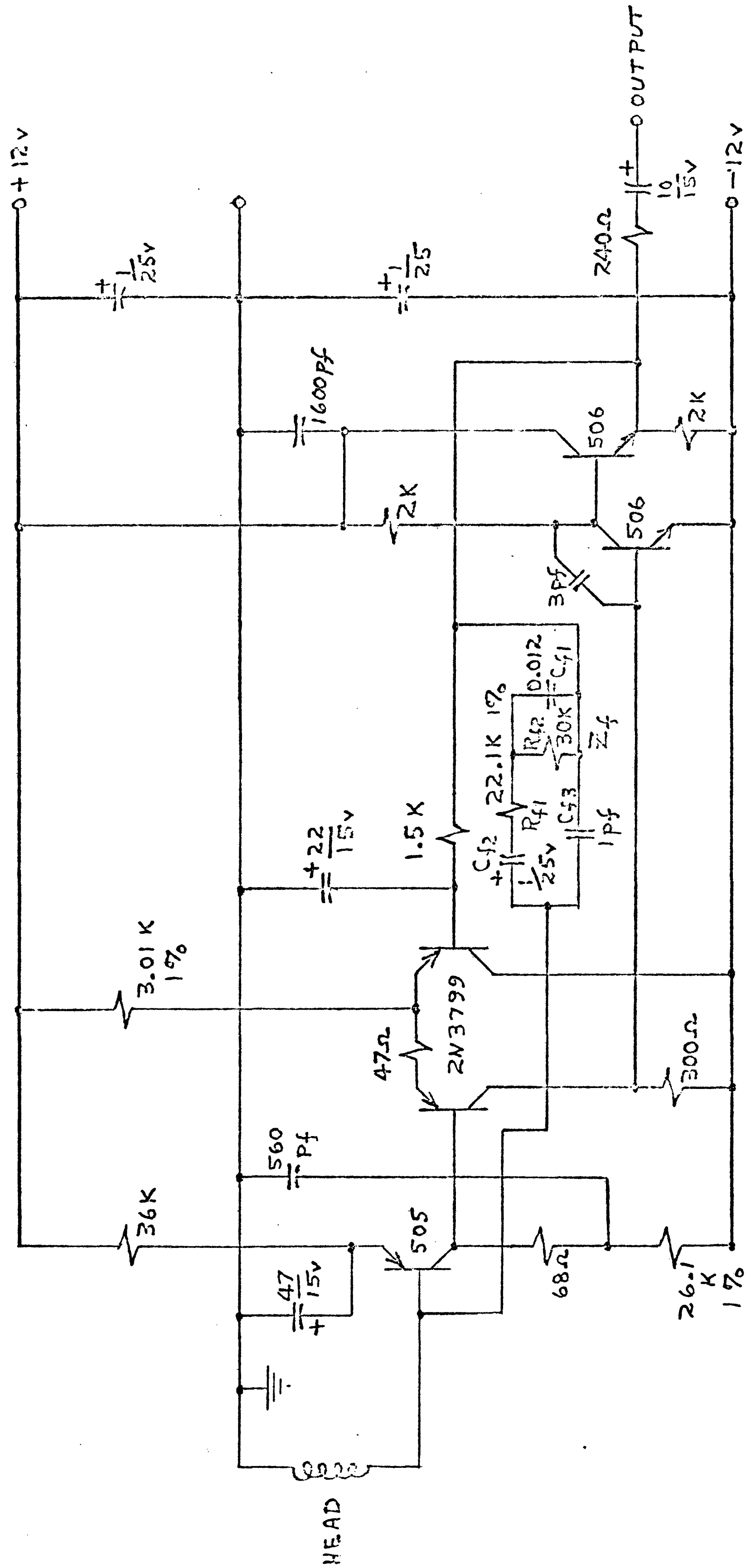

Figure 4. Current Mode Preamp.

compared to that of R_p. R_p is the equivalent parallel loss resistance of the head as measured on a bridge.

The shape of the frequency response of the transimpedance of the amplifier is directly proportional to the magnitude of Z_f. There is a low frequency pole-zero pair contributed by R_{f2} and C_{f1} which was convenient for matching the equalizer used to equalize the output of the preamp. The midband value of Z_f is therefore

$$Z_f = R_{f1} + \frac{1}{\frac{1}{R_{f2}} + pC_{f1}} \qquad \text{Eq. 15}$$

which simplifies to

$$Z_f = R_{f1} \frac{\left(\frac{1}{R_{f1}} + \frac{1}{R_{f2}}\right) + pC_{f1}}{\frac{1}{R_{f2}} + pC_{f1}} \qquad \text{Eq. 16}$$

The high frequency value of Z_f is

$$Z_f \Big|_{high} = R_{f1} = 22.1 \text{ K} \qquad \text{Eq. 17}$$

The low frequency value of Z_f is

$$Z_f \Big|_{low} = R_{f1} + R_{f2} = 52.1 \text{ K} \qquad \text{Eq. 18}$$

The zero of Z_f is

$$\text{zero} = \frac{1}{2\pi} \frac{1}{\frac{R_{f1} R_{f2} C_{f1}}{R_{f1} + R_{f2}}} = 1.04 \text{ khz} \qquad \text{Eq. 19}$$

The pole of Z_f is

$$\text{pole} = \frac{1}{2\pi} \quad \frac{1}{R_{f2}C_{f1}} = 443 \text{ hz} \qquad \text{Eq. 20}$$

An additional pole occurs at

$$\text{pole} = \frac{1}{2\pi} \quad \frac{1}{R_{f1}C_{f3}} = 7.2 \text{ mhz} \qquad \text{Eq. 21}$$

An additional zero occurs at

$$\text{zero} = \frac{1}{2\pi} \quad \frac{1}{(R_{f1}+R_{f2})\ C_{f2}} = 3.05 \text{ hz} \qquad \text{Eq. 22}$$

The straight line approximation to the frequency response of the trans-impedance of the current mode preamp is then given in Figure 5.

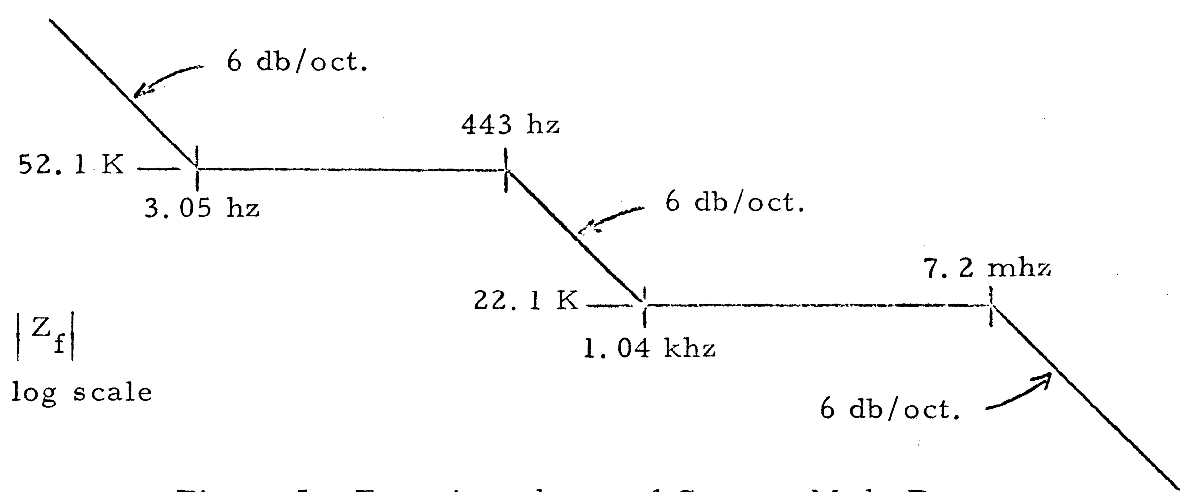

Figure 5. Transimpedance of Current Mode Preamp

This circuit is not an easy one to design (see Section 7), especially with the head connected directly across the input. The design in Figure 4 should not be taken as a production design; it is given here for information only -- satisfactory for testing purposes.

5.3 Version C (New Patented Voltage Mode)

Since it was believed that the front end arrived at for the current mode preamp was as nearly optimum as possible with present day devices, it was decided to try to design a voltage mode preamp using the same front end. The reason this first stage is better is that it has voltage gain, current gain, and power gain while the FR-1400 first stage (Version A) has only current gain and power gain (the emitter follower has no voltage gain).

Since the current mode preamp looks like a short circuit at its input, head resonance is eliminated because the head is shorted out. With the voltage mode preamp, resonance is still a problem so the circuit must be arranged so as to not increase the input capacitance through a Miller effect on the C_{ob} of the first stage. This implies that the collector of the first stage must operate into virtually a short circuit (as if it were an emitter follower). Furthermore, according to Middlebrook, the noise figure or signal-to-noise is independent of feedback. The trick is, therefore, to design the circuit for optimum noise performance by using considerable voltage gain in the first stage and then, with feedback loops, reduce the voltage gain of the first stage to as small a value as possible to eliminate the Miller capacitance. This requires two feedback loops, a shunt (virtual ground) loop to the collector of the first stage to reduce impedance there and a normal series feedback loop to the emitter of the first stage to stabilize the overall gain of the amplifier. But there is a special requirement on the shunt feedback loop. To understand this, we must consider the equivalent circuit of the first stage of the preamp.

The new voltage mode preamp., version C, is composed of two separate amplifiers as in version A. The output amplifier in version A drives the cable to the electronics tray and consists of the components to the right of coupling capacitor C5 in Schematic #1246526. This identical amplifier was used in the new preamp., version C. The first amplifier in version A (all components to the left of capacitor C5 in Schematic #1246526) is replaced in the new preamp., version C, by the amplifier given in Figure 6. It is not possible to refer to a drafting schematic at this writing, because the drafting package is not complete. The component designations in Figure 6 will, therefore, not correlate to those in the final drafting package.

In order to understand the requirements on the shunt feedback loop, we consider an equivalent circuit for the first stage Q1 in Figure 6. This equivalent circuit,[1] shown in Figure 7, is equivalent only to the input impedance seen looking into the base of Q1.

By an arguement similar to that used for computing the transfer function of a common emitter stage with an emitter resistor[2], we may say that

$$r_e = R_e + \frac{kT}{qI_e} \qquad \text{Eq. 23}$$

That is, r_e in Figure 7, is the sum of the emitter resistor R_e and the usual dynamic emitter resistance kT/qI_e of the transistor. Since the

1. Class Notes for Linear Wideband Transistor Amplifiers, by Ray Moore, Ampex, July 1965, p. 24.

2. Ibid, p. 26.

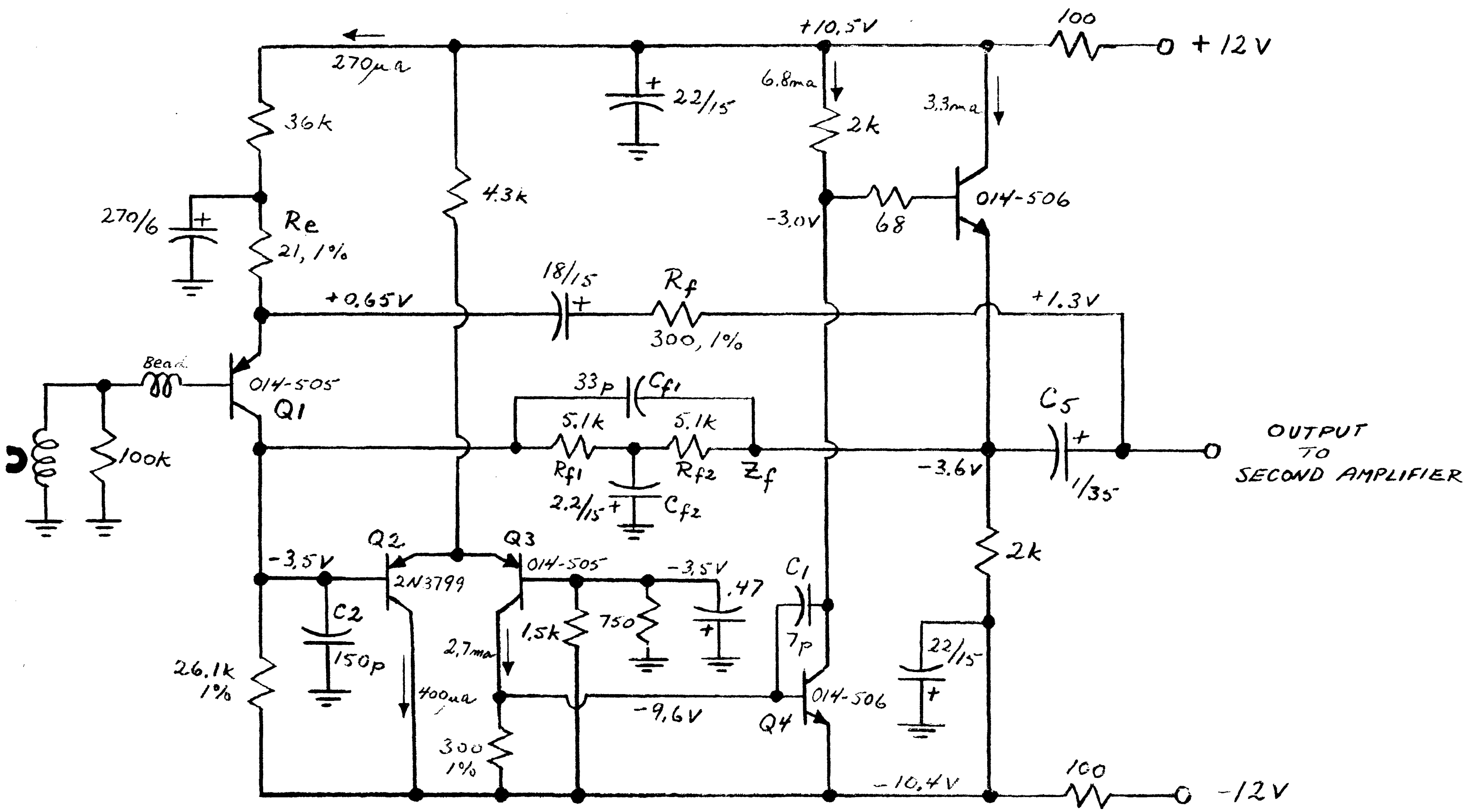

Figure 6. New Voltage Mode Preamplifier

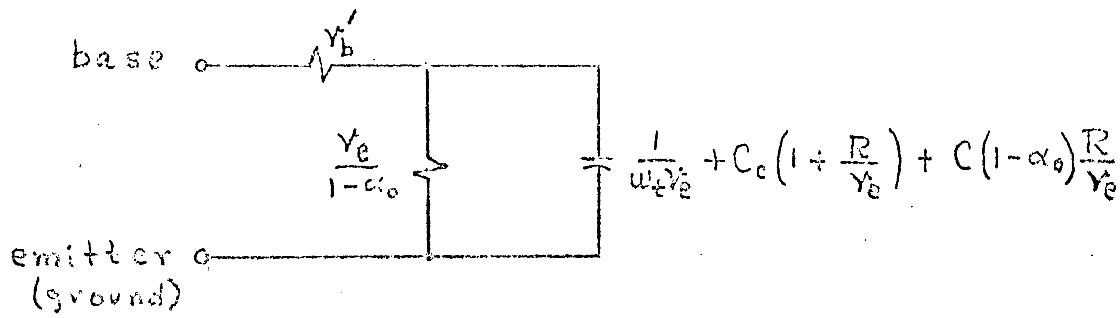

Figure 7. Equivalent Circuit of Input Impedance of Common Emitter Stage Q1.

emitter current I_e is 0.278 ma,

$$r_e = 21\Omega + \frac{27\Omega}{0.278} = 118\Omega \qquad \text{Eq. 24}$$

For the 014-505, at 0.278 ma, let us assume that $1-\alpha_o$ is 0.01, r'_b is 40 ohms, ω_t is 250 mhz, and C_c is 3 pf. Further, R and C are the resistance and capacitance which the collector of Q1 drives and are not yet known. A new equivalent circuit, Figure 8, may be drawn, putting in the known values.

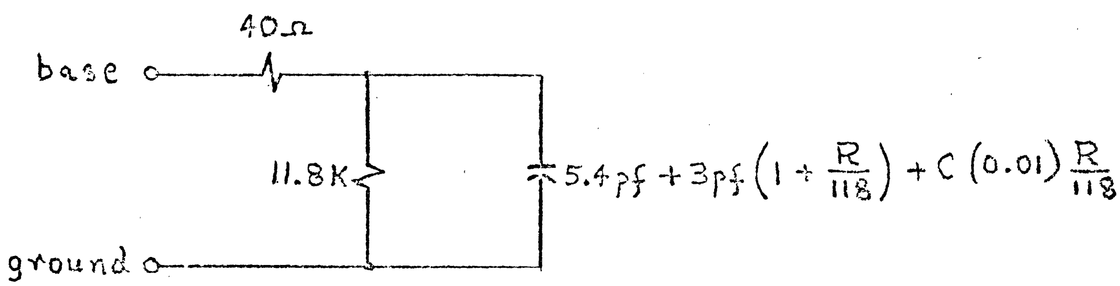

Figure 8. Equivalent Circuit of Input Impedance of Common Emitter Stage Q1.

Since we are concerned about head resonance at about 2 mhz, we must examine this circuit at that frequency. Assuming the total capacitance to be less than 80 pf, the capacitive reactance is greater than 1 kΩ; the 40 Ω r'_b is insignificant. We must now try to minimize the capacitance in order to keep head resonance high. Since I_e in Q1 was picked to optimize noise figure, there is not much one can do about the 5.4 pf. The controllable factors are, however, the second and third term in the expression for capacitance in Figure 8.

Examination of Figure 8 reveals a curious but crucial fact: the capacitance is <u>increased</u> by increasing C. This is curious because increasing C lowers the magnitude of the impedance driven by the collector of Q1 and, one would expect, therefore, lower the Miller capacitance of Q1. This result is not at all obvious -- only a careful analysis reveals it.

The conclusion, then, is that to minimize the capacitance seen looking into the base of Q1, one must keep C small (the $1-\alpha_o$ term helps) but, most important, the collector load impedance must be kept resistive and its R value small (equal to or less than 118Ω). Then the second term for capacitance in Figure 8, the true Miller term, is kept small and so also is the third term. This is the strange requirement placed upon the shunt feedback loop to the collector of Q1 in Figure 6: to present a small, resistive load to Q1. This requirement is satisfied in a strange way.

The amplifier, with the feedback network Z_f disconnected, is from the base of Q2 to the output emitter Q5, in the region of 2 mhz, a single time constant amplifier. An ac equivalent circuit of Q2, Q3 and Q4 is shown in Figure 9.

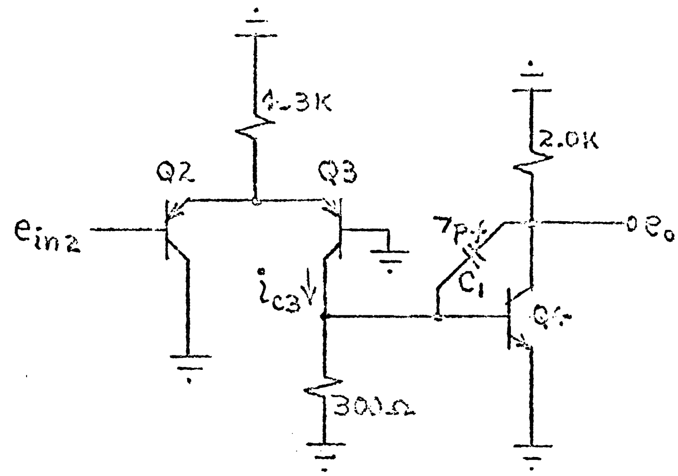

Figure 9. Equivalent Circuit of Q2, Q3, and Q4 for ac.

For an input voltage e_{in2} the ac current i_{c3} coming out of the collector of Q3 is, to a good approximation,

$$i_{c3} = \frac{e_{in2}}{r_{e2} + r_{e3}} \qquad \text{Eq. 25}$$

where r_{e2} and r_{e3} are the usual equivalent ac emitter resistors kT/qI_{e2} and kT/qI_{e3}. Since I_{e2} is 1.12 ma, r_{e2} is 24 ohms; since I_{e3} is 2 ma, r_{e3} is 14 ohms.

The Miller capacitance formed by C_1 plus the C_{ob4} of Q4 (total about 10 pf) predominates the current to voltage transfer of Q4. Very simply, and to a good approximation,

$$\frac{e_o}{i_{c3}} = \frac{-1}{p(C_1 + C_{ob4})} \qquad \text{Eq. 26}$$

where p is the complex frequency variable. Combining Eqs. 25 and 26,

$$\frac{e_o}{e_{in2}} = \frac{-1}{p(C_1 + C_{ob4})(r_{e2} + r_{e3})} \qquad \text{Eq. 27}$$

Since, in the shunt feedback impedance Z_f, C_{f2} is a short for ac, the only ac feedback is C_{f1}. The ac equivalent circuit for Q2, 3, 4, 5 and the feedback is therefore given in Figure 10 (forgetting C_2 for the moment).

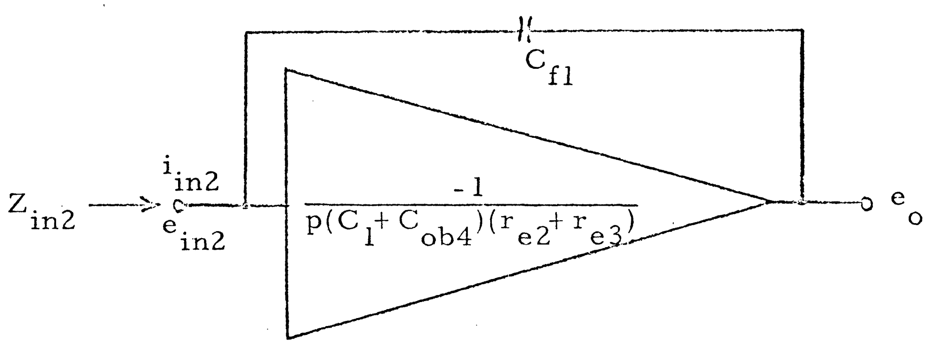

Figure 10. Equivalent Circuit for Q2, 3, 4, 5, and C_{f1}.

It is the input impedance Z_{in2} which we are interested in.

$$i_{in2} = (e_{in2} - e_o)\, pC_{f1} \qquad \text{Eq. 28}$$

and, substituting Eq. 27 for e_o

$$i_{in2} = e_{in2}\left[1 + \frac{1}{p(C_1 + C_{ob4})(r_{e2} + r_{e3})}\right] pC_{f1} \qquad \text{Eq. 29}$$

then

$$Z_{in2} = \frac{e_{in2}}{i_{in2}} = \frac{1}{pC_{f1}\left[1 + \frac{1}{p(C_1 + C_{ob4})(r_{e2} + r_{e3})}\right]} \qquad \text{Eq. 30}$$

Since Eq. 27 is much larger than one, we may drop the one in the denominator of Eq. 30:

$$Z_{in2} \cong \frac{p(C_1 + C_{ob4})(r_{e2} + r_{e3})}{pC_{f1}} = \frac{380 \text{ pf } \Omega}{33 \text{ pf}} = 11.5\,\Omega \qquad \text{Eq. 31}$$

Surprisingly, Z_{in2} is resistive and quite small. This is exactly what was desired: R in Figure 8 is Z_{in} or 11.5 ohms. The R/118 term becomes about 0.1. The second term in the capacitance expression becomes 3 x 1.1 or 3.3 pf which is very reasonable. Furthermore, C_2 plus the input capacitance 46 pf of Q2 (computed using Figure 7) is about 200 pf. The third term in the capacitance expression of Figure 8 for Q1 is therefore

$$C \times 0.01 \times \frac{R}{118} = 200 \times .01 \times 0.1 = 0.2 \text{ pf} \qquad \text{Eq. 32}$$

which is negligible.

The purpose of C_2 is to keep the shunt feedback loop from oscillating. It has no other function. It forms a capacitive divider to ground with C_{f1}, thereby reducing the feedback for all frequencies but, most important, at the gain crossover frequency.

The remaining problem is to find out whether the overall bandwidth of the amplifier is adequate. In the 2 mhz region we may draw an ac equivalent circuit for the whole amplifier with an input voltage e_{in} and an output voltage e_o (Figure 11). The transimpedance T_{r2} of the portion of the amplifier in Figure 10 (Q2, 3, 4, 5) is computed as follows. e_{in2} from Eq. 27 is substituted into Eq. 28:

$$i_{in2} = -e_o \left[p(C_1 + C_{ob4})(r_{e2} + r_{e3}) + 1 \right] pC_{f1} \qquad \text{Eq. 33}$$

Again the approximation is made that $p(C_1 + C_{ob4})(r_{e2} + r_{e3})$ is much less than one in the region of 2 mhz. Then

$$T_{r2} \triangleq \frac{e_o}{i_{in2}} \cong - \frac{1}{pC_{f1}} \qquad \text{Eq. 34}$$

An equivalent circuit for the whole amplifier is given in Figure 11 neglecting the effects of source impedance.

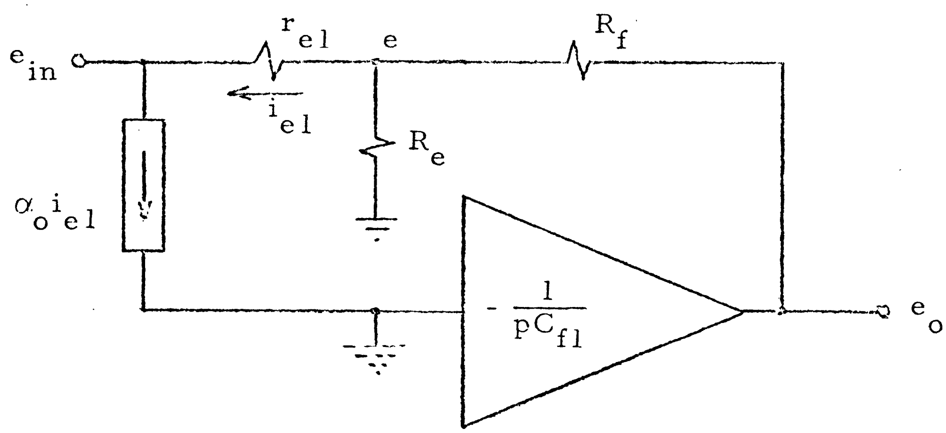

Figure 11. Equivalent Circuit of Amplifier

Writing a node equation for node e

$$\frac{e - e_{in}}{r_{e1}} + \frac{e}{R_e} + \frac{e - e_o}{R_f} = 0 \qquad \text{Eq. 35}$$

Also, assuming $\alpha_o \cong 1$,

$$i_{e1} = \frac{e - e_{in}}{r_{e1}} \qquad \text{Eq. 36}$$

and

$$\frac{e_o}{i_{el}} = - \frac{1}{pC_{f1}} \qquad \text{Eq. 37}$$

Combining Eqs. 36 and 37,

$$pC_{f1}\, e_o = \frac{e_{in} - e}{r_{el}} \qquad \text{Eq. 38}$$

Combining Eqs. 35 and 38 yields

$$\frac{e_o}{e_{in}} = \frac{R_e + R_f}{R_e} \quad \frac{1}{1 + pC_{f1}\, r_{el}\, R_f \left(\frac{1}{R_e} + \frac{1}{r_{el}} + \frac{1}{R_f} \right)} \qquad \text{Eq. 39}$$

The configuration has the usual low frequency gain for a series feedback loop.

$$\left. \frac{e_o}{e_{in}} \right|_{\text{low freq.}} = \frac{R_e + R_f}{R_e} \qquad \text{Eq. 40}$$

It also has a single high frequency pole at

$$f_{pole} = \frac{1}{2\pi C_{f1}} \quad \frac{1}{\frac{r_{el} R_f}{R_e} + R_f + r_{el}} \qquad \text{Eq. 41}$$

which for the values in Figure 6 makes the gain 15.3 and the pole frequency 2.82 mhz.

R_{f1} and R_{f2}, bypassed by C_{f2}, bias the output voltage to -3.5 v. The other dc voltages and currents are given in Figure 6.

5.4 Comparison of Performance

The performance of the three different types of preamps may be compared by examining Figures 12, 13and 14. In each figure, the equalized output is plotted along with the 200 cycle slot noise as a function of frequency with tape in motion and at rest. The original ES-200 preamp., modified from the FR-1400, is called version A and its curves are plotted in Figure 12. The current mode preamp, discussed in Section 5.2, is called version B and its curves are plotted in Figure 13. The new voltage mode preamp., discussed in Section 5.3 is called version C and its curves are plotted in Figure 14.

The noise figure of each preamp was measured, with the same head as a source, and was 1 db or less in all cases at 1.5 mhz. Therefore, since most of the noise is from the head at 1.5 mhz, the signal-to-slot-noise for all three should be identical. With the tape stopped, the signal-to-slot-noise is 69.5 db for A, 67 db for B, and 69 db for C. A and C are equal within experimental error but the current mode preamp., version B, seems to be about 2 db worse than the other two.

At 500 khz, since there is a rather large difference between the noise curves for tape moving and stopped, the noise is primarily tape generated. Therefore, the signal-to-slot-noise with tape moving should be identical for all three cases. It is 77 db for version A, 75 db for version B, and 77 db for version C. Again the current mode, version B, is 2 db worse than the other two. It appears that a 2 db error has come into the current mode measurement. This may have been caused somehow by the equalizer, or by a slight error in setting up record current.

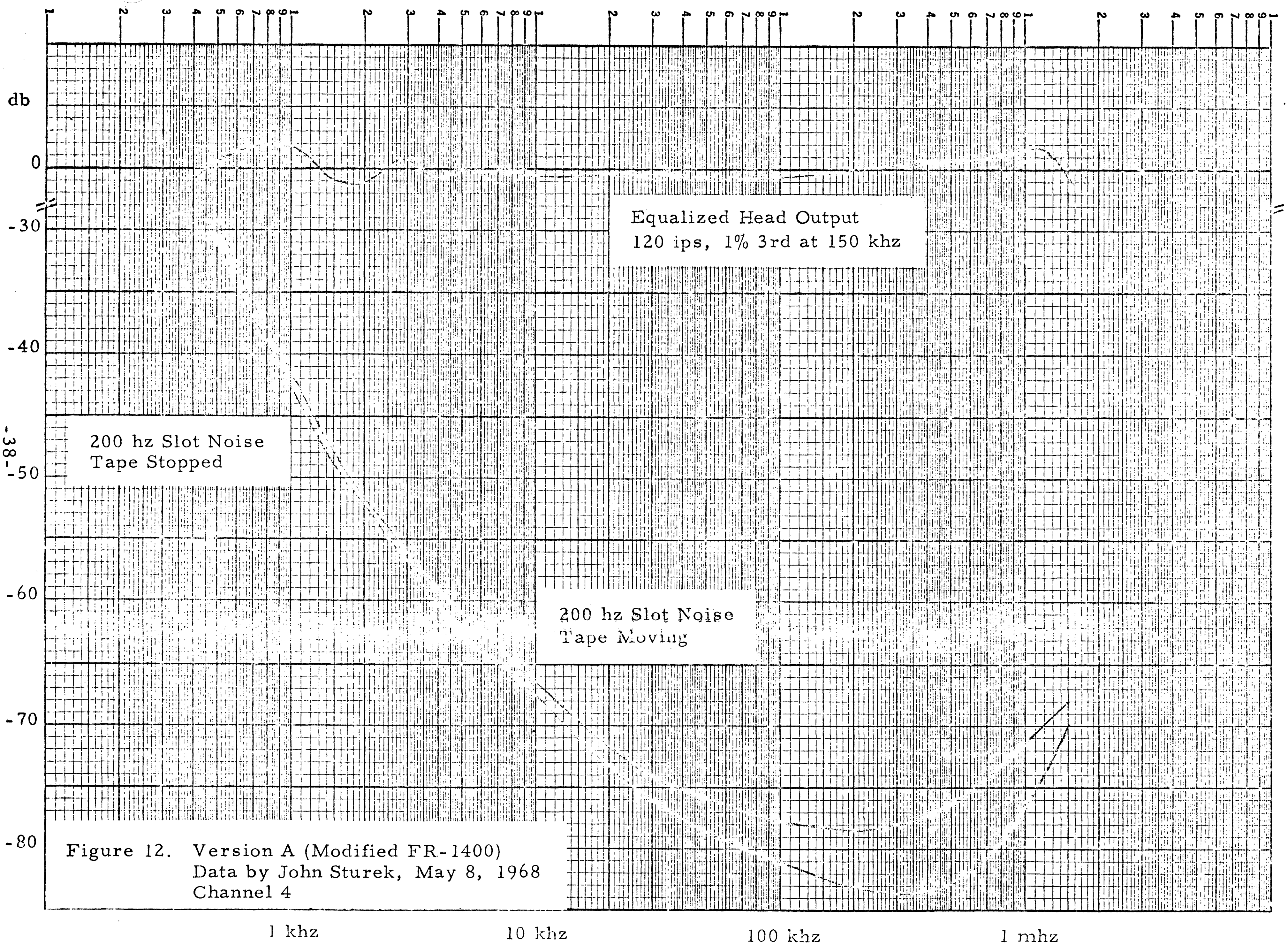

Figure 12. Version A (Modified FR-1400)
Data by John Sturek, May 8, 1968
Channel 4

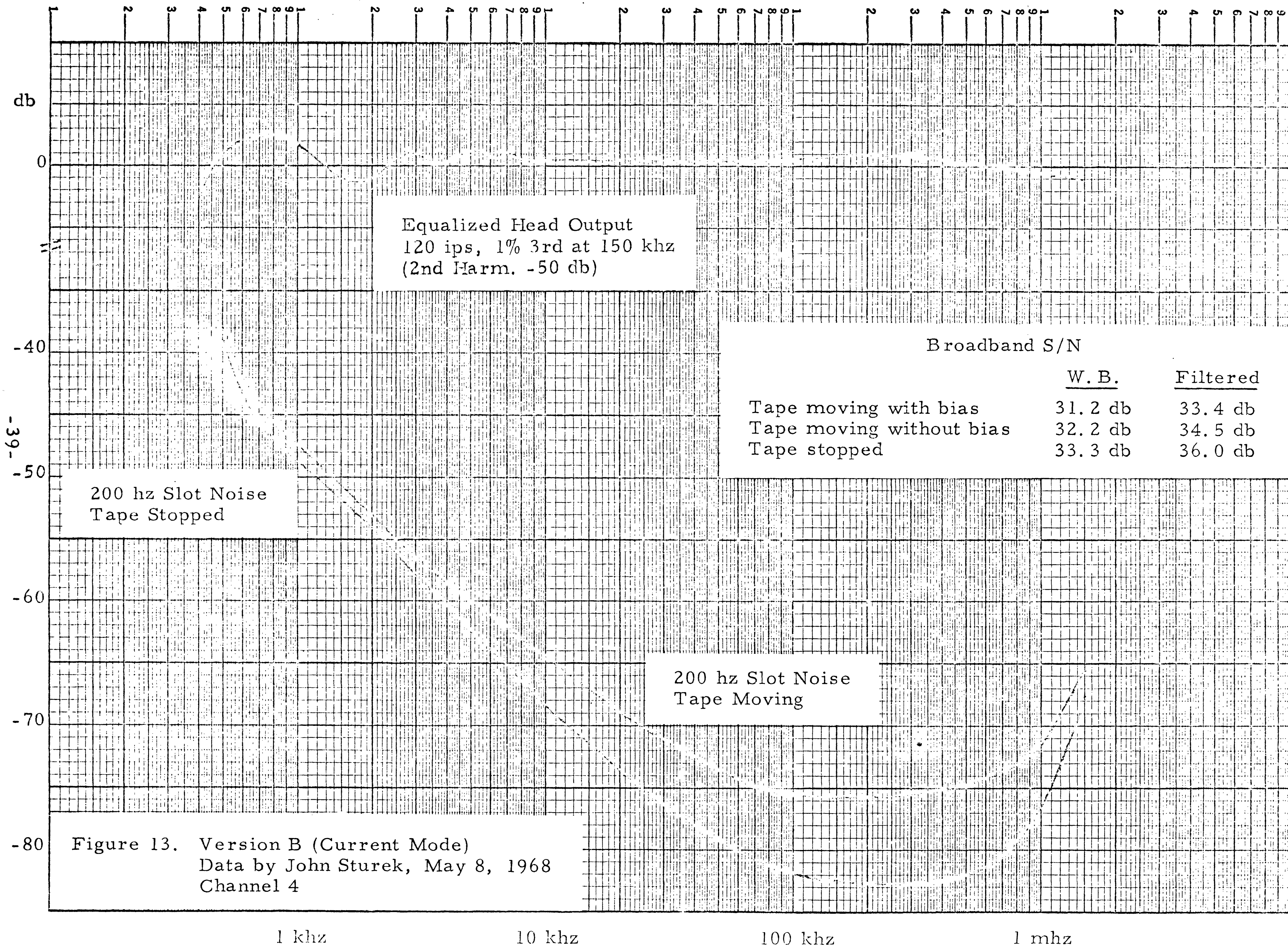

Figure 13. Version B (Current Mode)
Data by John Sturek, May 8, 1968
Channel 4

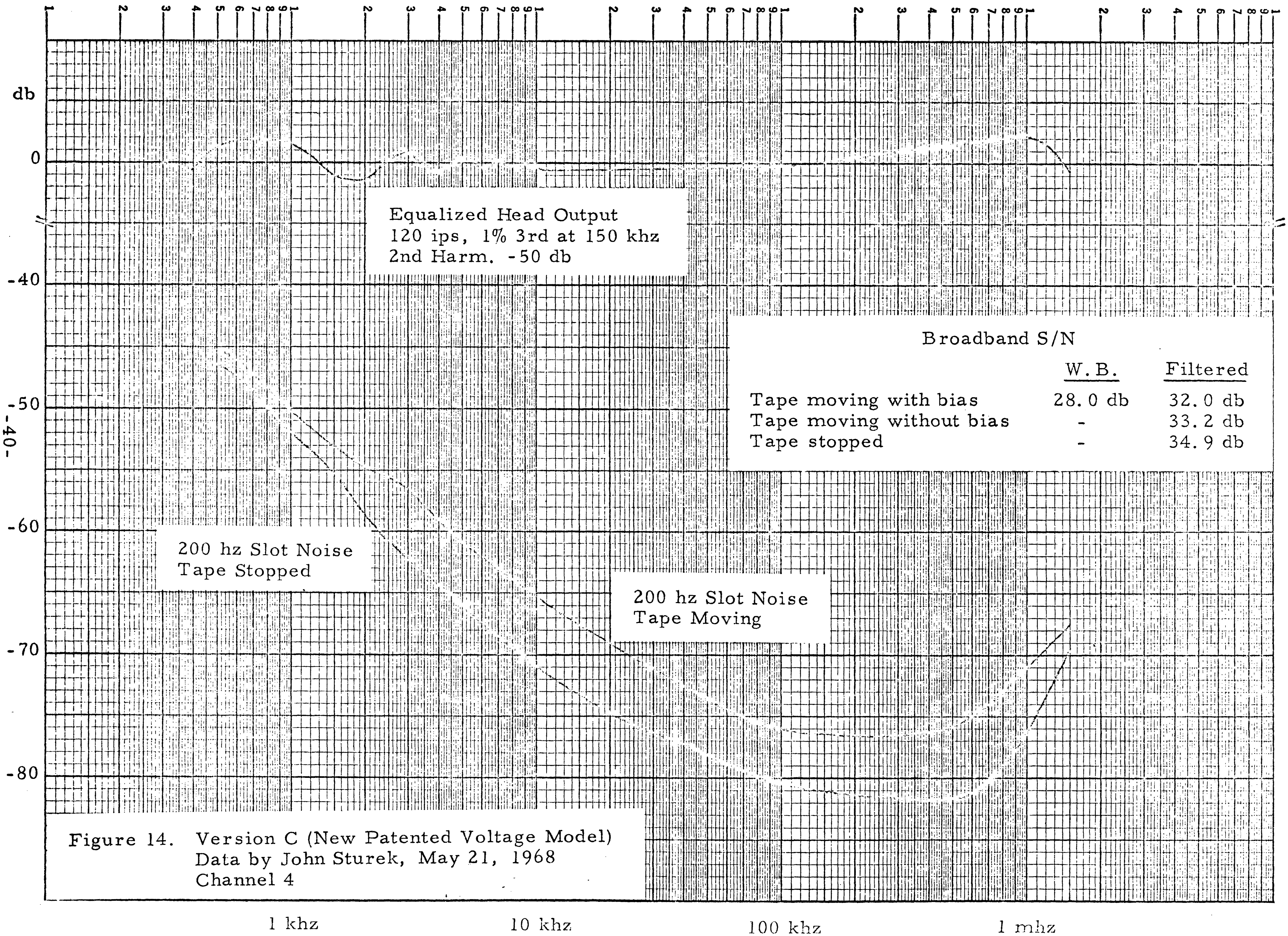

Figure 14. Version C (New Patented Voltage Model)
Data by John Sturek, May 21, 1968
Channel 4

However, the significant differences between the preamps. occur at the low frequency end of the spectrum. At 700 hz (the curves may not be reliable at 500 hz due to frequency drift in the HP 410A), the signal-to-slot-noise with the tape stopped is 39 db for version A, 48 db for version B, and 50.5 db for version C. As expected the new voltage mode preamp., version C, is 11.5 db better than the old voltage mode, version A. And if we can add the seemingly consistent 2 db error to the reading for the current mode, version B, it is within 0.5 db of the new voltage mode, version C, and 11 db improved over the old voltage mode, version A. The magnitude of the improvements is increasing as frequency goes down.

It is curious that in spite of the fact that the slot signal-to-noise for the current mode is consistently 2 db lower than that for the new voltage mode, the broadband signal-to-noise is 1 db better for the current mode with the tape stopped. This is explained, I believe, by the fact that the equalizer for the current mode falls off more steeply above 1.5 mhz than for the voltage mode. But it does tend to refute the 2 db discrepancy in signal-to-slot-noise.

I believe that the above data, plus the noise figure measurements, prove that there is no essential difference in signal-to-noise performance between the voltage and current modes (Hewlett-Packard alleges the current mode to be superior). The data also shows that the new voltage mode preamp. is a significant improvement over the old voltage mode preamp. in the low frequencies.

5.5 General Theory of Noise Figure and Its Application to Reproduce Heads

Terman[3] defines noise figure as

$$F = \frac{\text{signal-to-noise power ratio of ideal system}}{\text{actual signal-to-noise power ratio of output}} \qquad \text{Eq. 42}$$

By "ideal system" he means one in which the source (see Figure 15) contributes noise but the amplifier does not. That is, the amplifier acts in all respects like the actual amplifier except that it contributes no noise. Since the signal and load resistance is, for any given preamp., the same in both numerator and denominator, we may rewrite Eq. 42 .

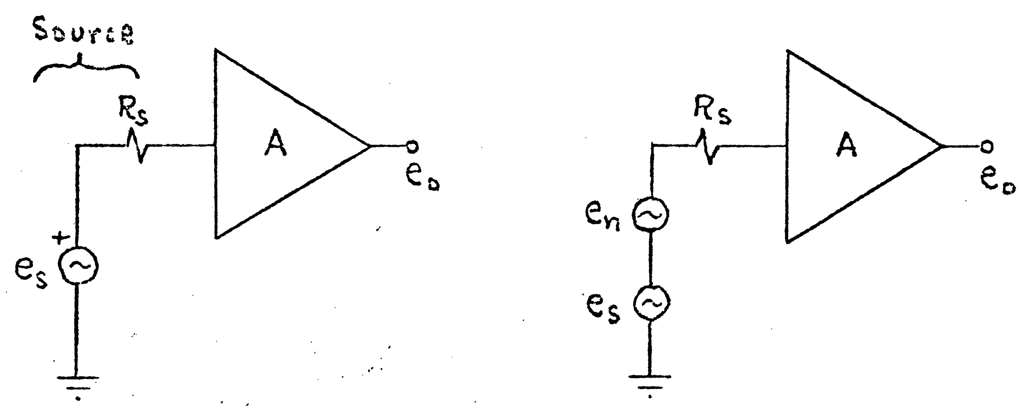

Figure 15. Noise Figure Block Diagram

$$F = \frac{(\text{actual output noise voltage})^2}{(\text{output noise voltage of ideal system})^2} \qquad \text{Eq. 43}$$

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3. Electronic and Radio Engineering, Frederick Emmons Terman, 4th Edition, McGraw-Hill, 1955, p. 440.

The noise figure F is usually given in db and, since it is a ratio of powers or voltage squared, one must remember to take ten times the $\log_{10} F$ rather than twenty.

In Figure 15, an equivalent circuit model is used for the source: a Thevenin equivalent voltage generator e_s in series with its Thevenin equivalent output impedance R_s. But, in the case of the reproduce head, matters are not so simple. The head acts as a current generator i_p with its own inductance and resistance shunting this generator. This equivalent circuit is given in Figure 16.

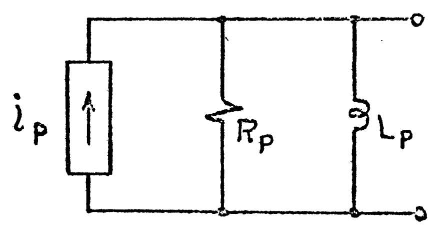

Figure 16. Current Equivalent Circuit of the Head

The current generator i_p is proportional to the remnant magnetization on the tape as it passes the head. It is therefore, except for losses, proportional to the record current and therefore, except for pre-emphasis, proportional to the record input voltage. There may also be a parallel capacitance C_p, but hopefully this will not be significant in the frequency range of interest. L_p may be relatively constant with frequency but R_p decreases with increasing frequency.

In working with a voltage preamp., it is convenient to put the circuit into an equivalent series form. This may be done at one frequency only -- that frequency at which the measurement of R_p and L_p are made. The single-frequency Thevenin equivalent of the circuit given in Figure 16 is derived by first computing the open-circuit impedance for real frequencies looping into the output terminals.

$$Z = \frac{R_p \, j\omega L_p}{R_p + j\omega L_p} \qquad \text{Eq. 44}$$

Then

$$Z = \frac{R_p \, j\omega L_p}{R_p + j\omega L_p} \quad \frac{R_p - j\omega L_p}{R_p - j\omega L_p} \qquad \text{Eq. 45}$$

And simplifying

$$Z = \frac{R_p (\omega L_p)^2}{R_p^2 + (\omega L_p)^2} + j \frac{\omega L_p (R_p)^2}{R_p^2 + (\omega L_p)^2} \qquad \text{Eq. 46}$$

The Thevenin open circuit voltage is (for one frequency)

$$e_s = i_p Z = i_p \frac{R_p \, j\omega L_p}{R_p + j\omega L_p} \qquad \text{Eq. 47}$$

If the Q is large (5 or so) we may simplify

$$e_s \cong i_p \, j\omega L_p \qquad \text{Eq. 48}$$

which shows the normal rising 6 db/oct. response in open circuit head voltage output.

We may now form a new series equivalent circuit (Figure 17), where e_s is given in Eqs. 47 and 48 and R_s and L_s are deduced from

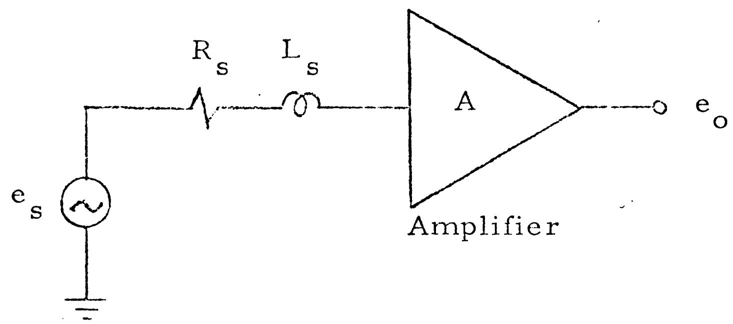

Figure 17. Series Equivalent Circuit of Head and Preamp.

Eq. 46 (for one frequency only).

$$R_s = \frac{R_p(\omega L_p)^2}{R_p^2 + (\omega L_p)^2} \qquad \text{Eq. 49}$$

and

$$L_s = \frac{L_p R_p^2}{R_p^2 + (\omega L_p)^2} \qquad \text{Eq. 50}$$

Then, inserting the noise generator which accompanies R_s, we have Figure 18 where

$$e_n^2 = 4kTBR_s \qquad \text{Eq. 51}$$

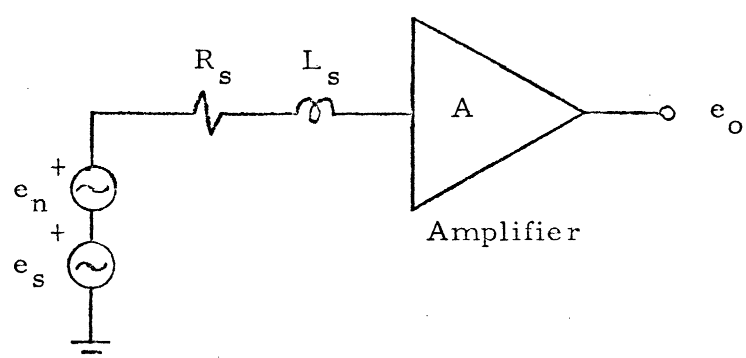

Figure 18. Series Noise Equivalent Circuit of Head and Preamp.

and where k is Boltzman's constant, T is absolute temperature in degrees Kelvin, B is the noise bandwidth and R_s is given in Eq. 49.

To find the noise at the output in Terman's ideal system one then measures R_p and L_p, computes R_s, measures the gain from e_n to e_o, computes Eq. 51 and multiplies e_n^2 by the gain squared. It may be possible to measure R_s and L_s directly -- this depends upon the type of bridge available. The gain is easily measured by inserting a small resistor in series with the head and inserting a current into the resistor. The gain A is then the ratio of e_o to the value of voltage across the resistor. We may now compute the noise figure F.

$$F = \frac{(\text{actual output noise voltage})^2}{4\,kTBR_s\,A^2} \qquad \text{Eq. 52}$$

The actual output noise voltage is measured with a voltmeter having bandwidth B (a slot meter, usually). It must be remembered that if the meter is an average reading meter the magnitude of true RMS noise is 1 db larger than that indicated by the average reading meter. Therefore, 1 db must be added to the noise figure to make this correction.

In the case where one is dealing with a current mode preamp., it is perhaps easier to use the parallel form of the equivalent circuit (Figure 19), where

$$i_n^2 = \frac{4\,kTB}{R_p} \qquad \text{Eq. 53}$$

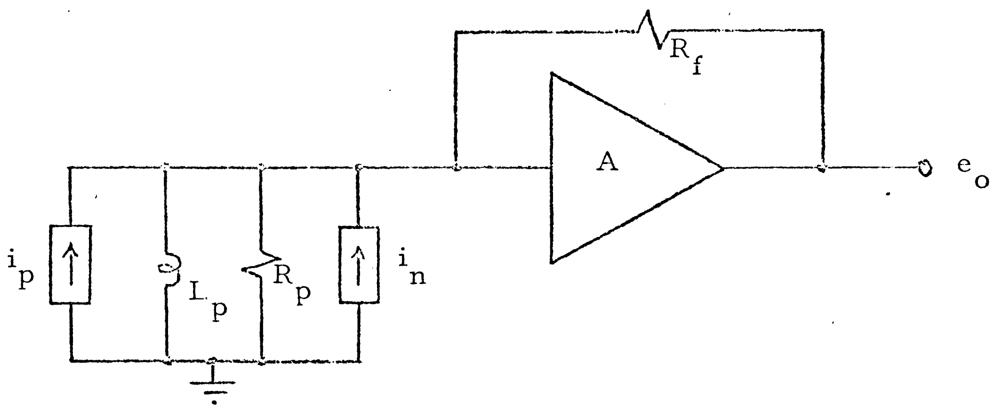

Figure 19. Parallel Noise Equivalent Circuit of Head and Preamp.

Now the transimpedance T_r of the amplifier (approximately R_f) is measured by inserting a known current tnrough a resistor much larger (one hundred times) than the input impedance of the amplifier. The noise figure F is then

$$F = \frac{(\text{actual output noise voltage})^2}{\frac{4kTB}{R} \; T_r^{\;2}} \qquad \text{Eq. 54}$$

remembering the correction for the average reading meter.

Some statements may be made about preamp. design. In both the voltage mode and current mode amplifier it is essential that any resistor which connects to the amplifier input be five or ten times larger than R_p because such resistors directly degrade the noise figure. This includes bias resistors and in the case of the current mode preamp. the feedback resistor R_f. The rule does not apply to the impedance looking into the base (input lead) of the first stage of the amplifier. The noise performance of this stage is governed by an equation such as that for resistive sources given by Schwartz[4] which does not imply that the impedance looking into

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4. "Information Transmission, Modulation, and Noise," Mischa Schwartz, McGraw-Hill, 1959, p. 253, Eq. 5-147.

the transistor be much higher than the Norton Equivalent R_p of the source resistance. This is a commonly held misconception among preamp. designers!

Furthermore, the noise figure is the most useful measure of preamp. performance at the high frequency end of the spectrum. The tape normally does not make a contribution here, so it is a question of comparing the preamp. to the head, and this is what the noise figure does. The main contribution to broadband noise (head and preamp.) is at the high end. If the design is not a proper one, however, the preamp. may make a significant contribution to broadband noise from the low end of the spectrum.

6. REPRODUCE AMPLIFIER

(See Schematics 1246501, 1246503, and 1246505)

The ES-200 reproduce amplifier was taken from the design done for the Lem-Appollo system done by Ray Moore and Maung Gyi during the summer of 1965. The choice of values for the different equalizer components, which is the heart of the design, was made by Rich Smith during the ES-200 program. During this development, Rich wrote a program for the Tyme-Share Computer which automatically produces equalizer values when given certain parameters of the head curve to be equalized. This was an invaluable aid, since it occurs that the most changing and unpredictable factor in an instrumentation electronics development is the shape of the head curves. The fact that, in production, the equalizers actually equalize the reproduce heads, seems little short of miraculous.

6.1 Amplitude Equalizer Design

Since the equalizer design was primarily Rich Smith's, I will not discuss it in great detail, except to point out where the biggest difficulties were encountered. The form of the amplitude equalizer is not basically unique. (See Figure 20.) The only unusual components are C_5 and C_{R1}. They permit C_{39}, the element which determines the rising 6 db/oct. portion of the curve, to be common to all seven equalizers, while at the same time permitting R_1, the element which determines the low end 3 db frequency, to be different for different equalizers.

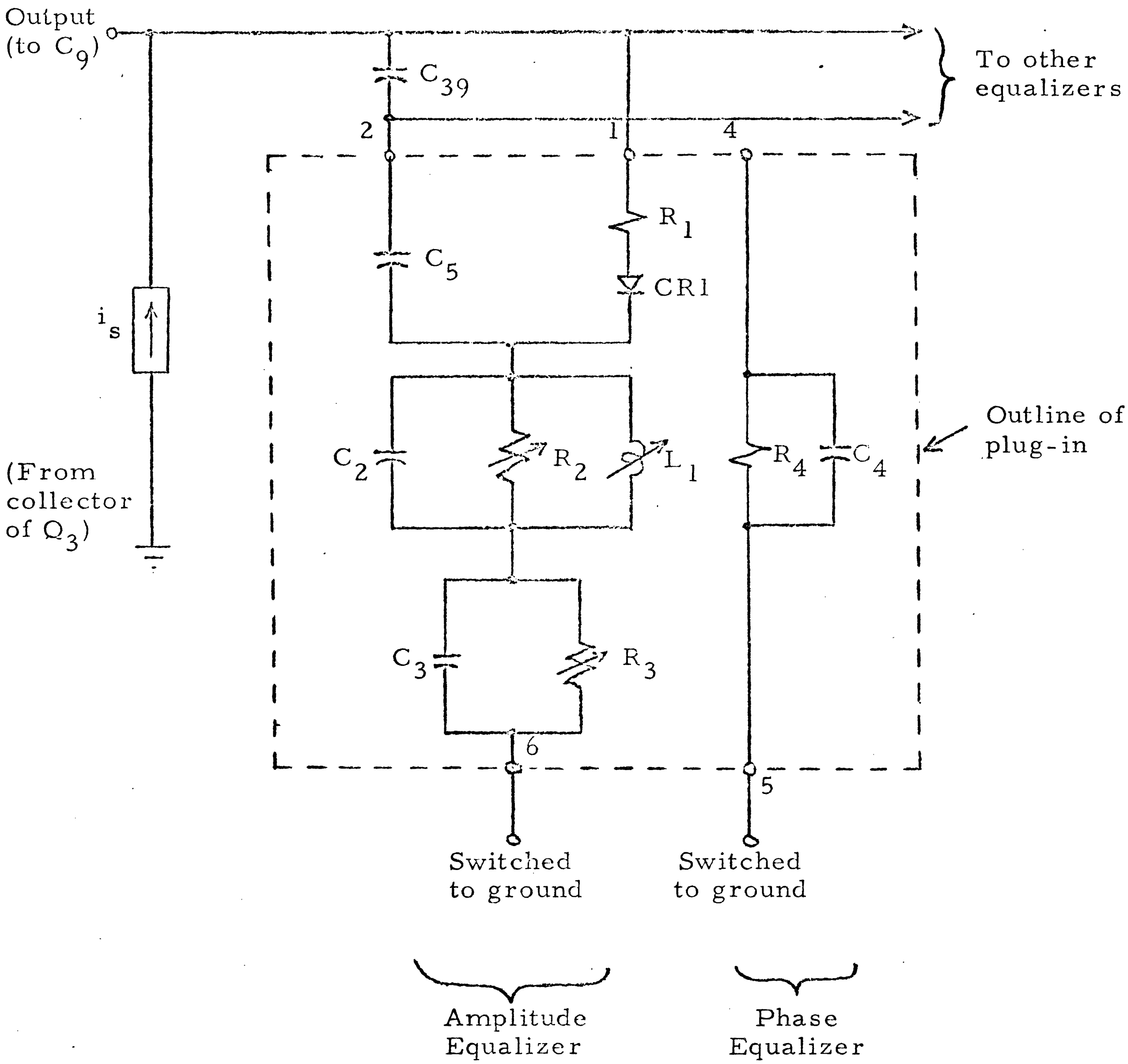


Figure 20. Equalizer

C_5 is a coupling capacitor; CR1 blocks the ac signals from reaching the bottom of C_{45} from any but the actuated equalizer. Therefore, only the R_1 from the actuated equalizer is paralleled with C_{45}.

R_3, which is adjustable, shapes the equalizer response to fit the maximum of the head curve (which may occur anywhere from one-sixth to one-tenth of the upper band edge frequency). C_3 helps remove high frequency noise. The upper band edge is equalized by the rising impedance of the tuned circuit L_1, C_2 and R_2. Since C_2 is not adjustable, L_1 must be adjusted to tune the circuit to the upper band edge frequency. Then R_2 is adjusted to give the right response at that point.

The problem with this arrangement is that there is no control available in the region of two-thirds the band edge frequency, but some variation in head output is encountered there which does not correlate with the shape of the curve at one-tenth the upper band edge and at the upper band edge. If C_2 was adjustable, L_1 could be adjusted for the two-thirds maximum frequency, and the circuit then resonated by adjusting C_2. But there is no room for a larger capacitor and there are already 294 equalizer adjustments in a 14 channel system. It was to solve these problems that patent ID #1766, "Electrically Speed Switchable Direct Reproduce Equalizer," and patent ID #1767, "Switchable Resistance/Inductance Multiplier" were filed by Ray Moore (they are not part of the ES-200 system).

6.2 System Considerations (See Schematic 1246501)

A block diagram of the ES-200 Reproduce Amplifier is given in Figure 21. There are two basic configurations of the reproduce amplifier, the H system and the I system. The basic parameters are given in Table 3. The equalizers are plugged in; therefore, they determine the bandwidth of

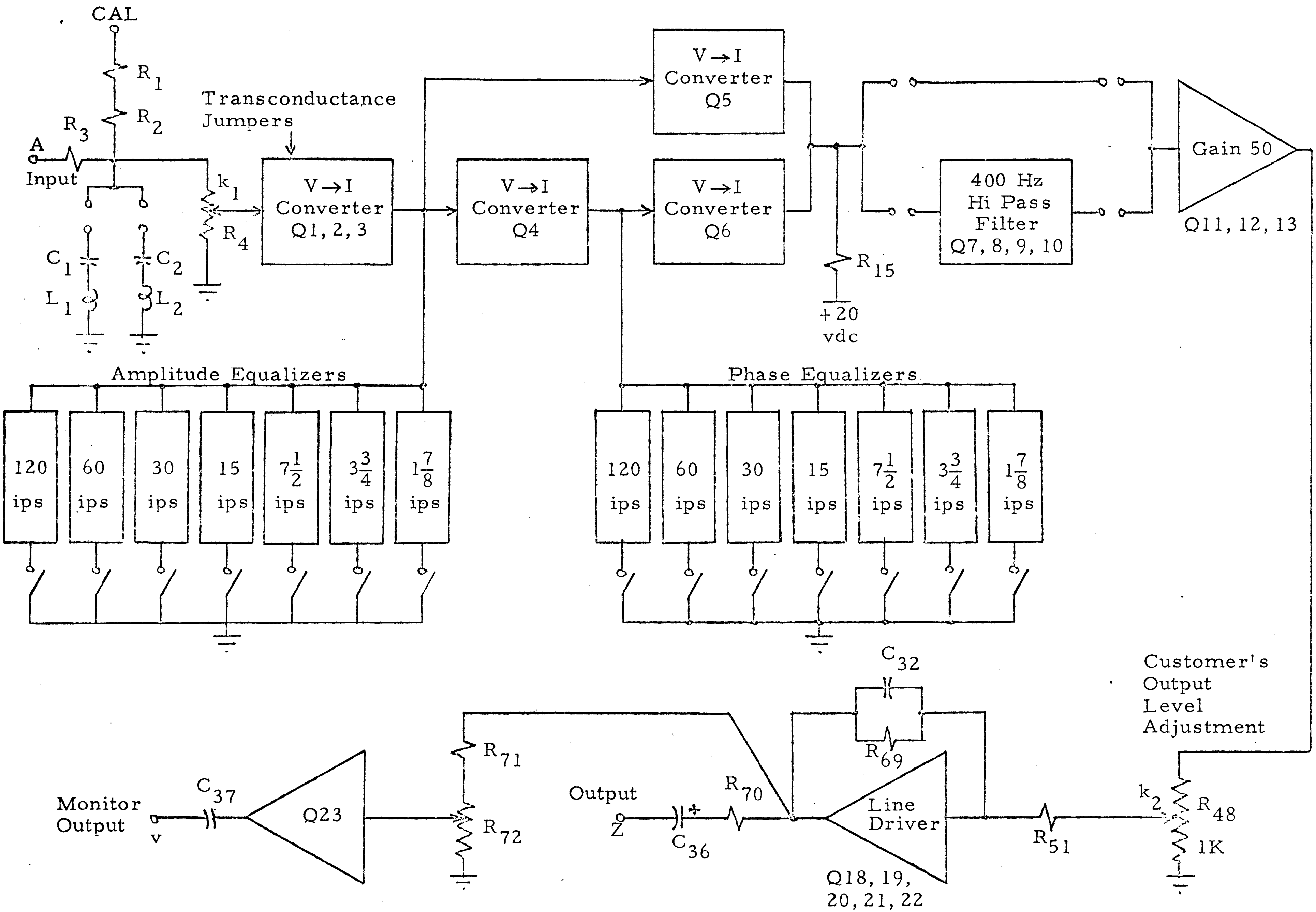


Figure 21. ES-200 Reproduce Amplifier Block Diagram

	H	I
Bias Frequency	7.7 mhz	3.4 mhz
Bandwidths (120 ips)	1.5 mhz, 2 mhz, (2.4 mhz)	600 khz (750 khz)
400 hz High Pass Filter	Yes	Normally No (optional)

Table 3. The Two Configurations of the Reproduce Amplifier

the system. Some jumpering is required also. The proper bias trap must be jumper selected to match the bias frequency. The g_m of the first $V \rightarrow I$ converter is jumper selected to match the output of either the H or I head. The 400 hz high pass filter may be jumpered in or out (it is required for H system performance).

Potentiometer R_4 is adjusted to normalize the outputs of the various heads. The first $V \rightarrow I$ converter is a standard series feedback amplifier except that the output is taken as the collector current of emitter follower Q3, rather than its emitter voltage as is normally the case. Such an elaborate amplifier is required because of the tremendous variety of signal levels coming out of the preamp. Since the highest level signals are at about the right frequency to put distortion products at the peak of the equalizer curve (+20 db, for example), it is imperative that this amplifier have exceptionally low distortion.

However, once the signal is equalized, it may be processed through single transistor stages (Q4, 5, 6) at a low level (4 or 8 mvrms normally) without introduction of distortion products. It is then taken up to 200 mvrms by the simple series feedback amplifiers Q11, 12, and 13. A 6 db loss is normally taken at the customer's level control R_{48}, dropping the signal to 100 mvrms. The line driver has a gain of 20, bringing the signal to 2 vrms which is reduced to 1 vrms by the divider formed by the terminated 75 ohm output line and R_{70}. The normal output to the customer is, therefore, 1 vrms with the pot set at half way. Once the reproduce amplifier is set up in this way, the input (and therefore the output) may be raised 10 db without limiting or significant distortion. This capability is provided primarily by the special shunt feedback line driver, which will be discussed in detail in Section 7. This amplifier is basically the same as the head driver amplifier and also the impedance converting amplifier in the record amplifier.

The gain through the system is given in Eq. 55.

Reproduce Amplifier Gain

$$= \frac{R_4}{R_3+R_4} \quad k_1 \quad \underset{(\text{or } R_{11})}{\frac{1}{R_9}} \quad \left|Z_A\right| \quad \left|\frac{Z_\phi}{R_{20}} - 1\right| \quad \frac{R_{43}+R_{39}}{R_{39}} \quad k_2 \quad \frac{R_{69}}{R_{51}} \quad \left|\frac{Z_L}{R_{70}+Z_L}\right| \qquad \text{Eq. 55}$$

$$= \quad 0.91 \quad k_1 \quad \frac{1}{75\,\Omega} \quad Z_A \quad \underset{(\text{low freqs.})}{0.5} \quad 52 \quad 0.5 \quad 20 \quad 0.5 \qquad \text{Eq. 56}$$

where $|Z_A|$ is the magnitude of the amplitude equalizer impedance, k_1 is the gain of pot R_4, Z_ϕ is the phaze equalizer impedance, k_2 is the loaded gain of pot R_{48}, and Z_L is the load impedance.

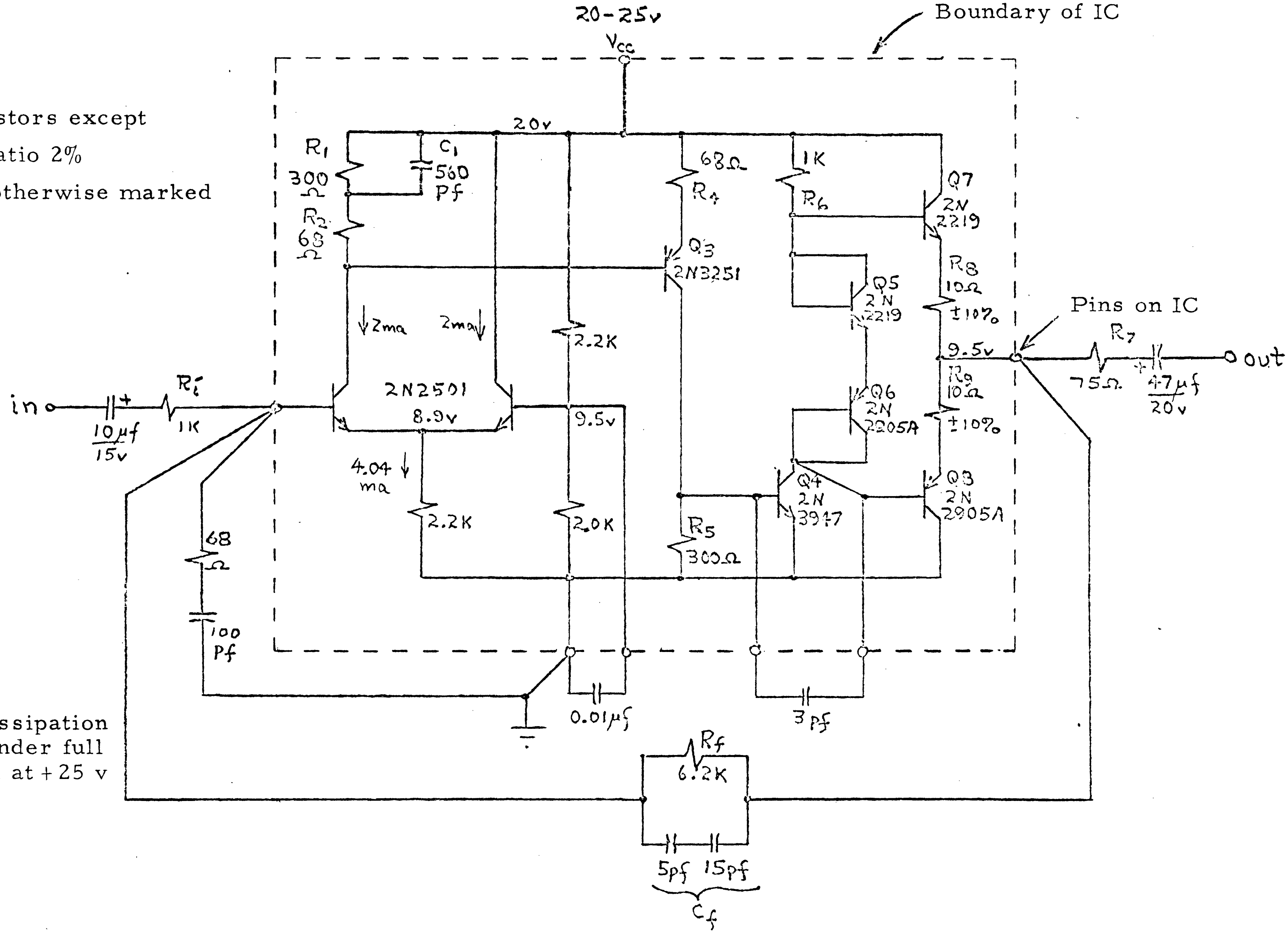
Phase correction is accomplished by the usual left-half-plane-pole -- right-half-plane-zero method. The phase correcting capacitor C_4 may be experimentally chosen to minimize group delay or to optimize pulse response. The two are not usually the same.

7. THE SHUNT FEEDBACK TRIPLE

It is this amplifier, used in three places in each direct channel, which makes possible some of the unique features of the ES-200 system. In the record amplifier, it makes it possible to simultaneously provide a constant 20 KΩ input impedance, a 50 hz - 2.4 mhz bandwidth, a 20 db overdrive capability, and a record level adjustment. In the record driver it makes it possible to provide the head with up to 150 ma peak-to-peak at an output voltage of 15 v ptp with negligible distortion from a single -20 v power supply with class A-B efficiency. In the reproduce amplifier, it makes it possible to drive a 75 ohm line from a 75 ohm source impedance at 10 db over 1 vrms. With a standard series feedback amplifier these things cannot be done.

As can be seen from an examination of each of these three amplifiers, they are each biased quite differently. This fact is a result of the evolutionary process which the design passed through -- the last version being the best from the standpoints of cost and simplicity. The last ES-200 version is the head driver amplifier; it should be noted that there are no emitters bypassed and therefore a minimum of tantalum which is the most expensive circuit ingredient. The final version of this amplifier was done in anticipation of having a hybrid integrated circuit made; the schematic for this version is given in Figure 22. The main new features of this design are the use of transistors Q5 and Q6 in place of diodes 013-677 and the use of a 2N3947 in place of an 014-506 for Q4 to obtain a 25 v B_{VCEO}. The 2N2501 is an 014-506; the 2N3251 is an 014-505; the 2N2219 is an 014-247; the 2N2905A is an 014-364.

5% resistors except
$\frac{2.2K}{2.0K}$ ratio 2%
unless otherwise marked



Total dissipation
1 watt under full
load and at +25 v

Figure 22. Hybrid Shunt Feedback Triple

In all respects, at the frequency of gain crossover, all four versions of this amplifier are basically the same. That is, the ac equivalent circuits are all nearly identical. Since the biasing is relatively straightforward (different in each version), I will not discuss it except to say that the dc bias is <u>very stably</u> accomplished through the ac feedback resistor R_f. We may then proceed to the more complex question of stability from oscillation and consider the high frequency equivalent of the amplifier given in Figure 23.

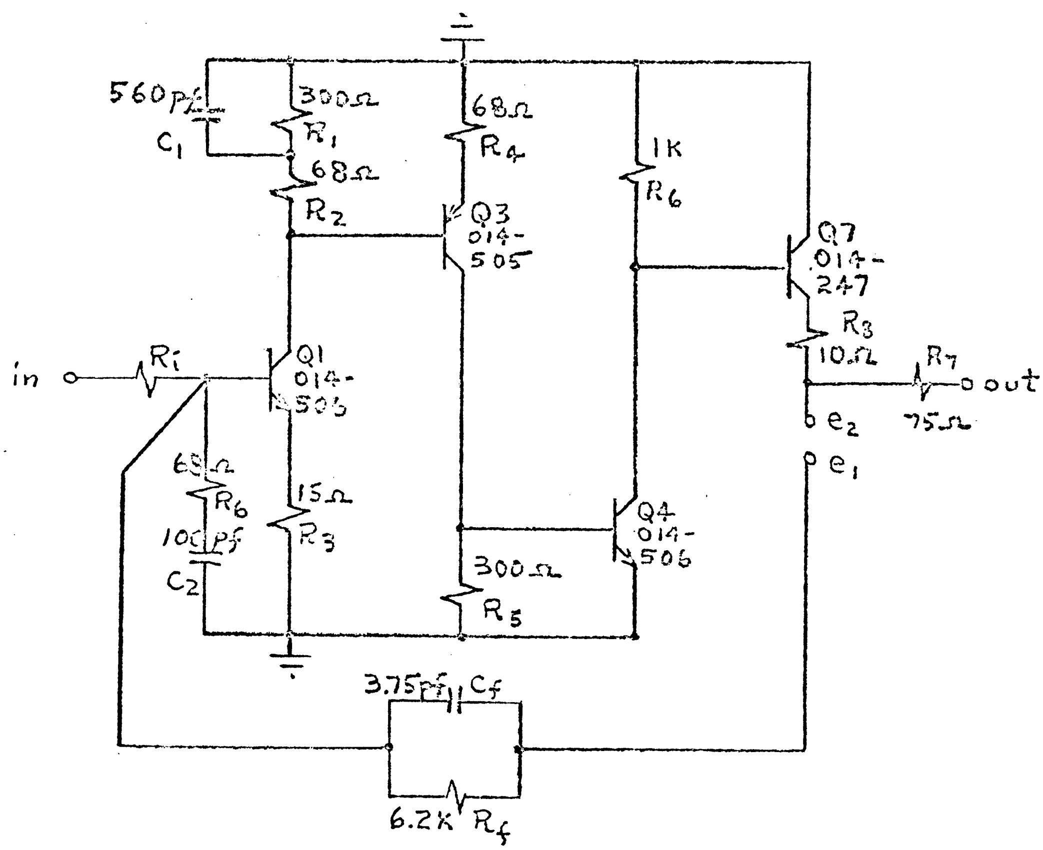


Figure 23. High Frequency Equivalent Circuit of Shunt Feedback Triple

In order to determine whether the amplifier is stable or not we refer to the basic feedback theory advanced by Bode.[5] We compute the feedback factor which as Bode states at the top of p. 33 is the "transmission around the complete loop" or in Figure 23, e_2/e_1 (the loop may be broken at any convenient point). The feedback factor is often misnamed "loop gain." Then by examining the phase curve where the feedback factor is unity, we may discover the margin of safety or phase margin.

Beginning at e_1, we may form an equivalent circuit for the feedback network and first stage Q1. (See Figure 24.) Assume that the input is open.

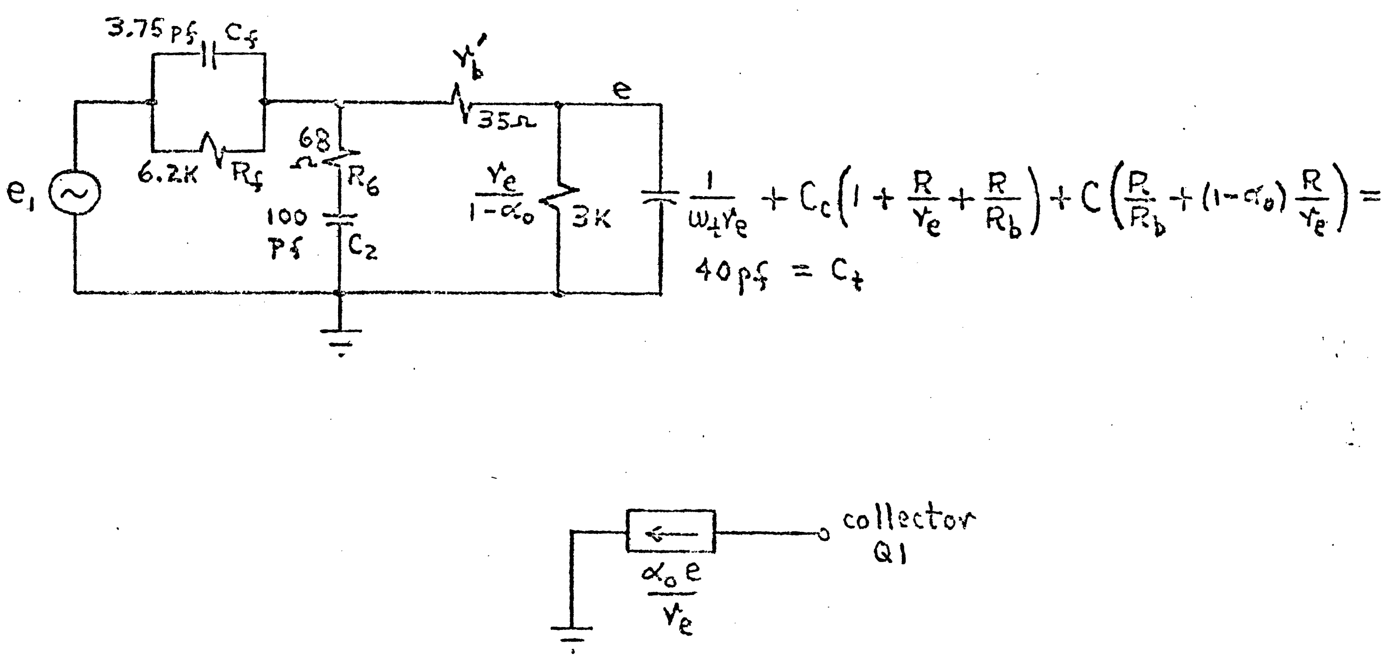


Figure 24. Equivalent Circuit for Feedback and First Stage

5. Network Analysis and Feedback Amplifier Design, by Hendrik W. Bode, D. Van Nostrand Co., Twelfth Printing, pp. 31-33.

The transistor model used in Figure 24 is the simplified first-pole equivalent for the transfer of a common emitter stage.[6] r_e includes both R_3 and the dynamic emitter resistance kT/qI_e which for Q1 will be about 15 ohms; r_e is therefore 30 ohms. Let us assume that $1-\alpha_o$ is 0.01 and that f_t is 350 mhz, and C_c is 4 pf. R is the collector load resistance, C is the collector load capacitance. At gain crossover (about 20 mhz), C_1 is virtually a short circuit such that R must be R_2 or 68 ohms. Furthermore, C cannot be much larger than $(R_5/R_4)\,C_{c3}$ or 15 pf. R_b, at gain crossover, is roughly r_b' plus R_6, since C_2 is getting small. Therefore

$$\frac{1}{\omega_t r_e} + C_c\left(1 + \frac{R}{R_e} + \frac{R}{R_b}\right) + C\left(\frac{R}{R_b} + (1-\alpha_o)\frac{R}{r_e}\right) =$$

$$15.2 \text{ pf} + 4 \text{ pf}\left(1 + \frac{68}{30} + \frac{68}{103}\right) + 15\left(\frac{68}{103} + (.01)\frac{68}{30}\right) =$$

$$15.2 \text{ pf} + 15.7 \text{ pf} + 9.1 \text{ pf} = 40.0 \text{ pf} \qquad \text{Eq. 57}$$

The major contributors are $\frac{1}{\omega_t r_e}$ and $C_c(1+\frac{R}{R_e})$ which justifies the inaccuracies in estimating C and R_b. $\frac{r_e}{1-\alpha_o}$ is 3.0 K.

The transfer of e_1 to e may be computed at mid-band frequencies (neglecting r_b') as

6. Class Notes from Wideband Transistor Amplifiers, Ampex, 1965, Ray Moore, p. 23.

$$\left.\frac{e}{e_1}\right|_{\text{Midband}} = \frac{\frac{r_e}{1-\alpha_o}}{\frac{r_e}{1-\alpha_o} + R_f} = \frac{3K}{3K + 6.2K} = 0.326 \qquad \text{Eq. 58}$$

The first pole may be calculated by assuming R_6, r_b' and C_f small (zero). The pole is then determined by the parallel combination of R_f and $\frac{r_e}{1-\alpha_o}$ and the sum of C_2 and the C_t from the transistor. Note that C_2 was chosen to be more than twice the capacitance of the transistor; C_2, therefore, predominates the gain above the first pole. This is in line with the general approach to amplifier design of making the significant high frequency (gain crossover) characteristics independent of transistor parameters, thereby reducing the probability that the amplifier will oscillate. This first pole is

$$\text{pole} = \frac{1}{2\pi \underbrace{\frac{r_e}{1-\alpha_o} \parallel R_f}_{2.02\text{ K}}} \quad \frac{1}{C_2 + C_t} = 560 \text{ khz} \qquad \text{Eq. 59}$$

There is a zero determined by R_f and C_f (again, both passive components).

$$\text{zero} = \frac{1}{2\pi R_f C_f} = 6.85 \text{ mhz} \qquad \text{Eq. 60}$$

A shunt feedback zero of this form is called a phantom zero.

Another, higher frequency zero occurs at a frequency determined primarily by R_6 and C_2. This zero will be approximately

$$\text{zero} = \frac{1}{2\pi R_6 C_2} = 23 \text{ mhz} \qquad \text{Eq. 61}$$

A more exact computation may be made by considering all the significant components at 23 mhz as shown in Figure 25.

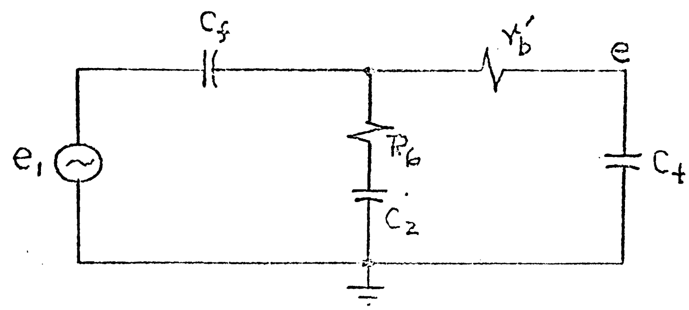


Figure 25. First Stage Equivalent at 23 mhz

An approximation helps. Since C_f is very small we may say that e_1 in series with C_f is a current source e_1pC_f. The resulting equivalent circuit is then

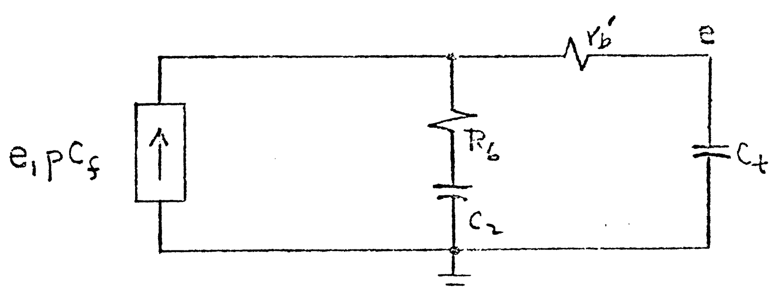


Figure 26. Modified First Stage Equivalent at 23 mhz

The transfer is given in Eq. 62.

$$\frac{e}{e_1} = \frac{C_f}{C_2 + C_t} \quad \frac{pR_6C_2 + 1}{p(r_b' + r_b)\,\dfrac{C_2C_t}{C_2C_t} + 1} \qquad \text{Eq. 62}$$

Eq. 61 is, therefore, quite accurate. But Eq. 62 also reveals an extra pole at

$$\text{pole} = \frac{1}{2\pi} \quad \frac{1}{r_b' + R_6} \quad \frac{C_2 + C_t}{C_2 C_t} = 54 \text{ mhz} \qquad \text{Eq. 63}$$

Therefore, in the region of gain crossover, the gain is determined by $\frac{C_f}{C_2 + C_t}$ as modified by a zero determined by R_6 and C_2 -- the response is predominantly determined by R_6, C_2 and C_f, all passive, predictable components. The pole in Eq. 63 strongly depends upon C_t but is high enough not to cause trouble.

It should further be noted that if the input resistor R_i (Figure 23) is larger than 500 ohms or so, the feedback factor in the region of gain crossover is virtually independent of the source. This is not true of series feedback and is a definite advantage of the shunt form.

The collector current of Q1, i_{c1}, is then from Eq. 58 and Figure 24

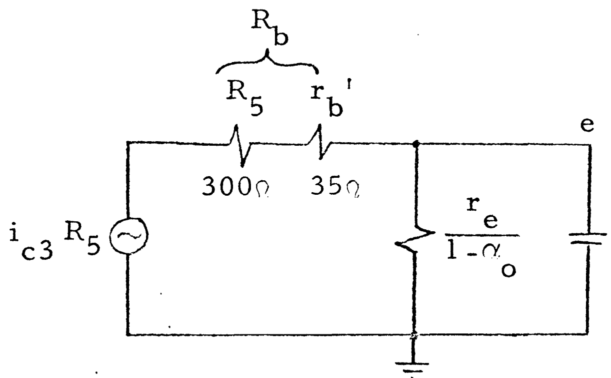
$$\frac{i_{c1}}{e_1} = \underbrace{\frac{\alpha_o}{r_e + (1-\alpha_o)R_f}}_{1/92\ \Omega} \quad \frac{(\text{zero Eq. 60})(\text{zero Eq. 61})}{(\text{pole Eq. 59})(\text{pole Eq. 63})} \qquad \text{Eq. 64}$$

At this point, it is convenient to move to the last stage and work back to the collector of Q1. To compute the load resistance and capacitance for Q4 (Figure 23) it is convenient to write the expression for the input impedance of an emitter follower (Q7)

$$Z_{in} = r_b' + \frac{r_e + R_L}{1-\alpha_o} \quad \frac{1}{1 + p\left(\frac{r_e + R_L}{1-\alpha_o}\right)\left(C_c + \frac{1}{\omega_\alpha(r_e + R_L)}\right)} \qquad \text{Eq. 65}$$

R_L, the emitter load resistance, will be at least 10Ω (R_8, Figure 22), plus 75Ω (R_7, Figure 22), plus the load of 75Ω which totals 160Ω. If r_e were 5Ω, $r_e + R_L$ would be 165Ω. Neglecting r_b' and assuming $1-\alpha_o$ equal to 0.01, the emitter follower then looks like a resistor 16.5K in parallel with a capacitor $C_c + \frac{1}{\omega_\alpha(r_e + R_L)}$ which, for ω_α equal to 300 mhz, and C_c equal to 7 pf, becomes about 10 pf.

Let us next compute the current to voltage transfer of Q4.[7] An equivalent circuit for Q4 may be drawn:



$$\frac{1}{\omega_t r_e} + C_c\left(1+\frac{R}{r_e}+\frac{R}{R_b}\right) + C\left[\frac{R}{R_b} + (1-\alpha_o)\frac{R}{r_e}\right] =$$

$$C_t = 11,060 \text{ pf}$$

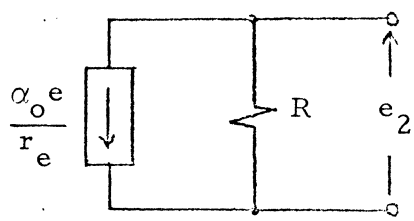


Figure 27. Equivalent Circuit of Q4.

A Thevenin equivalent of the collector current i_{c3} of Q3, taken as a current source, and R_5 yields the voltage source $i_{c3} R_5$ in series with R_5. R is R_6 in parallel with the 16.5K computed for the emitter follower Q7;

7. Ibid, p. 23.

R is 9.43 K. C is the 10 pf computed for emitter follower Q7. Since it is assumed that the emitter follower Q7 has a unity voltage gain, the output voltage in Figure 27 is e_2 from Figure 23. Assuming r_e is 5 ohms, $1-\alpha_o$ is 0.01, f_t is 350 mhz, C_c is 3 pf plus the external added 3 pf from Figure 22, R_b is 335 Ω, C_t for Q4 in Figure 27 becomes

$$C_t = \frac{1}{\omega_t r_e} + C_c\left(1 + \frac{R}{r_e} + \frac{R}{R_b}\right) + C\left(\frac{R}{R_b} + (1-\alpha_o)\frac{R}{r_e}\right)$$

$$= 91 \text{ pf} + 6 \text{ pf}\left(1 + \frac{9430}{5} + \frac{9430}{335}\right) + 10 \text{ pf}\left(\frac{9430}{335} + .01\,\frac{9430}{5}\right)$$

$$= 91 \text{ pf} + 6 \text{ pf}(1 + 1890 + 28) + 10 \text{ pf}(28 + 19)$$

$$= 91 \text{ pf} + 11,500 \text{ pf} + 470 \text{ pf} = 11,060 \text{ pf} \qquad \text{Eq. 66}$$

C_t is dominated by $C_c \times R/r_e$. Therefore, by putting a passive 3 pf capacitor in parallel with the C_c of the transistor, we are able to reduce the effect of variations in C_c from one transistor to another. Again, we are attempting to make the high frequency performance independent of transistor performance, without sacrificing too much bandwidth.

$\frac{r_e}{1-\alpha_o}$ is 500 ohms. There is a single pole at

$$\text{pole} = \frac{1}{2\pi}\;\frac{1}{\left(\frac{r_e}{1-\alpha_o} \;||\; R_b\right) C_t} = 71.6 \text{ khz} \qquad \text{Eq. 67}$$

The transfer from i_{c3} to e_2 is then, from Figure 27,

$$\frac{e_2}{i_{c3}} = \underbrace{\frac{\alpha_o RR_5}{r_e + (1-\alpha_o) R_b}}_{336\ K} \quad \frac{1}{(\text{pole Eq. } 67)} \qquad \text{Eq. 68}$$

The only remaining unknown factor is the transfer from the collector current of Q1, i_{c1}, to the collector current of Q3, i_{c3}. In order to compute the pole location for Q3, it is convenient to examine the input impedance[8] of the base of Q4, since it forms part of the load for Q3. The low frequency value is simply $r_b' + \frac{r_e}{1-\alpha_o}$ or about 535 ohms. There is a zero of input impedance at

$$\text{zero} = \frac{1}{2\pi} \; \frac{1}{R} \; \frac{1}{C + C_c} = 1.05 \text{ mhz} \qquad \text{Eq. 69}$$

There is a pole of input impedance at

$$\text{pole} = \frac{1}{2\pi} \; \frac{1-\alpha_o}{r_e} \; \frac{1}{\frac{1}{\omega_t r_e} + C_c(1+ \frac{R}{r_e}) + C(1-\alpha_o)\frac{R}{r_e}} = 27.3 \text{ khz} \qquad \text{Eq. 70}$$

Therefore, above 1 mhz, the input impedance becomes

$$\text{input impedance at base of Q4} = r_b' + \frac{r_e}{1-\alpha_o} \times \frac{27.3 \text{ khz}}{1.05 \text{ mhz}}$$

$$= 35\,\Omega + 500\,\Omega \times \frac{27.3}{1050} = 48\,\Omega \qquad \text{Eq. 71}$$

and it is <u>resistive</u>. Therefore, for Q3, the load resistance R is 48 Ω and C is zero above 1 mhz.

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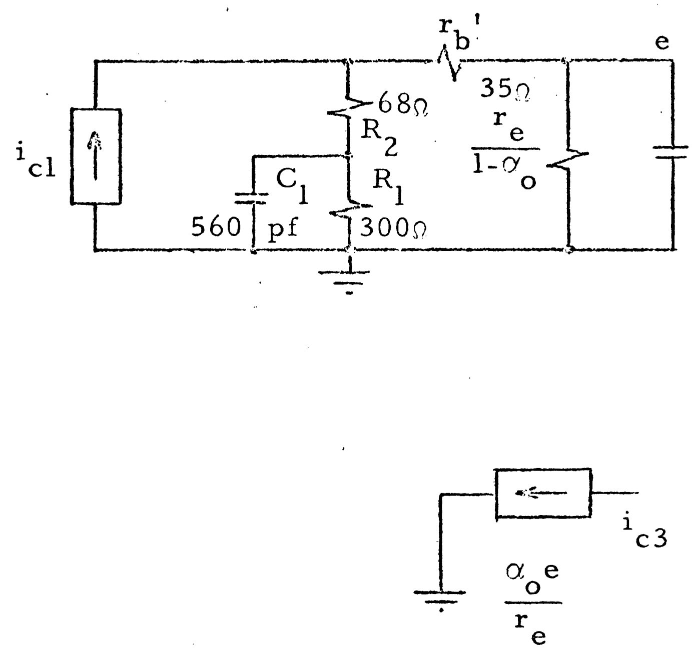
8. Ibid, p. 24.

We may now draw the equivalent circuit[9] of Q3 from the collector of Q1 to the collector of Q3 (Figure 28).



$$\frac{1}{\omega_t r_e} + C_c\left(1+\frac{R}{r_e}+\frac{R}{R_b}\right) + C\left[\frac{R}{R_b} + (1-\alpha_o)\frac{R}{r_e}\right] = C_t$$

Figure 28. Equivalent Circuit of Q3.

For Q3, $1-\alpha_o$ is 0.01, r_e is 15 ohms + R_4 or 83 ohms, f_t is 300 mhz, C_c is 3 pf, R is 48 ohms, R_b is $r_b{}' + R_2$ or 103 ohms.

$$C_t = 6.4 \text{ pf} + 3 \text{ pf}\left(1 + \frac{48}{83} + \frac{48}{103}\right) + 0 = 12.5 \text{ pf} \qquad \text{Eq. 72}$$

Since $\frac{r_e}{1-\alpha_o}$ is 8.3 K, the high frequency transistor pole is determined by $R_2 + r_b{}'$ and C_t

$$\text{pole} = \frac{1}{2\pi}\,\frac{1}{(R_2 + r_b{}')\,C_t} = 123 \text{ mhz} \qquad \text{Eq. 73}$$

9. Ibid, p. 23.

which is well out of the picture.

There is a pole-zero pair determined by R_1, C_1, and R_2.

$$\text{pole} = \frac{1}{2\pi} \frac{1}{R_1 C_1} = 0.946 \text{ mhz} \qquad \text{Eq. 74}$$

$$\text{zero} = \frac{1}{2\pi} \frac{R_1 + R_2}{R_1 R_2 C_1} = 5.14 \text{ mhz} \qquad \text{Eq. 75}$$

The transfer from i_{c1} to i_{c3} is then

$$\frac{i_{c3}}{i_{c1}} = \underbrace{\frac{\alpha_o}{r_e} (R_1 + R_2)}_{4.40} \frac{(\text{zero Eq. 75})}{(\text{pole Eq. 74})} \qquad \text{Eq. 76}$$

We may now combine Eq. 76 with Eqs. 68 and 64

$$\frac{e_2}{e_1} = \underbrace{\frac{i_{c1}}{e_1}}_{\text{Eq. 64}} \underbrace{\frac{i_{c3}}{i_{c1}}}_{\text{Eq. 76}} \underbrace{\frac{e_2}{i_{c3}}}_{\text{Eq. 68}} \qquad \text{Eq. 77}$$

Rewriting Eq. 77,

$$\frac{e_2}{e_1} = \frac{1}{92} \times 4.4 \times 336\text{ K} \; \frac{\overset{6.85 \text{ mhz}}{(\text{zero Eq. 60})}\overset{23 \text{ mhz}}{(\text{zero Eq. 61})}}{\underset{0.56 \text{ mhz}}{(\text{pole Eq. 59})}\underset{54 \text{ mhz}}{(\text{pole Eq. 63})}} \times$$

$$\frac{\overset{5.14 \text{ mhz}}{(\text{zero Eq. 75})}}{\underset{0.946 \text{ mhz}}{(\text{pole Eq. 74})}} \; \frac{1}{\underset{71.6 \text{ khz}}{(\text{pole Eq. 67})}} \; \frac{1}{\underset{123 \text{ mhz}}{(\text{pole Eq. 73})}} \qquad \text{Eq. 78}$$

The low frequency value of feedback factor is 16.1×10^3 or 84.1 db. The plot of Eq. 78 is given in Figure 29.

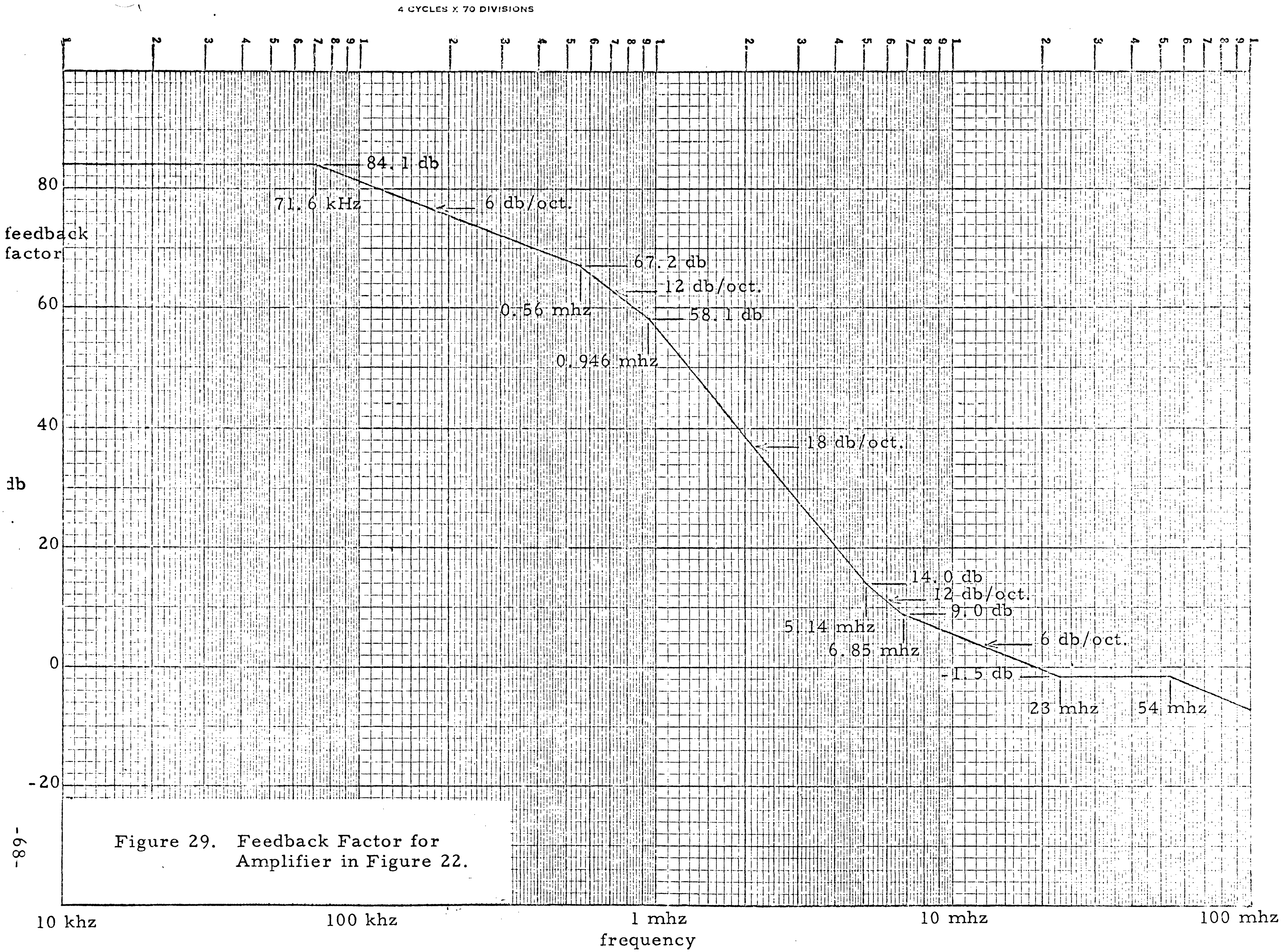


Figure 29. Feedback Factor for Amplifier in Figure 22.

As predicted, the gain crossover occurs at 20 mhz. The effect of the three zeros is to flatten the gain out above 5 mhz but before gain crossover is reached. There is another pole (Eq. 73) at 123 mhz and will be others at higher frequencies. Therefore, it is essential in this multi-time-constant amplifier to severely flatten the feedback factor curve in the region of gain crossover in order to reduce the phase lag to prevent oscillation. It is convenient to place the 23 mhz - 54 mhz shelf slightly below crossover to insure that the curve does not approach crossover at the higher frequency. Clearly, the feedback factor zero db crossover is at such a gentle slope that the phase cannot be much more than 90°. This fact is experimentally verified by transient response measurements: at low levels, there is never any overshoot or ringing. Only when the square wave is so large as to nearly cut off transistor Q4 does any overshoot begin to appear. But even then, there is only one overshoot with no subsequent ringing. This is positive experimental evidence that the phase margin is quite conservatively large.

The incredible aspect of the amplifier is that the feedback factor rises so steeply below 5 mhz. It is, in fact, conditionally stable with an 18 db/octave slope over a 5:1 range in frequency. The feedback factor at 2 mhz is <u>39 db</u> which would be unthinkable in a series feedback amplifier. A series feedback amplifier, with a 20 mhz crossover, cannot exceed a feedback factor of 20 db at 2 mhz, and with the transistor types used in this design, the series feedback would very likely be nearly oscillating with a crossover at 20 mhz. The differences become even greater as the frequency drops. At 1 mhz the feedback factor is 56 db.

It should be further emphasized that the important time constants are all controlled by passive components and are, to a large extent, independent of transistor parameters. Furthermore, the emitter of the final gain stage Q4 is grounded, so that the collector may swing the full supply voltage. This is not possible with any normal series feedback configuration. Also, the only tantalum capacitors are the input and output coupling capacitors. There are no bypass capacitors. The ac feedback and dc feedback are the same, so that the 84 db feedback factor goes all the way to dc. Thus, there cannot be any low frequency oscillation problems because there is no low frequency gain crossover. This is very good design practice in any amplifier.

The specifications which may be required of the amplifier in Figure 22 are given in Table 4. (The amplifier will be flat within $\pm .5$ db from 50 hz to 2.4 mhz.)

Transient Response

At 4 vptp into a 75 ohm load:

- 5% maximum overshoot
- No ringing
- Rise time 40 nsec maximum, 10 to 90%
- Fall time 50 nsec maximum, 10 to 90%

Distortion

At 2.5 vrms into 75 ohms:	V_{cc} 20 v	V_{cc} 25 v
2nd harm. of 1 mhz	-60 db	-65 db
3rd harm. of 0.667 mhz	-60 db	-65 db

Power Supply Current Drain +20%, -10%	V_{cc} 20 v	V_{cc} 25 v
No load	22.5 ma	27.5 ma
At 2.5 vrms into 75 ohms	34.5 ma	39.5 ma

Table 4. Specifications for Amplifier in Figure 22.

8. THE FM MODULATOR

Excellent block diagrams for the modulator along with explanations of its functioning were prepared for the manual by Loren Clark of Training. The ES-200 manual should, therefore, be a good source for this type of information. For the most general and complete presentation of the combinations and permutations of jumpers, I recommend SP1801609. The most useful thing I can do here, is to first give my version of how the basic modulator works, and second give a section of special design aspects.

8.1 The Functioning of the Basic Modulator (See Schematic #1245993)

The basic oscillator (modulator) consists of Q11, 12, 13, 14, 15, 16, 17, 18, 19 and 20. When reduced to essentials, the modulator is rather simple. Consider Figure 30.

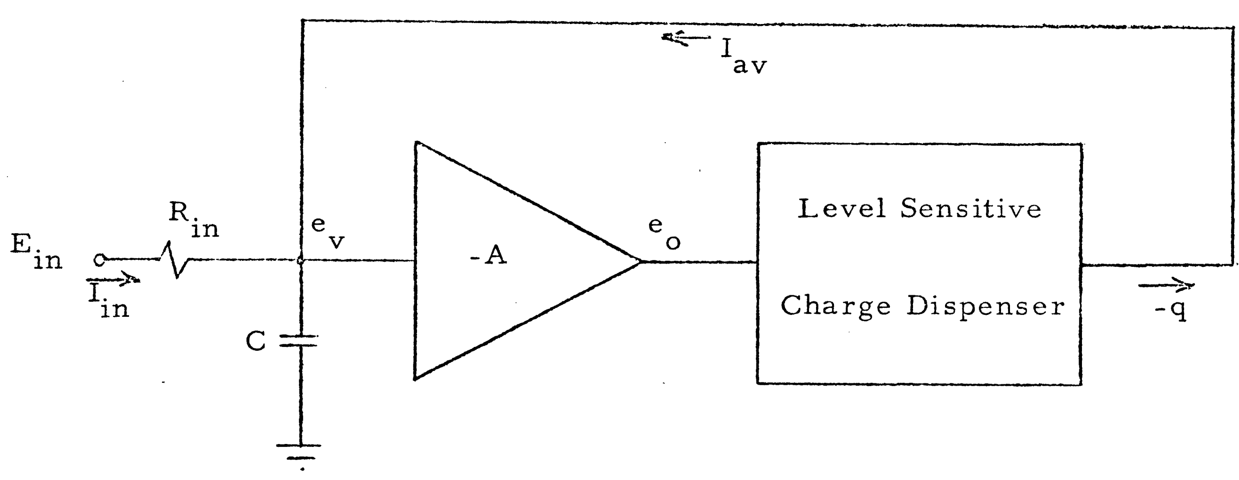


Figure 30a. Simplified Block Diagram of Modulator

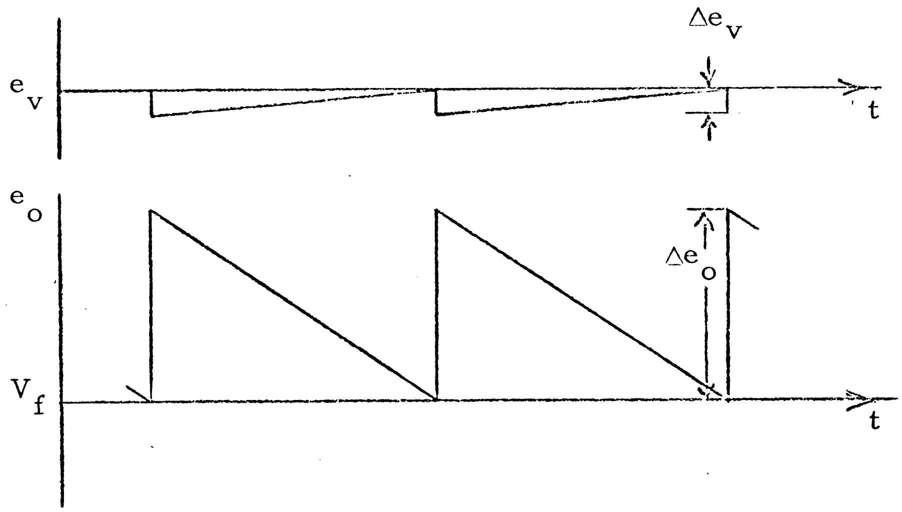


Figure 30b. Waveforms in Block Diagram of Modulator

Assuming a positive e_{in}, current flows through R_i into capacitor C causing the voltage to slowly rise. This voltage waveform is amplified and inverted by the amplifier with gain -A. Thus at the output of the amplifier, the voltage is slowly falling as C charges. Finally, this voltage reaches a critical firing level (as with a Schmitt trigger circuit) which actuates the "level sensitive charge dispenser." At this time, the charge dispenser emits a constant lump of negative charge -q back into the capacitor C, causing the amplifier input voltage to drop an amount Δe_v which is given by

$$\Delta e_v = \frac{-q}{C} \qquad \text{Eq. 79}$$

Thus at the amplifier's output the voltage suddenly moves away from the firing voltage by an amount Δe_o which is given by

$$\Delta e_o = A\frac{q}{C} \qquad \text{Eq. 80}$$

The positive input begins to slowly charge the capacitor again, resulting in a continuing cyclic functioning.

If we assume that capacitor C is large, and gain -A also appropriately large, both e_v and Δe_o become quite small -- virtually zero. Thus we have a virtual ground at the amplifier's input. The input current is, therefore, given by

$$I_{in} = \frac{E_{in}}{R_{in}} \qquad \text{Eq. 81}$$

The average current fed back by the charge dispenser is

$$I_{av} = fq \qquad \text{Eq. 82}$$

where f is the frequency of charge dispensing. If the amplifier takes no current, Eq. 81 must equal Eq. 82 so that

$$f = \frac{E_{in}}{R_{in}q} \qquad \text{Eq. 83}$$

A constant current I_{off} fed into the summing node e_v will cause a constant frequency to result. Then E_{in} may be positive or negative and will cause the modulator to deviate about the center frequency f_{cc}.

$$f_{cc} + f = \frac{I_{off}}{q} + \frac{E_{in}}{R_{in}q} \qquad \text{Eq. 84}$$

Eq. 84 is, therefore, the basic law of the modulator.

But there is a starting problem. If the amplifier is in limiting, there may be no charges fed back to the summing node. Therefore, rather than using a Schmitt trigger to sense the amplifier output, a gated astable multivibrator is used, that is, rather than firing a Schmitt, the amplifier inhibits or gates a multivibrator. Then if the amplifier is in limiting in the direction which indicates the need for charge, the multivibrator will free run until the loop is balanced and the amplifier takes over. A new block diagram may be drawn (Figure 31) including reference designations to Schematic #1245993.

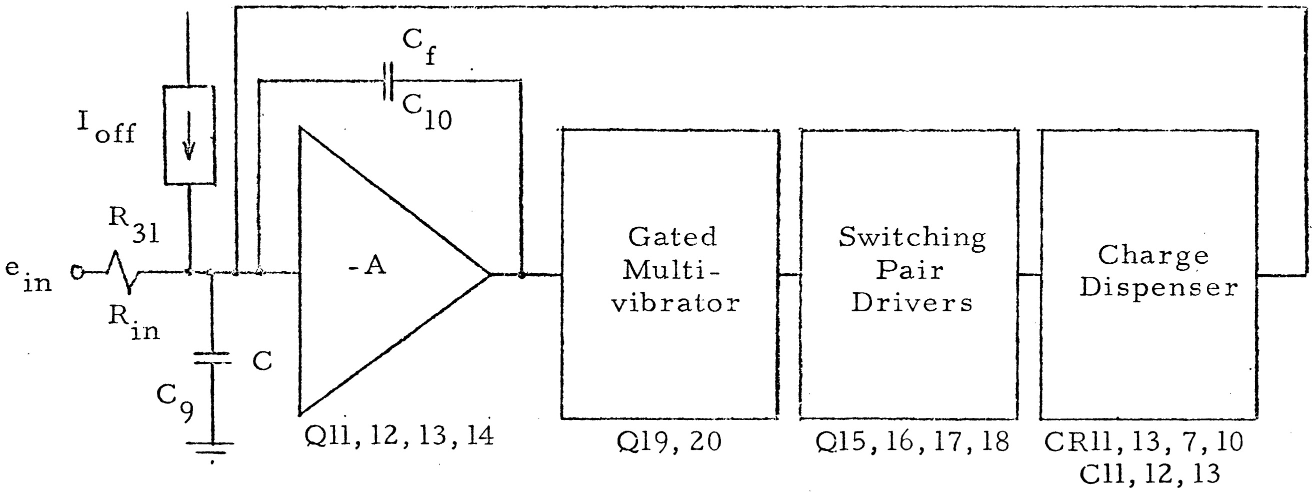


Figure 31. Block Diagram of Modulator

In order that $C(C_9)$ not be so large, the amplifier is used as a Miller amplifier. The actual C is then

$$C = C_9 + (1 + A)\, C_f$$

$$= 680 \text{ pf} + (501)\ 130 \text{ pf}$$

$$= 680 \text{ pf} + 65,000 \text{ pf} \qquad \text{Eq. 85}$$

From Eq. 85 we see that, since the predominant capacitor is $(1 + A)\, C_f$, all but about 1% of the charge q goes into C_f. Therefore, we must recompute the value of Δe_o in Figure 30b.

$$\Delta e_o = \frac{q}{C_f}$$ Eq. 86

Since q is about 315 pc and C_f is 130 pf, Δe_o is about 2.4 v.

The charge dispenser is shown simplified in Figure 32. The left side of a precision capacitor C_{prec} is clamped between ground and a precision

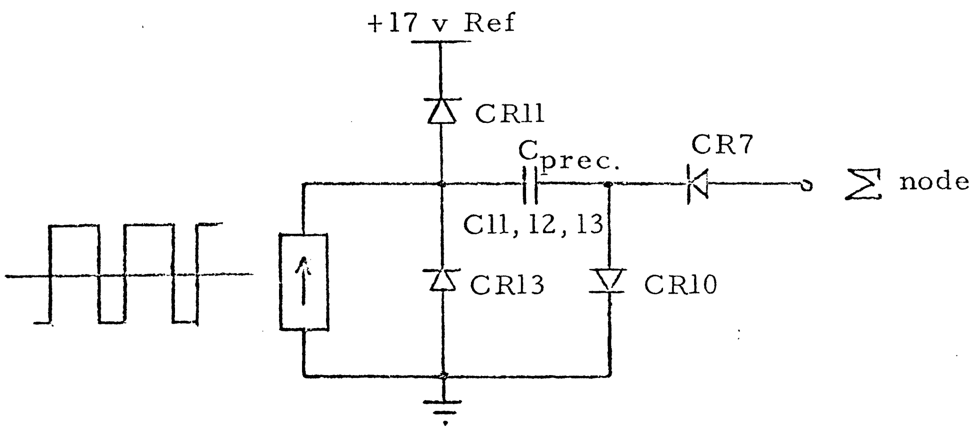


Figure 32. Charge Dispenser Simplified

17 v reference voltage using diodes CR13 and CR11. The charge is rectified by diodes CR10 and CR7 so that only the negative charge passes to the summing node. The cycle (determined by the astable) begins by driving the left side of the capacitor from + 17 v to ground; the charge

$$q = E_{ref}\ C_{prec}$$ Eq. 87

is dispensed to the summing node. Then, the left side of the capacitor is taken positive again in preparation for the next cycle. The secret of the charge dispenser is that q is determined only by E_{ref} and C_{prec} and the match of the four cancelling diode drops. q is therefore independent of the precision of the driving current or the preciseness of timing in the astable; it may be held extremely constant.

The drive for the charge dispenser is accomplished by current switching pairs Q15, 16 and Q17, 18 which are switched by the astable.

The astable provides the timing for the dispensing cycle and recharge cycle. It is inhibited or gated from the amplifier via diodes CR14 and CR16. CR15 inhibits the amplifier during the dispensing cycle so that the amplifier won't gate the astable off before the dispensing cycle is complete. It should be noted and emphasized that the frequency of the modulator has no relationship to the free-running frequency of the astable.

8.2 Special Design Aspects

The preamp, Q1, 2, 3, 4, 5, 7, 8, 9, 10 is a low-drift, series feedback, unity gain, dc amplifier. The only innovative aspect of the amplifier is the use of the base current of Q1 to cancel the base current of Q4A and the use of the base current of Q7 to cancel the base current of Q7. Since base currents typically vary 1%/$^{\circ}$C., once these cancelling currents are made equal to the currents being cancelled, they will track with temperature. Thus, without chopping, we achieve not only a low offset voltage drift but a nearly zero offset current. Since the V_{EB} drops for transistors track better with temperature if they are initially set up equal than if emitter currents are initially set up equal, the offset voltage adjustment is in the collectors of Q4A and B rather than the emitters. The 2K resistor R9 and capacitor C1 help isolate the feedback loop from the effects of differing source impedances.

R57 which is in series with CR13 is one of those unexplainable things. The modulator performs better in the oven with it than without it.

The second harmonic adjustments R71, R75, and R78 simply slide the slicing level of switching pair Q28, 29 up and down on the non-zero rises and falls of the waveforms coming out of the flip-flops. By this means, perfect symmetry in the output is accomplished, and therefore, a null of even harmonics.

A dc amplifier was used to output the carrier because tantalum is more expensive than transistors and sag is to be kept to a minimum. The carrier is eventually ac coupled to the head by the head driver but its cutoff is down around 10 hz, so its sag is not offensive.

By the use of many jumpers, the modulator has been made all-things-to-all-customers. It will accommodate any of the standard IRIG-106-66 bandwidths with a great variety of input levels and impedances.