national accelerator laboratory

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QUADRUPOLE POWER SUPPLY ACTIVE RIPPLE FILTERS

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ABSTRACT

A 40-kW active filter is used with each of the main accelerator's 12-phase, quadrupole power supplies to reduce the power supply's voltage ripple. The active filter operates on a voltage-bucking principle with the filter output transformer-coupled into the quadrupole bus. Filtering is accomplished by sensing the combined power supply-active filter output and acting upon that signal with the filter to cancel the ripple. The sensed signal is amplified and used to drive a transistor bank which drives the primary of the output transformer in a Class A mode. Under normal conditions, the quadrupole power supplies are operated in a ramped mode with slopes during acceleration of 10^4 V/sec. The electronics of the active filter is designed to respond only to the first three harmonic components of ripple in the power supply output. Attenuation of the voltage ripple during injection, acceleration, and extraction is approximately 30dB. Quadrupole bus current ripple is correspondingly reduced.

GENERAL OPERATION

The purpose of the NAL quadrupole active ripple filters is to reduce the voltage ripple produced by the pulsed quadrupole power supplies. Reducing the voltage ripple in turn reduces the current and magnetic field ripple that affects the proton beam. Since the SCR phase-controlled power supplies are twelve-phase supplies, the outputs contain 720 Hz and higher harmonic components of ripple. The filters are designed to remove the 720, 1440, and 2160 Hz components.

Figures 1 and 2 show simplified block diagrams of the active filter and how it is connected to the quad power supplies. While the power supply output is ramped from zero to 1000 V and back to zero, a ripple voltage of varying amplitude and harmonic content is present in the supply output. The active filter operates to reduce the power supply ripple by driving the reactor-transformer in such a way as to "buck-out" or cancel the ripple and thus present a "filtered" output to the quadrupole magnet load. Ripple is reduced during the entire quad power supply pulse cycle. However, the reduction in ripple is most important during extraction (time when power supplies are in flat top) so as to remove modulation effect in the extracted beam.

Power for the active filter is derived from the quad power supply's 350 V ac power transformer. The voltage is stepped down, rectified, and capacitor filtered to provide a 40 V dc source. A transistor bank with 16 current amplifier modules and 20 transistors/module operating in parallel, drive the reactor-transformer using the 40 V dc source for power. Base drive for the transistor bank is derived from a power amplifier which is driven by low level frequency selective amplifiers. The low level input signal to the amplifiers comes from a high impedence voltage divider connected as shown in Figure 1. The transistor bank, connected as an emitter-follower to the power amplifier, operates in a Class A mode with a total dc bias current of 500 to 800 A dc driving the reactor-transformer primary. Figure 3 presents pertinent active filter waveforms which illustrate how the quadrupole power supply is filtered. If the filter was not present in the circuit, the quad power supply voltage ripple would appear across the quad magnet load. With the filter connected as shown in Figure 1, the ripple is attenuated as shown in Figure 3e. The amount of attenuation is essentially equal to the open-loop gain of the filter at the various harmonic frequencies. Open-loop gain is set by the voltage divider, input amplifier, frequency shaping amplifiers, power amplifier, transistor bank, and turns ratio of the reactor-transformer. Nominally, the open loop gain is about 30, resulting in about a 30 db reduction in power supply ripple. The filter electronics is designed to provide stable operation under all normal conditions, and for ease in adjustment. A plot of the filter closed-loop gain versus frequency is shown in Figure 13.

FILTER FREQUENCY AND TIME RESPONSE

Figure 4 shows the components which affect the filter's frequency and time response. Assume that the attenuator, input amplifier summing amplifier and driver amplifier have a frequency response which is flat in the area of interest. Then the gains of these sections may be lumped together as B(s) with a constant value M. In the filter, the summing amplifier and driver amplifier are inverting amplifiers.

$$B(s) = M = \text{constant} \qquad (1)$$

Typically

$$M = \frac{1}{150} (+1.5)(-1.0)(-10) = +0.1 \qquad (2)$$

In Figure 5, the block diagram is simplified and drawn as a controls feedback system. (Note, the feedback loop has a net signal inversion and therefore the filter is actually a negative feedback system.)

The response of the frequency shaping amplifier A3 is designated as C(s), the compensating circuit A4 as F(s) and the emitter resistor-reactor combination as D(s). A further simplification of the feedback system is shown in Figure 6, where

$$H(s) = MC(s)F(s)D(s) \qquad (3)$$

H(s) is the open loop gain of the system.

To provide system stability and good performance over a variety of operating conditions, the frequency shaping amplifier has been chosen to be a set of three parallel bandpass filters tuned to the first three harmonics of the power supply ripple. Fast, high, and low frequency gain roll-off is provided by the bandpass circuits with a minimum of phase shift to the system. For ease of adjustment, a negative immittance converter realization of the bandpass circuit was used. A generalized NIC bandpass circuit shown in Figure 7 has the following transfer characteristic:

$$\frac{E_2(s)}{E_1(s)} = \frac{-KRCs}{(sRC)^2 + sRS(2-K) + 1} \qquad (4)$$

$$\frac{E_2(s)}{E_1(s)} = \frac{-Ks}{\omega_0}\left(\frac{1}{\left(\frac{s}{\omega_0}\right)^2 + 2\zeta\frac{s}{\omega_0} + 1}\right) \tag{5}$$

where $R_1 = R_2 = R$ (6)

$C_1 = C_2 = C$ (7)

$K < 2$ (8)

$\omega_0 = \frac{1}{RC}$ (9)

$\zeta = 1 - \frac{K}{2}$ (10)

and $Q = \frac{1}{2-K}$ (11)

Typical gain and phase response for an NIC bandpass amplifier is shown in Figures 8A and 8B.

The response for the three parallel bandpass circuits representing the frequency shaping amplifier A3 is given by:

$$C(s) = \frac{-K(RC)_1 s}{s^2(RC)_1{}^2 + s(RC)_1(2-K) + 1} \quad \frac{-K(RC)_2 s}{s^2(RC)_2{}^2 + s(RC)_2(2-K) + 1}$$

$$\frac{-K(RC)_3 s}{s^2(RC)_3{}^2 + s(RC)_3(2-K) + 1} \tag{12}$$

where $(RC)_1$, $(RC)_2$, and $(RC)_3$ correspond to the first three harmonics in the power supply output as given by Eq. (9) and where K is choosen to be the same for each of the circuits.

A lead compensating network is used to improve the filter's response to the voltage ramp generated by the quadrupole power supply. The transfer ratio for the generalized circuit shown in Figure 9 is:

$$F(s) = \frac{E_{out}}{E_{in}} = \frac{1}{a}\frac{(1+aTs)}{(1+Ts)} \tag{13}$$

$$a = \frac{R_1 + R_2}{R_2} > 1 \qquad (14)$$

$$T = \frac{R_1 R_2 C}{R_1 + R_2} \qquad (15)$$

Sketches of the lead network gain and phase are shown in Figure 10.

$$\sin \phi_m = \frac{a - 1}{a + 1} \qquad (16)$$

$$\omega_m = \frac{1}{\sqrt{a}\ T} \qquad (17)$$

The effect of the phase lead network upon the filter performance is discussed later in this section.

Amplifier A6 represents the gain-phase characteristic generated by the power transistor modules and reactor-transformer combination. The equivalent transistor emitter resistance and transformer primary resistance (the sum is called R_E) in series with the transformer primary inductance causes a low frequency gain roll-off. Since the equivalent emitter resistance changes with the number of modules which are on, the low frequency break point and associated phase shift also change. At high frequencies, the ac voltage gain caused by the transformer turns ratio is about 5.25. Amplifier A6 is represented by the transfer function D(s) given below

$$D(s) = \frac{E_{out}}{E_{in}} = N \frac{L}{R_E} \left(\frac{s}{1 + s \frac{L}{R_E}} \right) \qquad (18)$$

where
- N = Transformer voltage turns ratio
- L = Transformer primary inductance
- R_E = Equivalent transistor emitter resistance and transformer primary resistance.

In the active filter N = 5.25, L = 8 μH, and R_p = .33 mΩ. For 16 modules there are 320 parallel 1Ω emitter resistors giving an equivalent emitter resistance of 3.1 mΩ. With 16 modules, the resulting low frequency breakpoint is:

$$f_c = \frac{R_E}{2\pi L} = \frac{(3.1 + .33) \times 10^{-3}}{2\pi \times 8 \times 10^{-6}} = 68 \text{ Hz} \tag{19}$$

With one module, the low frequency breakpoint is 995 Hz. Figure 11 shows the gain and frequency characteristic of D(s) for twelve modules.

Some information regarding the stability of the active filter can be obtained from Bode plots of open loop gain and phase. These plots can be obtained from the individual plots of B(s), C(s), D(s), and F(s). Figures 12a and 12b show typical Bode plots for the system. From Figure 12b it is seen that the phase margin at unity gain with 12 modules is 72°. With fewer modules on, the phase margin is less.

The closed-loop response for the system shown in Figure 6 is given by Eq. (20). By placing

$$\frac{E_0(s)}{E_1(s)} = \frac{1}{1 - H(s)} \tag{20}$$

the individual expressions for B(s), D(s), and F(s) in Eq. (12) and substituting that expression for H(s) in Eq. (20), the closed loop response, given by Eq. (21), is obtained:

$$\frac{E_0(s)}{E_1(s)} = \frac{1}{1 - M\, C(s) \frac{1}{a} \left[\frac{1 + aTs}{1 + Ts}\right] NL \left[\frac{s}{R_E + sL}\right]} \tag{21}$$

Two factors are important in the closed loop response: (1) attentuation of ripple and (2) the filter time response to the power supply voltage ramp. Consider the ripple attenuation first. The frequencies at which Eq. (21) is of most interest are 720, 1440, and 2160 Hz. Equation (21) may be evaluated more easily at these frequencies by making several simplyfing approximations which are generally valid. As an example, consider 720 Hz. The compensating amplifier response given by F(s) is chosen in such a way that near 720 Hz and high frequencies

$$(1+aTs) \cong aTs \tag{22}$$

$$(1+Ts) \cong Ts \tag{23}$$

and therefore near 720 Hz

$$\frac{1}{a}\,\frac{(1+aTs)}{(1+Ts)} \cong 1 \tag{24}$$

Also, with a number of modules on and near or above 720 Hz, the expression for D(s) reduces to the following:

$$\frac{NLs}{R_E + sL} \cong \frac{NLs}{sL} = N \tag{25}$$

Finally, at 720 Hz the last two terms in Eq. (12) are small compared to the first term and the expression for C(s) reduces to:

$$C(s) \cong \frac{-K(RC)_1 s}{s^2 (RC)_1^{\,2} + s(RC)_1 (2-K) + 1} \tag{26}$$

The transfer function near 720 Hz therefore reduces to Eq. (27)

$$\frac{E_0(s)}{E_1(s)} \cong \frac{1}{1 + \dfrac{MNK(RC)_1 s}{s^2 (RC)_1^{\,2} + s(RC)_1 (2-K) + 1}} \tag{27}$$

In the bandpass circuit, $(RC)_1$ is chosen to give a resonance at 720 Hz (f_1 = 720 Hz) and K is choosen to be about 1.96. Keeping the same values for M and N as used previously:

$$\left.\frac{E_0(j\omega)}{E_1(j\omega)}\right|_{\omega=\omega_1} \cong \frac{1}{1 + \dfrac{(0,1)(5.25)(1.96)j\omega}{\omega_1\left[\left(\dfrac{j\omega}{\omega_1}\right)^2 + j\dfrac{\omega}{\omega_1}(2-1.96) + 1\right]}} = .037 \tag{28}$$

Attenuation for the above example corresponds to -28.6 db. Equation (29) is a simplified expression for system attenuation at 720 Hz, it is also valid at 1440 and 2160 Hz.

$$\left.\frac{E_0}{E_1}\right|_{f = f_1,\ f_2,\ \text{or}\ f_3} \cong \frac{2-K}{MNK} \tag{29}$$

The overall closed-loop response of the filter is shown in Figure 13.

The other closed loop factor, the filter time response, is now considered. Normal operating conditions in the quadrupole power supply cause the output voltage to ramp from 0 to 1000 V in 0.1 sec and from 1000 V to 0 in 8 msec. The positive ramp tends to cause a shift in the bias current into the reactor-transformer. Examine the input voltage, $b'(s)$, to A_6 during the + ramp <u>without</u> the compensating amplifier A_4 present.

$$b'(s) = B(s)\, C(s)\, E_0(s) \qquad (30)$$

$$b'(s) = \frac{B(s)\, C(s)\, E_1(s)}{1 - H(s)} \qquad (31)$$

For the + ramp

$$E_1(s) = \frac{A}{s^2} \qquad (32)$$

where typically $A = 10^4$ volts/ second.

Substituting the expressions for $B(s)$, $E_1(s)$ and $H(s)$ into Eq. (31) results in Eq. (33)

$$b'(s) = \frac{MAC(s)}{s^2 \left[1 - C(s)\, MNL \left(\frac{s}{R_E + sL}\right)\right]} \qquad (33)$$

To find the steady-state solution, apply the final value theorem.

$$b'_{ss} = \lim_{s \to 0} sb'(s) \qquad (34)$$

$$b'_{ss} = -MKA \left[(RC)_1 + (RC)_2 + (RC)_3\right] \qquad (35)$$

For the values $M = 0.1$, $K = 1.96$, $A = 10^4$, $(RC)_1 = .22 \times 10^{-3}$, $(RC)_2 = .11 \times 10^{-3}$, and $(RC)_3 = .0733 \times 10^{-3}$, b'_{ss} has a value of -.784 V which means that the bias voltage applied to the reactor-transformer and emitter-resistor combination is decreased by -.784 V during the ramp. Since the equivalent series resistance in A_6 is 3.4 milliohms (with 16 modules), the bias current through the primary decreases by the following amount during the ramp:

$$\Delta Ip = \frac{-.784}{.0034} = -230 \text{ amps} \qquad (36)$$

As an example, if the bias current was set at 500 A dc, it would drop to 270 A during the ramp, limiting the filters ability to filter when power supply ripple is at its worst and possibly shuting down the filter with an undercurrent trip. The problem is reduced by using the lead network shown in Figure 9.

Examine the voltage b(s) during the ramp with the compensating amplifier A_4 present.

$$b(s) = \frac{B(s)\ C(s)\ F(s)\ E_1(s)}{1 - H(s)} \tag{37}$$

Applying the final value theorem in Eq. (37), the steady state value of b(s) during the ramp is found to be

$$b_{ss} = \frac{-MKA}{a} \left[(RC)_1 + (RC)_2 + (RC)_3\right] \tag{38}$$

where "a" is the ratio of the break frequencies shown in Figure 10. If a = 4 is chosen, the resulting decrease in bias current during the ramp would be only 57.5 A which would be probably acceptable. On the other hand, no stability problems arise if $a = \infty$. Therefore, choose $a = \infty$ and b_{ss} during the ramp as given by Eq. (38) is zero. From Eq. (14) it is seen that "a" may be set equal to ∞ by simply choosing $R_1 = \infty$, i.e., removing R_1 from the circuit.

When $R_1 = \infty$, Eq. (13) reduces to

$$F(s) = \frac{T' s}{1 + T' s} \tag{39}$$

where

$$T' = R_2 C \tag{40}$$

All of the assumptions and closed loop conclusions found in this section remain unchanged when $R_1 = \infty$.

The quad power supply output voltage fall rate is considerably faster than the rise time. An analysis similar to the ramp analysis can be made for the fall time showing the resultant increase in bias current. However, it is not necessary, since the fall time is very short (~ 8 milliseconds), the increased bias current is limited by the filters' current limit circuit to a value acceptable to the transistor module circuit breakers, and operation of the filter is not important during the fall time.

EXPERIMENTAL RESULTS

The purpose of the active filter is to reduce the voltage ripple in the quadrupole power supply output and thus reduce the corresponding current ripple in the quad bus. This section examines the effectiveness of the filters in reducing ripple.

Figure 1 shows where the waveforms presented in this section were observed. Although various power supply-filter pairs have somewhat different waveshapes, one non-regulating power supply filter combination is examined closely since it is representative of most other units. Figure 14 shows the quad power supply and active filter waveforms for a full main ring cycle with the filter off and then on. The improvement in ripple output is clearly evident from Figures 14a and 14b. Figures 14 c and 14 d show where the filter is generating the most voltage. Figure 15 shows the effect of the filter during the ramp. The most prominent frequency component seen in Figure 15a is the fundamental at 720 Hz. Effectiveness of the filter is easily seen in Figure 15b. The operation of the filter during extraction is shown in Figure 16. Strong harmonics components at the first and third harmonics (720 Hz and 2160 Hz) are seen across the power supply (magnet load voltage) when the filter is off. The effect of the filter during extraction is seen in Figure 16b. For all the above photos, attenuation of the 1st, 2nd and 3rd harmonics was set at 25.

As can be seen, the harmonic content of the power supply output changes during the main ring cycle. Also, the harmonic content changes with the current in the quadrupole bus (energy level) due to phase overlap. In all cases which have been observed, however, the predominant components are the 1st, 2nd and 3rd harmonics.

Due to the programmed voltage fluctuations in a regulating quad power supply, the waveforms shown in Figures 14 though 16 are somewhat different for the regulating supply. However, the improvement in ripple shown in these photos also occurs with the regulating supply.

Finally, Figure 17 shows the reduction of current ripple in the quadrupole horizontal focusing bus at 3325 A (300 GeV). When all of the quad power supplies in the bus have an active filter, the 1st, 2nd, and 3rd harmonics are practically eliminated in the current waveform. If any of the active filters are off, significant amounts of higher frequency ripple may be present in the quadrupole current. With the filters off, the peak to peak higher frequency current ripple is about 1/6600. When all the filters are on, the observed ripple is on the order of 1/75000 or less. From the above results, it is seen that the active filters are effective in reducing the 720, 1440, and 2160 Hz ripple components in the power supply output.

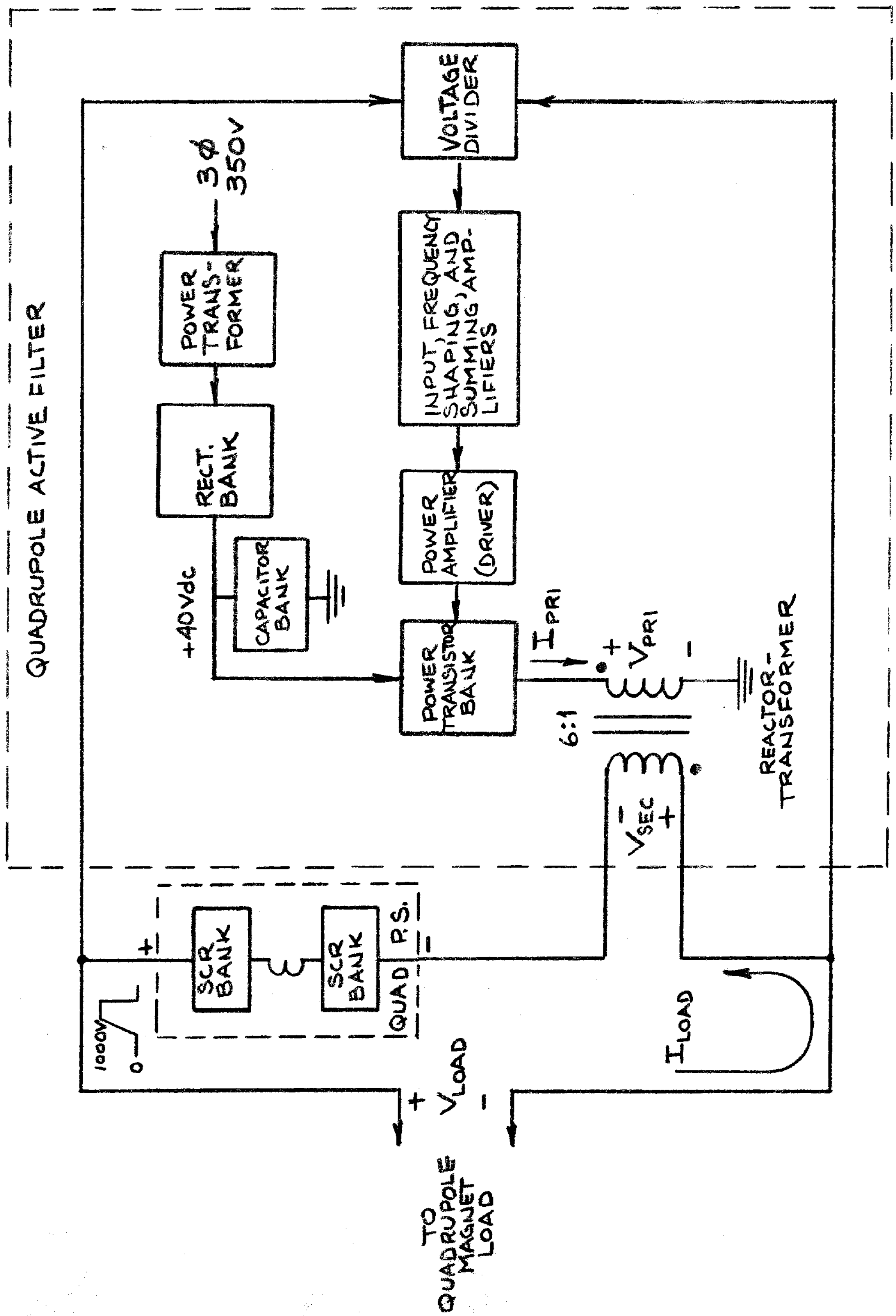


FIGURE 1 - ACTIVE FILTER SIMPLIFIED BLOCK DIAGRAM

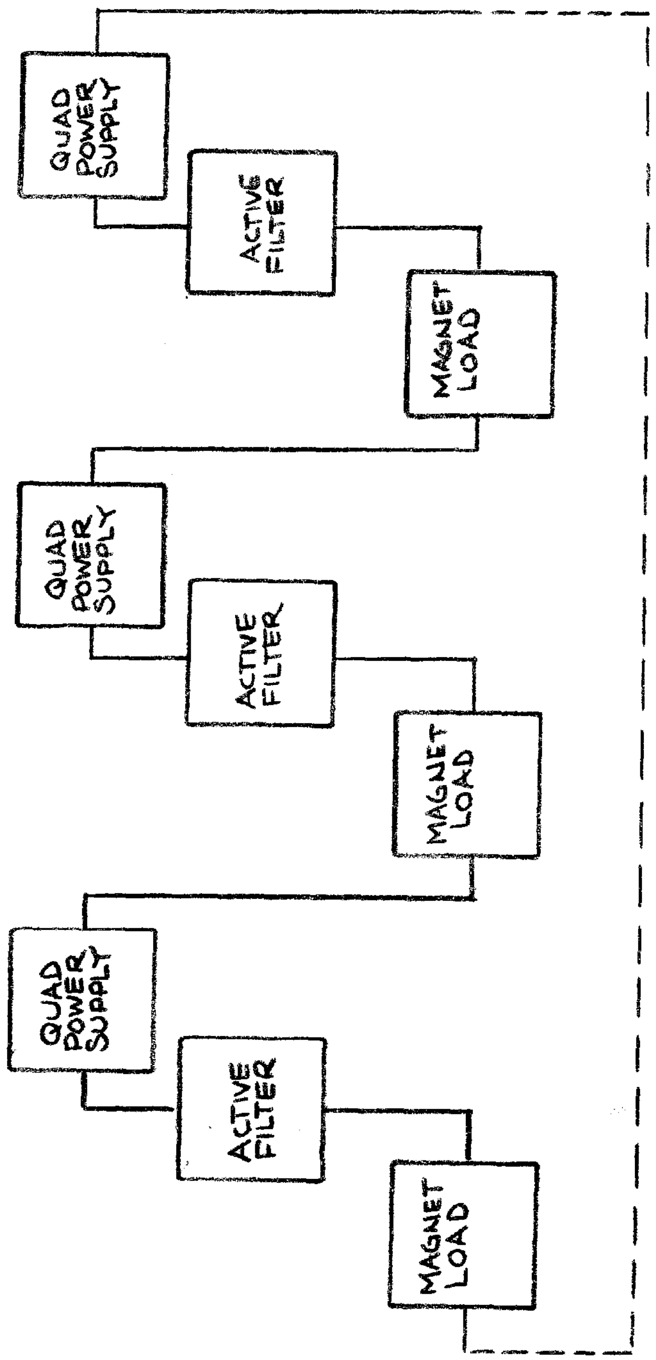


FIGURE 2 - QUADRUPOLE POWER SUPPLY, ACTIVE FILTER, LOAD CONNECTIONS

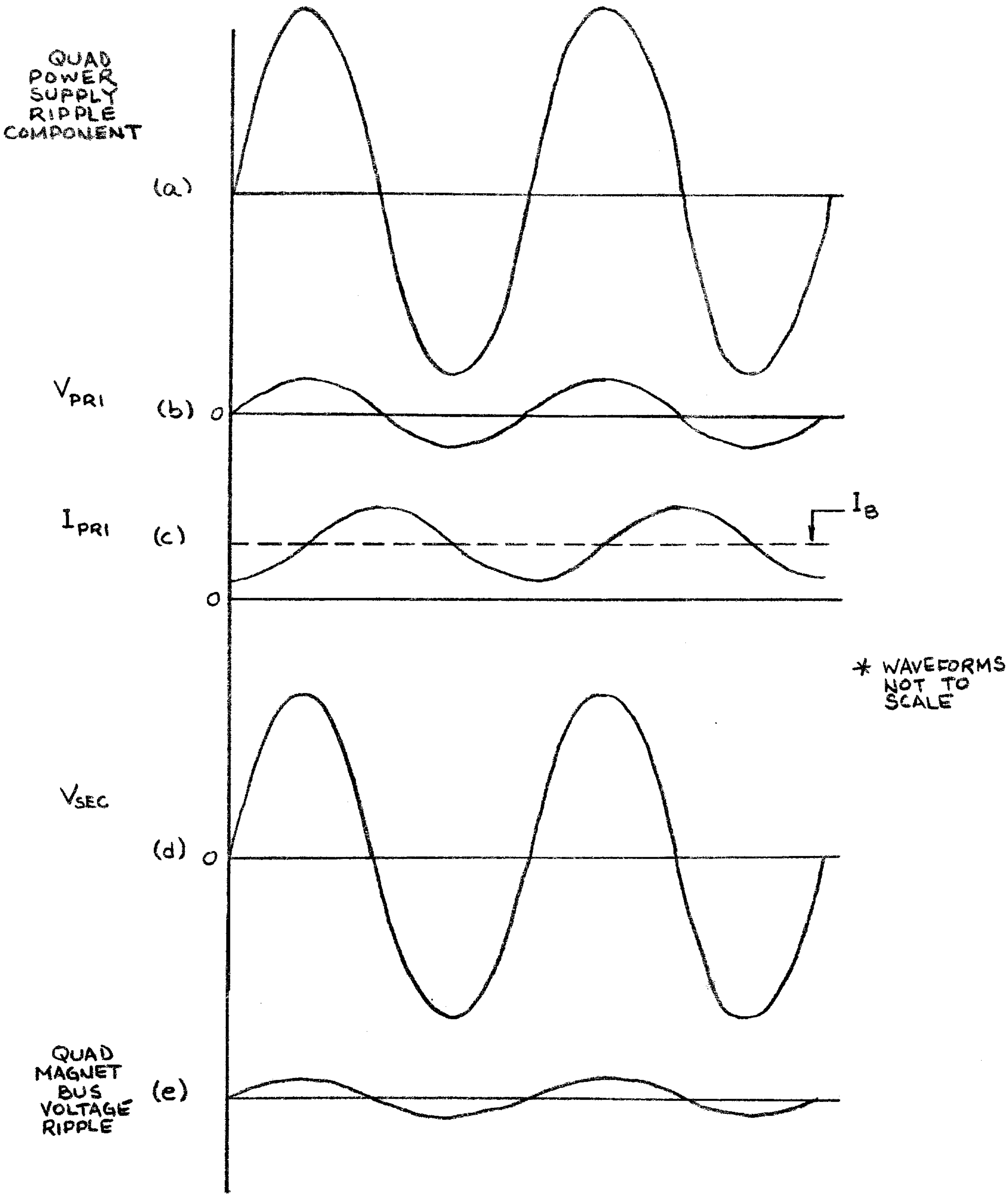


FIGURE 3 - ACTIVE FILTER WAVEFORMS*

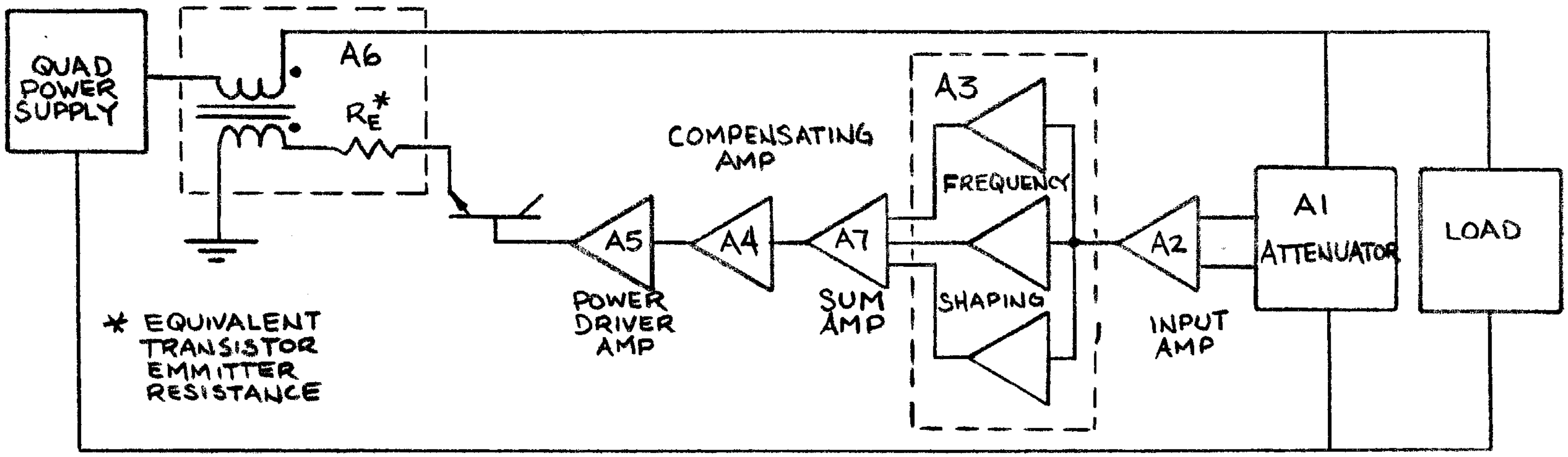


FIGURE 4 – FEEDBACK BLOCK DIAGRAM

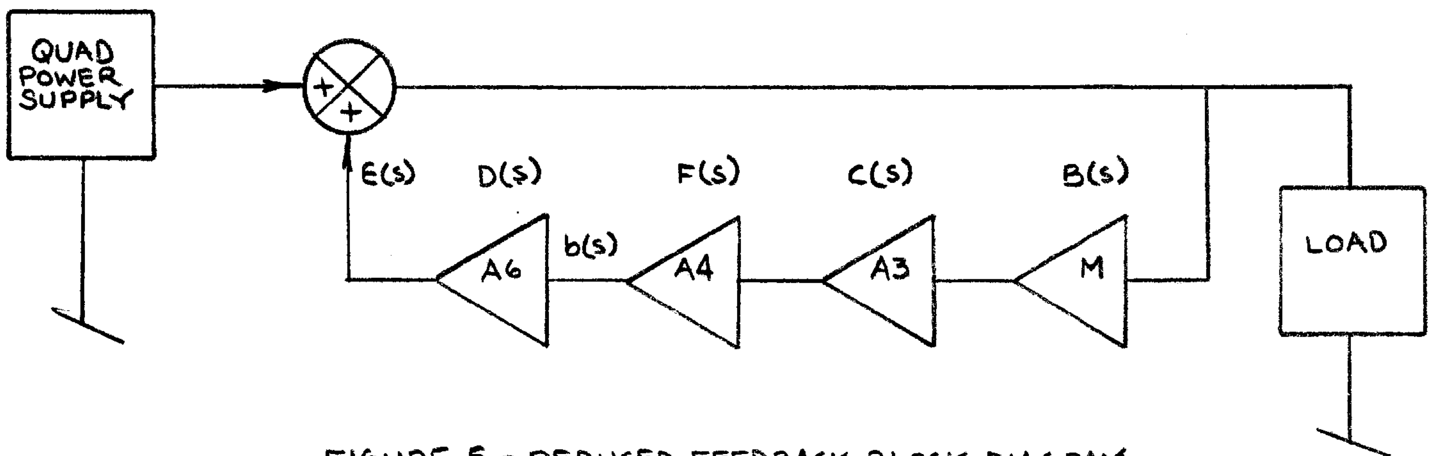


FIGURE 5 - REDUCED FEEDBACK BLOCK DIAGRAM

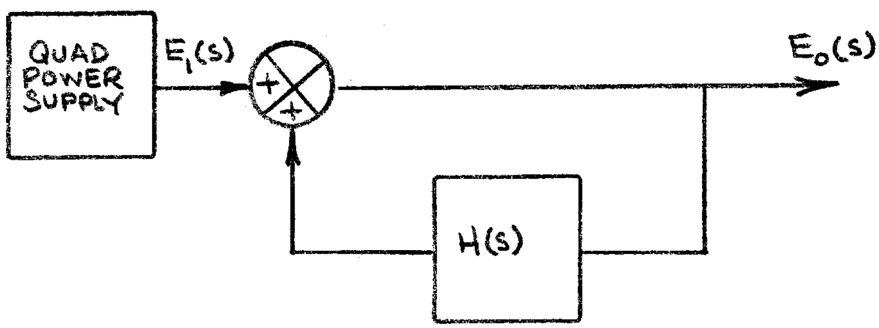


FIGURE 6 - SIMPLIFIED FEEDBACK BLOCK DIAGRAM

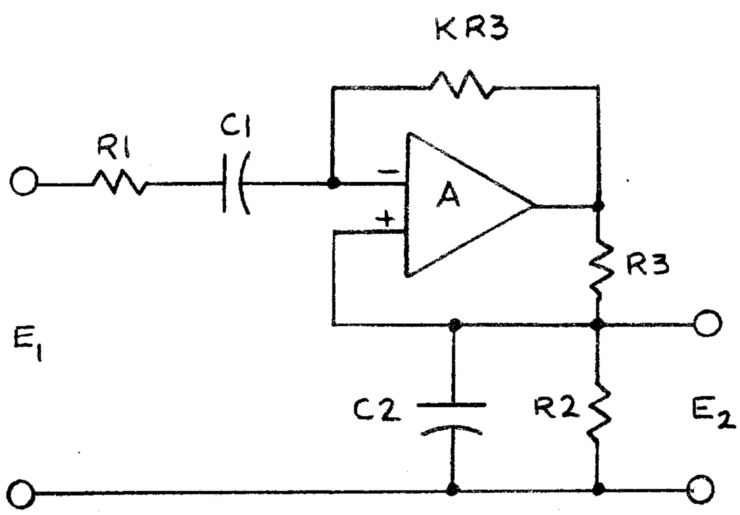


FIGURE 7 - IDEAL NEGATIVE IMMITANCE CONVERTER BANDPASS FILTER

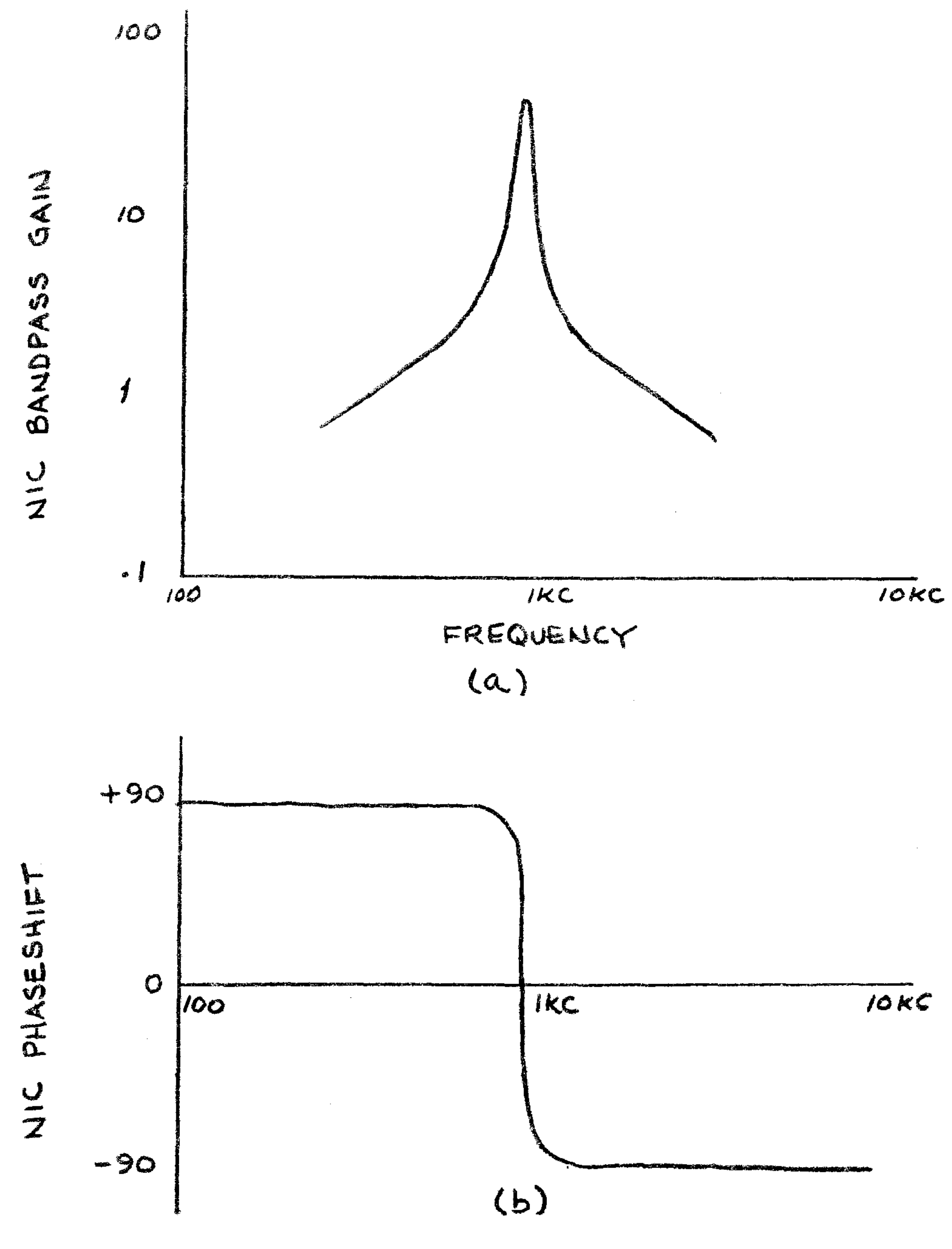


FIGURE 8 – NIC FREQUENCY RESPONSE

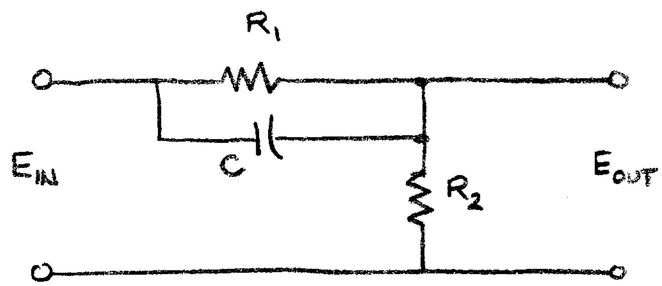


FIGURE 9 – GENERAL PHASE-LEAD NETWORK

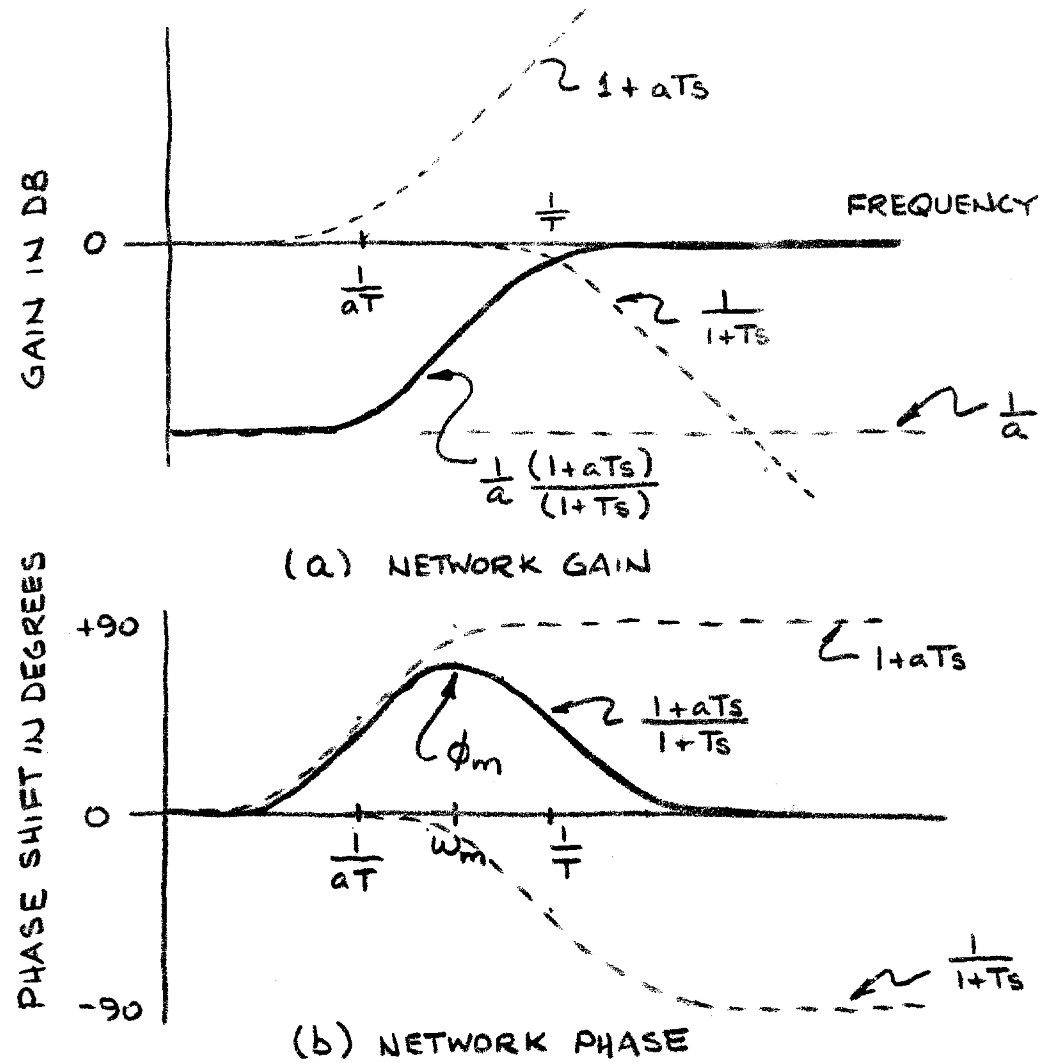
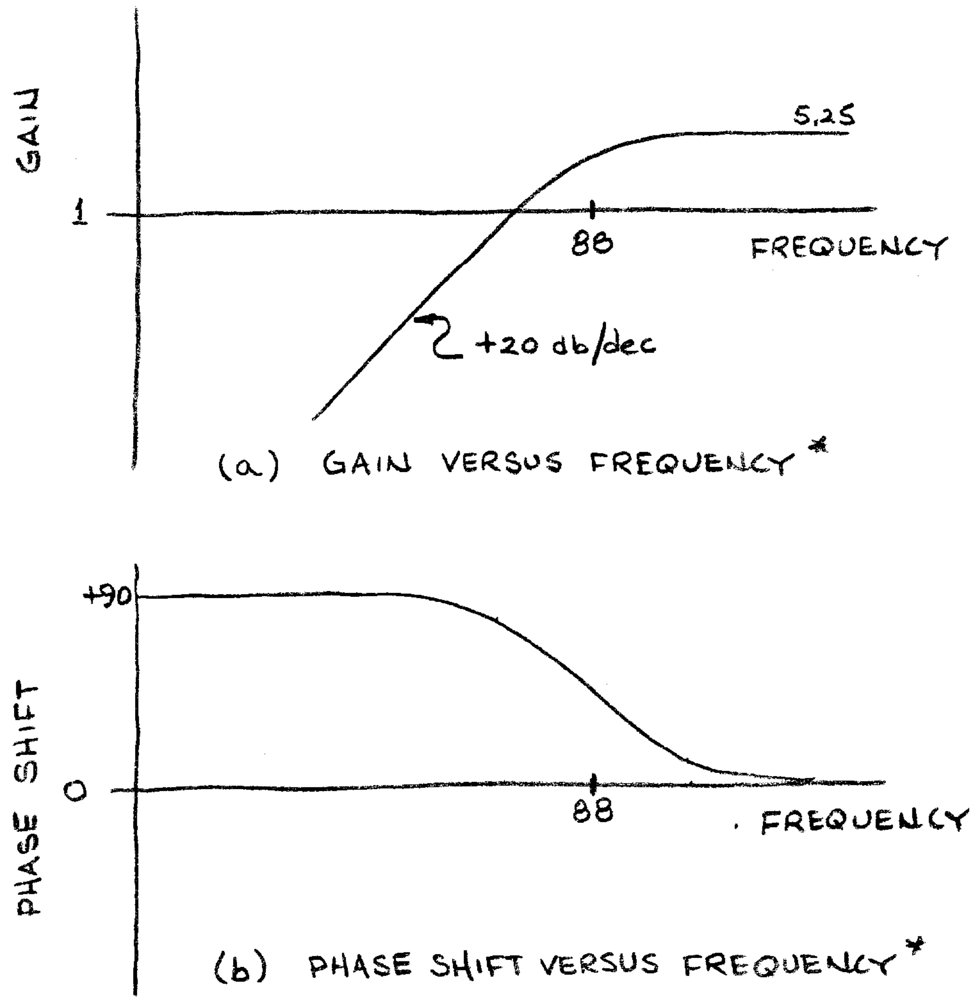


FIGURE 10 – LEAD NETWORK GAIN AND PHASE

SUBJECT | NAME | DATE | REVISION DATE



* FOR 12 MODULES

FIGURE 11 - TRANSISTOR MODULE AND REACTOR-TRANSFORMER FREQUENCY RESPONSE*

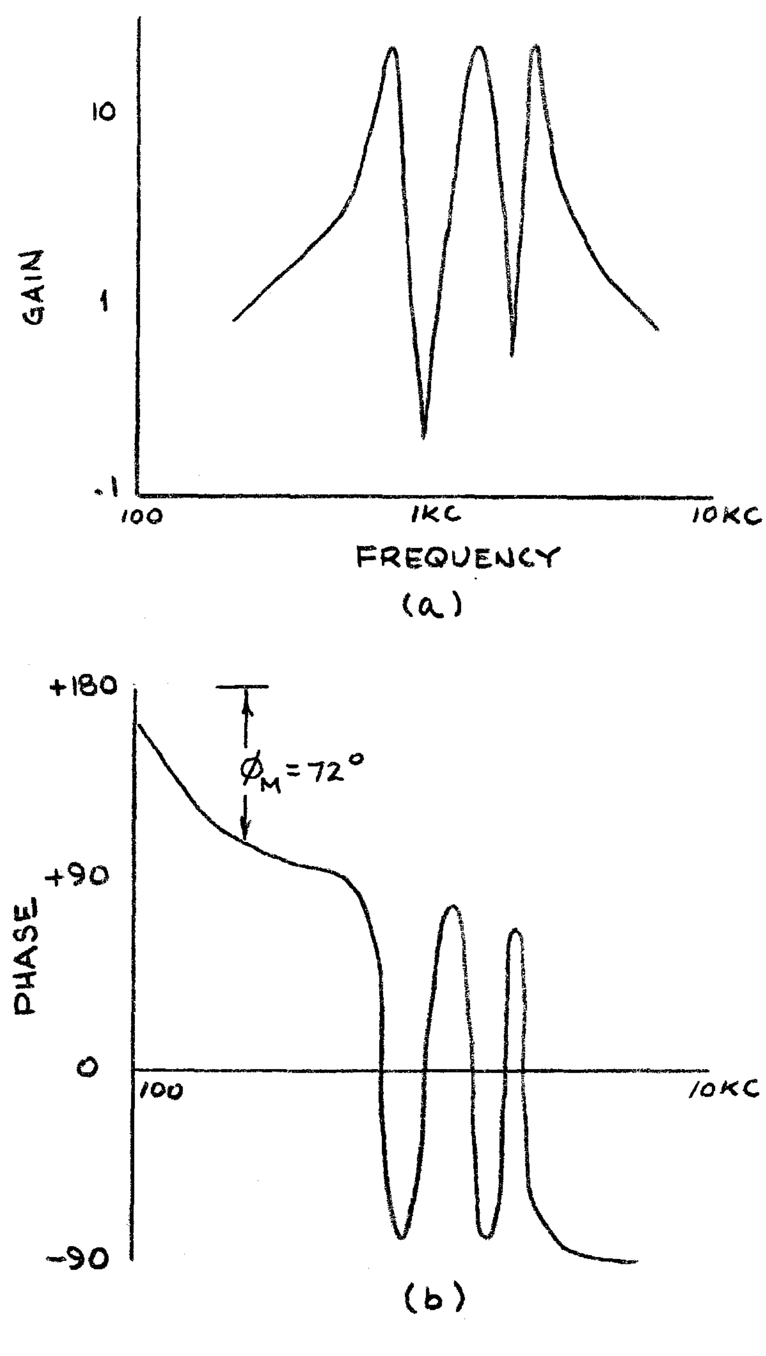


FIGURE 12 - OPEN LOOP FREQUENCY RESPONSE

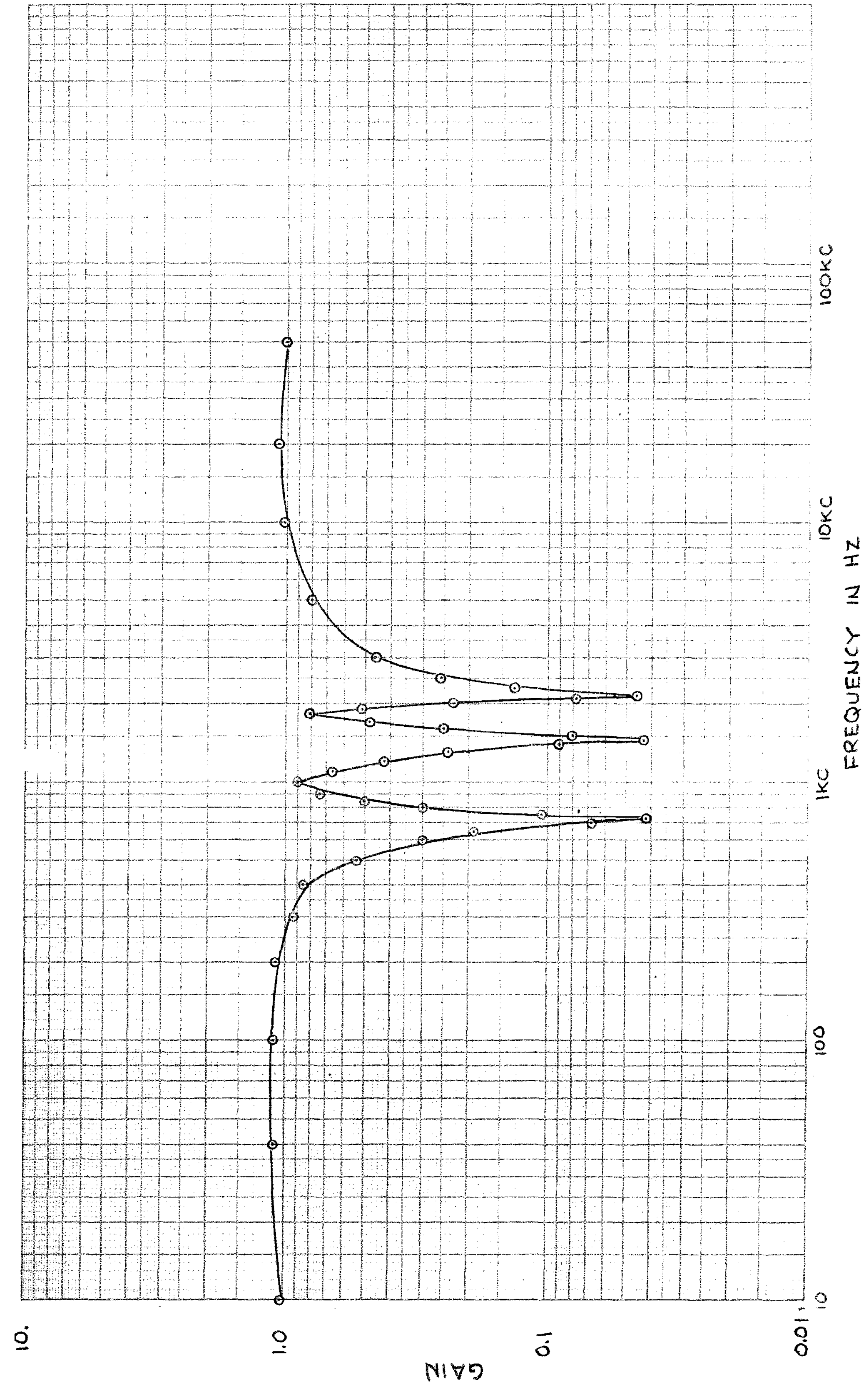


FIGURE 13 - ACTIVE FILTER CLOSED-LOOP RESPONSE

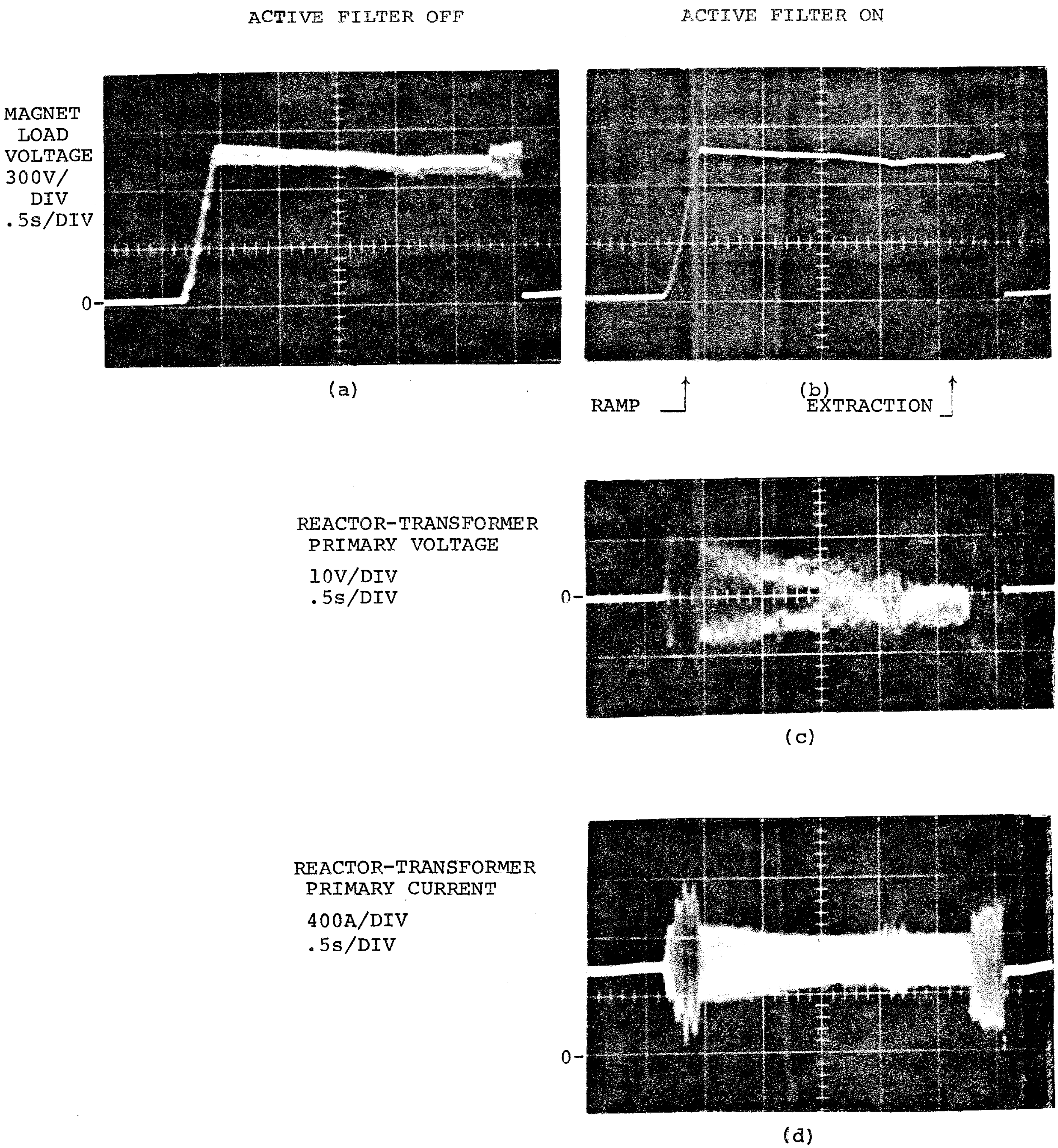


FIGURE 14- QUAD POWER SUPPLY AND ACTIVE FILTER FULL CYCLE WAVEFORMS

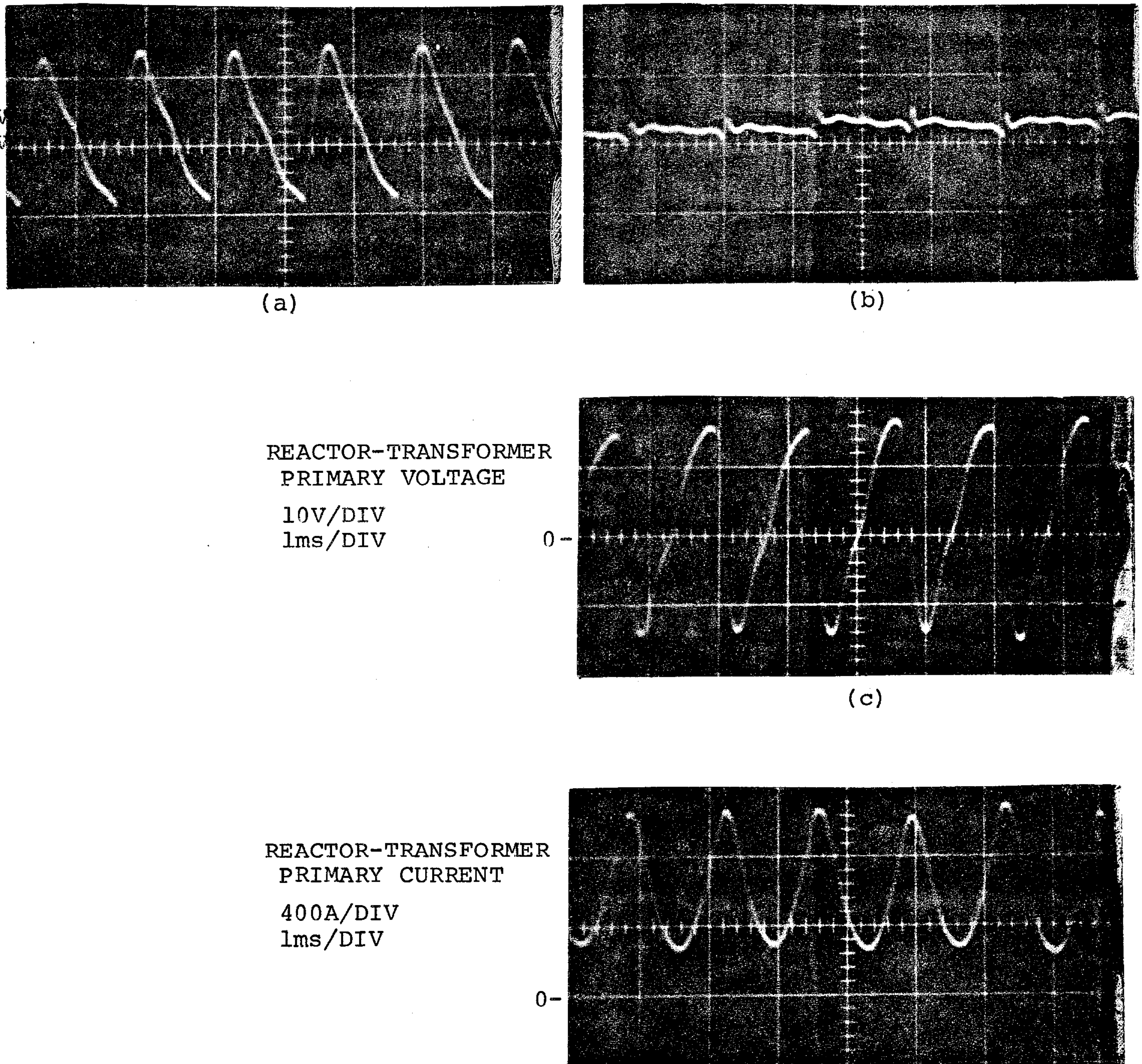
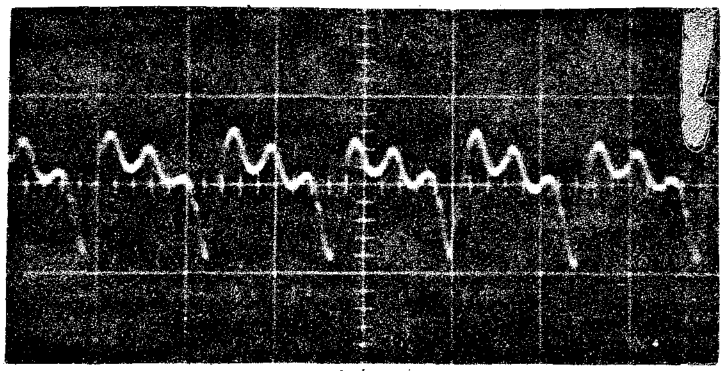


FIGURE 15 - QUAD POWER SUPPLY AND ACTIVE FILTER WAVEFORMS DURING RAMP

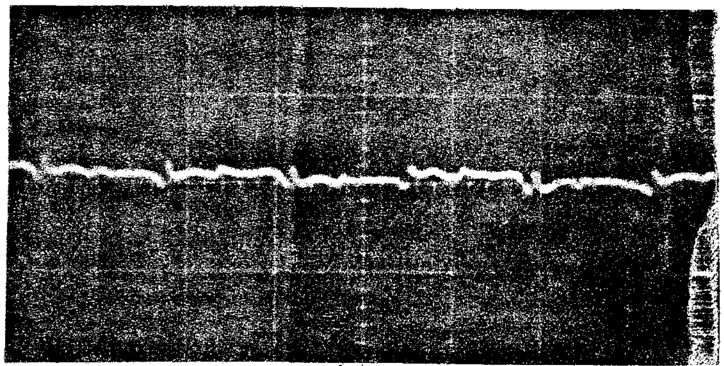
ACTIVE FILTER OFF

ACTIVE FILTER ON

MAGNET
LOAD
VOLTAGE
75V/DIV
1ms/DIV



(a)

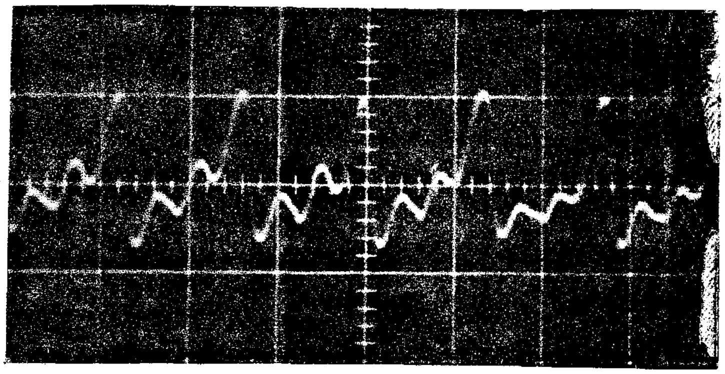


(b)

REACTOR-TRANSFORMER
PRIMARY VOLTAGE

10V/DIV
1ms/DIV

0–

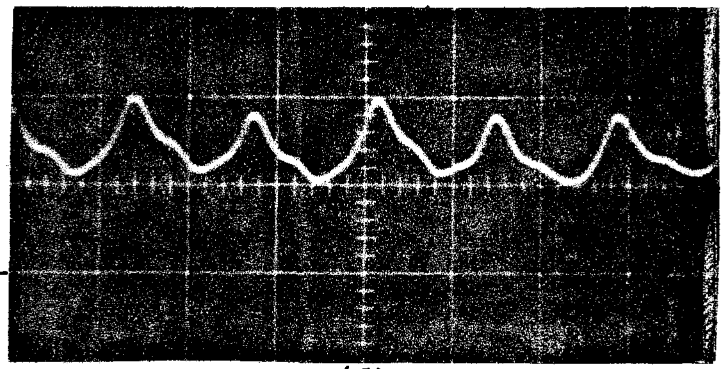


(c)

REACTOR-TRANSFORMER
PRIMARY CURRENT

400A/DIV
1ms/DIV

0–

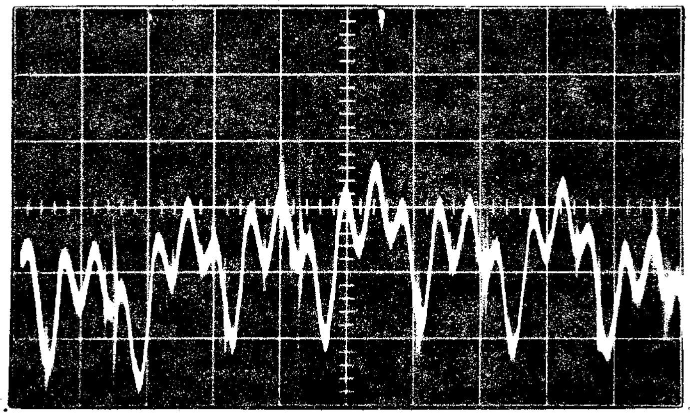


(d)

FIGURE 16 - QUAD POWER SUPPLY AND ACTIVE FILTER WAVEFORMS DURING EXTRACTION

FILTERS OFF

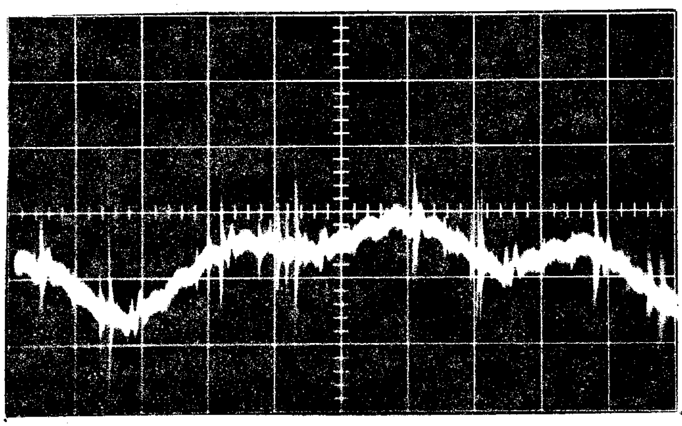
QUAD
BUS
CURRENT
RIPPLE
.25 A/DIV
1 ms/DIV



(a)

FILTERS ON

QUAD
BUS
CURRENT
RIPPLE
.25 A/DIV
1 ms/DIV



(b)

FIGURE 17 - QUADRUPOLE HORIZONTAL FOCUSING BUS CURRENT RIPPLE AT 300 GEV