# AM TRANSMITTER PERFORMANCE WITHOUT FILTER CHOKES

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## COMMON PROBLEMS IN AM TRANSMITTERS DUE TO FILTER CHOKES AND TUBES, ETC.

Some of the problems in transmitter design, manufacture and operation that are related to the iron core components are summarized herewith in order to show the areas of work to be done in eliminating or reducing these problems.

Inasmuch as this paper is premised on the use of solid state rectification and for the present, silicon diodes in particular, we must give notice to the obvious advantages of these rectifiers over the mercury vapor rectifier tubes and their attendant sockets and filament transformers, filter chokes and temperature problems. This comparison has been adequately covered by others and, hence, will not be repeated here.[5]

An easily overlooked problem with mercury vapor tubes is the low ratio of peak to average current capability, which in the MV tube is generally around 4 to 1 whereas in the silicon diode it is 15 or 25 to 1. In high power rectifiers the lower ratio practically precludes the use of capacitor input filters because the capacitor charging currents could damage the tubes even in polyphase rectifiers. This need not be the case when using silicon diodes conservatively. Hence, we have a whole new sphere of operation possible when using solid state rectification in polyphase circuits.

The requirement to use filter chokes in order to protect the diodes from high peak currents, in common rectifier circuits, makes the filter network a high pass filter as seen from the dc terminals. This is a serious disadvantage when one considers that the audio signal of a high-level modulated AM transmitter must pass through this network because the power supply is a series device in the audio power circuit. (Fig. 1)

Further complicating the problem we often find that the filter network can cause a serious low frequency distortion either or both due to parallel resonance of the filter reactor and capacitor and non-linearity of the filter choke inductance when subjected to heavy ac signals.

Additionally the filter capacitor and reactor combine with the modulation transformer, reactor and blocking capacitor network to form an unbalanced mesh that defies resolution because of non-linearity of inductive reactances varying with signal level and frequency. This is generally the culprit when we find distortion problems between 30 and 100 cps when things are otherwise supposed to be working well. (Fig. 4) Many hours of costly labor are expended to bring in transmitters that do not perform correctly at the 50 cycle or lower frequency. The answer quite often is different for systems that are supposed to be alike. Many things are blamed that are not the causes of trouble. One of the most common symptoms of this trouble is the unbalanced modulator plate currents at low frequencies which may be caused by one modulator tube actually working at a higher plate voltage than the other. This is due to the fact that audio signal passing through the power supply capacitor increases the dc voltage on one half of the cycle and decreases it on the other half. (Fig. 1b). This increase and decrease shows up on the modulators in synchronism with the signal, positive for one tube and negative for the other. This problem is minimized when the modulator supply is decoupled from the final amplifier supply either by having two separate supplies or by returning the low end of the modulation transformer to B Neg. instead of B plus, or by using a power supply with low output impedance. (Fig. 1a, Fig. 2).

Another important problem arising out of the use of a filter choke is the fact that when the instantaneous dc current through the choke is changed (due to modulation) that the choke inductance may change as much as two or three to one in value. When additional current is drawn, the inductance value drops much or little depending upon the design of the choke. When the audio signal reduces the instantaneous current through the choke the inductance value may increase. Thus the resonance frequency and Q of the choke will be different depending upon the polarity of the signal. This has the effect of having a different choke inductance when one modulator tube is conducting than for the other. Additionally the low alternating voltage capability of the choke at low frequencies may result in core saturation and, hence, harmonic distortion of the signal. This effect is more noticeable at high modulation percentages than at low percentages due to the fact of higher ac voltage drops across the choke. See Fig. 4 for an illustration of the different reflected load on the two modulator tubes due to power supply filter non-linearity.

Historically the literature concerning the design of modulation systems and transformers has largely ignored the effect that the power supply output impedance can have on the system performance. We have generally assumed that the supply behaved like a battery and had no appreciable filtering effects. Accordingly the classic design of modulation transformers and reactors did not take this into account; in fact the audio designer may be totally ignorant of the power supply and filter characteristics. Little wonder then that chaos has often reigned in the factory before a transmitter is born.

## OUTPUT IMPEDANCE OF POWER SUPPLY[9]

The output impedance of the power supply may be calculated as follows: Refer to Fig. 3.

At any frequency $$Z = \left[\frac{(R_1^2 + \omega^2 L^2)\,(R_C^2 + \frac{1}{\omega^2 C^2})}{(R_1 + R_C)^2 + (\omega L - \frac{1}{\omega C})^2}\right]^{\frac{1}{2}} \text{ ohm}$$

Maximum impedance occurs at resonance where Z reduces to

$$Z = \frac{R_1 R_C + \omega^2 L^2}{R_1 + R_C} \quad \text{ohms}$$

and when $LC = 1$

where L is equal to the sum of all series inductances.

where $R_1$ is equal to the sum of all resistances in the inductive branch.



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All values referred to the dc impedance level.
Minimum low frequency impedance approaches $R_1$.
Minimum high frequency impedance approaches $R_c$.

In order to minimize the resonant impedance one should;

(1) Keep inductance values at an absolute minimum.
(2) Use as small a value filter capacitors as is commensurate with other requirements.
(3) Add a resistance in series with the capacitor at least equal to the sum of the other resistances.
(4) Eliminate the filter reactor and use higher rectification phases to minimize the ripple before filtering.

Example: Fig. 1a. In a typical 5 KW transmitter using a 10 KW power source at 5000 volts dc and an average load current of 2 amperes the power supply and source impedance on a low impedance transformer would be about 3%. This would be further divided up to 1% series resistance and 2.9% reactance. Average load resistance on power supply is $\frac{5000}{2} =$ 2500 ohms, so 1% resistance is 25 ohms and 2.9% reactance is 72 ohms. $L_s$ effectively is equal to $\frac{72}{2\pi \times 60} = 0.19$ henry.

A conventional system may have a 3 henry 10 ohm filter choke and 10 mfd of capacitor, with no resistance in series with the filter capacitor.

$L = 3.0 + 0.20 = 3.2H.$

At resonance $\omega^2 = \frac{1}{LC} = \frac{10^6}{3.2x10} = 31000$

$\omega = 177$ and $f_r = \frac{127}{6.28} = 28$ cps which is not usually measured and plotted on the frequency response and distortion curves.

At resonance $Z = 9000$ ohms.

At 100% modulation the rated secondary current is 1 ampere rms, which obviously would not pass through the filter network at any frequency approaching the resonant frequency of 28 cycles. This is why we usually return the modulation transformer secondary current through a bypass to ground. The modulator (primary) currents cannot escape, however, but the signal frequency at the primary center tap is doubled, hence, the response above 40 or 50 cycles is not much affected. The dangers, however, lie in the switching transients which can place very high reverse voltage transients on the diodes. See Fig. 5.

At 50 cycles the output impedance of the above network turns out to be 1450 ohms. Thus a 25 cycle 1.4 ampere signal would encounter 2000 volt rms or 2800 volt peak drop at 100% modulation due to the primary modulator currents alone. The 1 ampere secondary current also encounters a 1550 ohm impedance in the customary 2 mfd blocking capacitor, giving rise to a 1550 volt drop in the capacitor. Thus we can see that the modulator tubes see a large series impedance in addition to the normal load of about 3600 ohms resistance, when a filter reactor is employed. This leads to many undesirable results readily apparent to the transmitter engineer.

When the filter reactor is eliminated and a 12 or more phase rectifier is employed (Fig. 2) the transformer parameters remain the same but the total L and C are changed. The L value drops to 0.20 henry and we can use as little as 4 mfd of capacitance. We will choose a 100 ohm value for $R_c$ as this approximates the low frequency impedance of the supply.

Applying these new numbers to the output impedance formulae given above we find that

$\omega^2 = \frac{1}{LC} = \frac{10^6}{0.20x4} = 1.25x10^6$ and $\omega = 1120$ so $f_r =$ 178 cps which, of course, is in the pass band.

At resonance $Z_r = \frac{R_1 R_c + \omega^2 L^2}{R_1 + R_c} =$

$\frac{25x100 + 1.25x10^6 x.04}{25 + 100} = 420$ ohms

At frequencies above and below 178 to 90 cycles the series impedance will be less and there will be no problem in passing clipped or square waves or long duration pulses through the power supply.

At 50 cycles $Z = \left[\frac{(25^2 + 3960)(10^4 + 635000)}{(25 + 100)^2 - (314x.20-800)^2}\right]^{\frac{1}{2}} =$ 75 ohms.

Thus we see that the series impedance at 50 cycles is on the order of 3% of the circuit impedance.

At frequencies below 50 cps the output impedance will approach a limit where the inductive reactance is negligible and the series resistive elements of the transformer secondary and of the diodes are the basic impedance.

At frequencies above resonance an increasing proportion of the signal current will transverse the RC branch of the network as the capacitive reactance approaches small values. The high frequency minimum impedance, of course, is the series resistance $R_c$. The resistor $R_c$ should be capable of handling the total ac current and the losses should not be excessive for the system. This resistor should have very little inductance so as not to impede transient current flow.

The maximum ac current in the RCL network occurs near resonance and the value will be equal to (Q x signal current) through the network where ($Q = \frac{\omega L}{R}$). R being the total effective series resistance in the resonant circuit.

The determination of the capacitor size is dependent upon several factors. One factor is the need for transient suppression which is greatly aided by the RC network across the dc terminals. Another factor is the amount of ripple and the line unbalance reduction needed. A third is the reduction or elimination of commutating short circuits in the rectifiers. A fourth is the output impedance requirements of the system. The complete analysis of these factors is under consideration and will be the subject of a future paper.

## NOISE

It has been found in practice that the amount of ripple resultant from an RC network across the dc terminals that is based on the amount of capacitance needed to keep the maximum transient due to switching down to 150% of the dc voltage, is also adequate to reduce the transmitter noise to less than -60 db. This presumes a nominal amount of line unbalance also.

## POWER SUPPLY REGULATION

The author believes that it is highly advantageous to use low reactance power supply system for a transmitter for the following reasons. Fig. 2.

It is generally conceded that it is less expensive to use a common power supply for the final amplifier and the modulator provided interactions can be minimized sufficiently. The stability of the modulator voltage is enhanced by the combined power rating of the two loads, especially when the one is steady. The problem with a common supply is usually two-fold. (a) The varying modulator current causes the dc voltage to vary and, hence, carrier shift can ensue due to the variation of the final input voltage. (b) The output impedance of the power supply is common to the modulator circuit and the final amplifier circuit in the audio system and can be a cause for distortion of signal.

It is quite obvious that the problem then is relative rather than absolute because both problems are based upon power supply impedance. Thus if the power supply regulation is low enough then the carrier shift tolerance will not be exceeded, and if the output impedance is low enough the distortion limits will not be exceeded. When filter chokes are used in the filter network the signal current is also largely prevented from traversing the choke and transformer rectifier path due to the high impedance path of the choke, hence, the filter capacitor alone must carry the signal current from one dc terminal to the other (see Fig. 1a). Economics usually limits the impedance of the capacitor path but also it is undesirable to store much energy in the capacitors. As a result, we find that we have an output impedance characteristic (see Fig. 3) having a very high impedance from 40 or 50 cps down. This problem is further compounded by the fact that the shunt reactance of the modulation transformer and inductor is also beginning to increase the modulator currents at frequencies below 50 cps. We, therefore, have the problem of increasing current and increasing series loss impedance at one and the same time when we employ a filter choke in the usual mode.

By eliminating the filter choke (Fig. 2a) and employing low leakage reactance power transformers the problem is largely resolved as may be seen by the other curves in Fig. 4. It is equally obvious that the curves for a 3% reactance transformer are substantially preferable to those for a 10% reactance transformer. There are also other problems in commutation and ripple that arise with the higher transformer reactances.[11]

Inasmuch as it is practically unavoidable to have the effects of power supply resonance and the resultant higher insertion loss at certain frequencies occur in the pass band it would be reasonable to accept the fact and plan for it rather than to try to brush the problem under the carpet. It is better to accept the facts of a small percentage insertion loss in the power supply and choose a set of parameters that will give the most uniform performance over the specified band width. It is necessary to have a surplus of power available from the modulators anyway unless we accept distortion and overloading at certain frequencies.

When we use an appropriate value of resistance in series with the filter capacitor, this becomes a minimum value for all frequencies above resonance and, hence, will tend to improve the linearity of the performance at the high frequency end of the pass band as well. It appears possible to obtain an excellent series impedance characteristic by using an $R_c$ value of 5 to 10% of the total load resistance.

At the lower frequencies the increasing currents due to the modulation transformer and reactor inductance is partly balanced off by the decreasing series loss impedance of the power supply, resulting in an extended range of excellent low frequency performance. Current reports from the field indicate an improvement of almost one octave additional low frequency useful range using the same modulation components and merely changing power supplies. No problem from resonance was encountered.

The matter of carrier shift using a common power supply is cared for by the relationship that the shift is approximately in proportion to the voltage deviation. When we have a power supply with an overall impedance of 5% or better, which is easy to obtain, we have a voltage variation due to modulation of less than 2½% because half of the power supply load is steady. Hence, the carrier shift of less than 3% is obtained with a common power supply having a regulation of 5% or less.

RECTIFIER SELECTION

Of course, a low impedance power supply calls for larger and more expensive diodes but I will try to show that the cost of this proposed system is in fact more economical as well as superior to the others which include the use of chokes and also high reactance supplies.

There are three basic factors among others that control the selection of rectifier diodes. The peak reverse voltage that the diodes must withstand, the peak fault current and the duration of it, and the continuous maximum thermal rating. One of the most troublesome factors in the present practice of diode selection is that many designers select diodes on the basis of average current capability as we did with rectifier tubes and selenium cells, and with little attention being paid to possible reverse voltage transients and fault current values and duration. They start with a diode system that will carry the average dc current and then (maybe) specify the circuit impedance to fit that diode system. This often results in poor regulation and high transients due to series inductance in the power supply. Thus failures due to switching transients may very likely occur because of the large amount of energy stored in the filter choke or series leakage inductance. To offset this you need to buy more diodes to increase the PRV rating of the diode string. So now we have a choice between putting our money into more small diodes or fewer large diodes. More properly one should determine in advance just what the regulation of the supply should be and give a tolerance on this value. The long term average power consumption of the system can usually determine the rating of the power transformers. With this data one can determine the fault current that a diode string may see and the overload protective device determines the duration of the fault (short circuit of power supply). Overload devices are readily available that will limit the fault duration to 20 to 100 milliseconds and fuse type protection is available that will act faster.

One then selects diodes that will meet the fault current magnetude and durations. This diode will generally be more than adequate to meet the steady state demands of the system. This arrangement has the basis for high reliability and long life for the diodes because the diodes can be operated at a very small temperature rise far below the diode temperature limits. We are told that the life expectancy of a diode is a function of magnetude and number of thermal excursions. When the thermal excursion is relatively low, as per the herein recommended basis, the number of excursions per life time is extremely high. Thus the lifetime of the rectifier stack is adequately long, providing voltage transients are under control. This, of course, eliminates a diode replacement charge. Furthermore, we are told that a cool running diode will withstand more transient voltage than a hot one.

The proper selection of the diode string peak reverse voltage rating is a rather complex procedure unless one wishes to pick an adequately safe figure with your eyes closed. This, of course, is not a satisfactory engineering procedure. A common practice is to use a stack rating of 150 to 250% of the working PRV of the circuit, depending upon the transient protection used.

At present there are two basic diode types to choose from and it would appear that the safety factor figure can be quite different depending upon the diode type. The presently most common type has no appreciable capacity for reverse current and, hence, no transient absorbing capability. Transient absorption must be completely provided for aside from the diodes. The other type is the bulk avalanche diode which has a zener voltage range and will safely handle or pass certain amounts of transient current for specified times. They also have a certain continuous reverse current power rating. This capability means that low power transients and part of higher power transients can be avoided by providing a discharge path within the avalanche voltage rating of the diode string for the stored energy responsible. Transients other than switching transients may not be cared for this way.

## BLOCKING CAPACITOR

A final item to consider is the choice of ($C_b$) blocking capacitor in the (Fig. 2) secondary circuit of the modulation transformer. Ostensibly the purpose of the capacitor is to block the dc current from the modulation transformer in order to avoid core saturation. For many years it has been customary to make the shunt inductance of the secondary of the modulation transformer approximately equal to the value of inductance of the modulation reactor.[1] A capacitor is then selected which will correct the power factor of the network at some low frequency usually between 20 and 60 cps. This, of course, is like a network and has the impedance characteristic of one. Economics dictates the maximum open circuit inductance of the reactor and, hence, the necessity of employing resonance to get a reasonably good low frequency performance.

The advent of peak clipping techniques in the audio amplifier network, for the purpose of increasing the average audio power content in the modulated carrier, has required a new look at all of the series and shunt devices in the audio system. The result of the peak clipping is flat topped signal excursions requiring the same treatment as in pulse techniques. In order to avoid excessive droop in the flat top portion of the wave and, hence, additional distortion we must have much lower frequency capabilities than heretofore.

A consideration of Fig. 5 will indicate the necessity of avoiding a filter choke in the power supply or even a high reactance transformer because of the high power supply resonant impedance at the low frequencies. When speech clipping is employed it would appear to be a distinct advantage to have the resonant frequency of the power supply well above 50 cycles so there will be no interference from the power supply on the flat top characteristic of the signal.

The function that describes the charge or voltage drop on capacitor as a result of a step-function change in current

tion change in current $V = E(1-\epsilon^{-\frac{t}{RC}})$ where E is the total voltage in the function, and t is the lapsed time involved and R and C being the controlling values of components.[2] Assuming that we have assigned values of voltage, time and resistance, the variable C has a marked effect on capacitor droop:

For example; if $E = 10000$, $t = .001$ sec. $R = 1000$ ohms and $C = 10\text{x}10^{-6}$ fd.

Droop $V = 10000\ (1-\epsilon^{-\frac{.001}{1000\text{x}10\text{x}10^{-6}}})$ and $X = 0.10$ $\epsilon^{-x} = 0.9048$ $V = 10{,}000 \text{ x } .0952 =$ 952 volts or $9\frac{1}{2}$% which is a rather unacceptable value. If C is increased to 50 mfd $X = .02$ and $\epsilon^{-x} = 0.9802$ $V = 10{,}000 \text{ x } .02 =$ 200 volts of 2%.

If we are only concerned with sine waves we can compute the reactance of the capacitor at the lowest frequency of interest and multiply this value by the square of the maximum signal current that it has to pass to obtain the reactive voltamperes involved. This product should only be equal to a minor fraction of the audio power if we wish to avoid wide swings in the load reflected to the modulator tubes. It is quite common in the existing transmitters for this capacitive voltamperes to equal and exceed the power rating of the modulator within the pass band of the transmitter because these capacitive voltamperes will be counter balanced by the inductive voltamperes of the modulation transformer and reactor. The problem is, of course, that at other frequencies this counterbalancing is not complete so that there can be a large value of reactive voltamperes in the tube load. So actually one can compute the value of capacitor needed to offset the inductive reactance by selecting a frequency for resonance and, hence, equality of reactive voltamperes and computing the inductive voltamperes using the known inductance values.

Now, however, with the advent of low output impedance power supplies, particularly at low frequencies we can do away with the $\pi$ network concept and economically design the modulation transformer to have an inductance several times higher than the modulation reactor inductance knowing that we can use a large value of blocking capacitor as in circuit Fig. 2b without excess cost. In circuit (b) the capacitor does not have to block the total dc voltage but only a very small dc voltage drop plus whatever ac voltage drop occurs due to the passage of audio current. Thus we can use a low voltage ac type capacitor of a relatively large capacitance and, hence, a very small ac or pulse drop across it. This type capacitor is usually much less expensive than the high voltage dc capacitor.

It can be shown that if the shunt inductance values of the modulation transformer and reactor are chosen for good 50 cycle performance without the necessity of power factor correction by capacitors, that this value will be also suitable for the requirements of low phase shift and droop under the 4 millisecond flattop condition specified for the speech clipping or trapezoidal signals. Therefore, we can say that the inductance values are determined by the minimum frequency requirements for 100% sine wave operation whereas the blocking capacitor value is determined by the allowable droop of the capacitor under flattop signal conditions.

The droop in voltage due to shunt inductance is adequately covered in papers[8-1] on pulse transformer design under flattop considerations. The shunt inductance of the transformer and reactor in parallel is the total inductance involved.

It is quite feasible to have the combined droop over a 3 to 4 millisecond period kept well below a 10% figure, making it unnecessary to use compensating pre-distortion in the signal in order to meet an overall droop limitation, when using the power supply without a filter reactor.

The use of $C_b$ in the (b) position shown in circuits requires a word of caution, however. Any ripple components present across the filter capacitor will show up with very little attenuation at the RF load and may result in excess noise level. This would probably be due to unbalanced 3 phase incoming line voltage or unbalanced impedance in the dc supply transformer. When $C_b$ is returned to B- or ground, the modulation reactor will serve to attenuate the ripple adequately before it reaches the load, due to the relatively high reactance of the inductor. When, however, the B+ is capacitively coupled to the secondary of the modulation transformer there is very little impedance in this path to the RF load. As a result it probably is more economical to buy the more expensive capacitor $C_b$ for circuit (A), for the smaller transmitters, than to arrange for nicely balanced three phase lines, as required for circuit (B). The 720 cycle ripple will usually be less than -50db due to the attenuation in the filter and leakage network but the 120 cycle ripple that occurs due to line unbalance is not diminished appreciably. For larger transmitters, then, depending on the total ripple to be -50db, it will probably be necessary to incorporate a line balancing device to reduce the noise from that source if the unbalance is likely to be in excess of limits.

## CONCLUSION

The elimination of filter chokes from the transmitter power supply can bring relief to the historically vexing problems of impedance mismatch, distortion, complex computation and performance at the low frequency end of the audio spectrum. The system herein discussed allows for vastly improved performance on clipped signal operation and extends the smooth performance range to lower frequencies with the same basic iron core components. Simplification of computations and construction is also afforded. The knowledgeable coordination of the power supply with the high power audio stage is urgently recommended.

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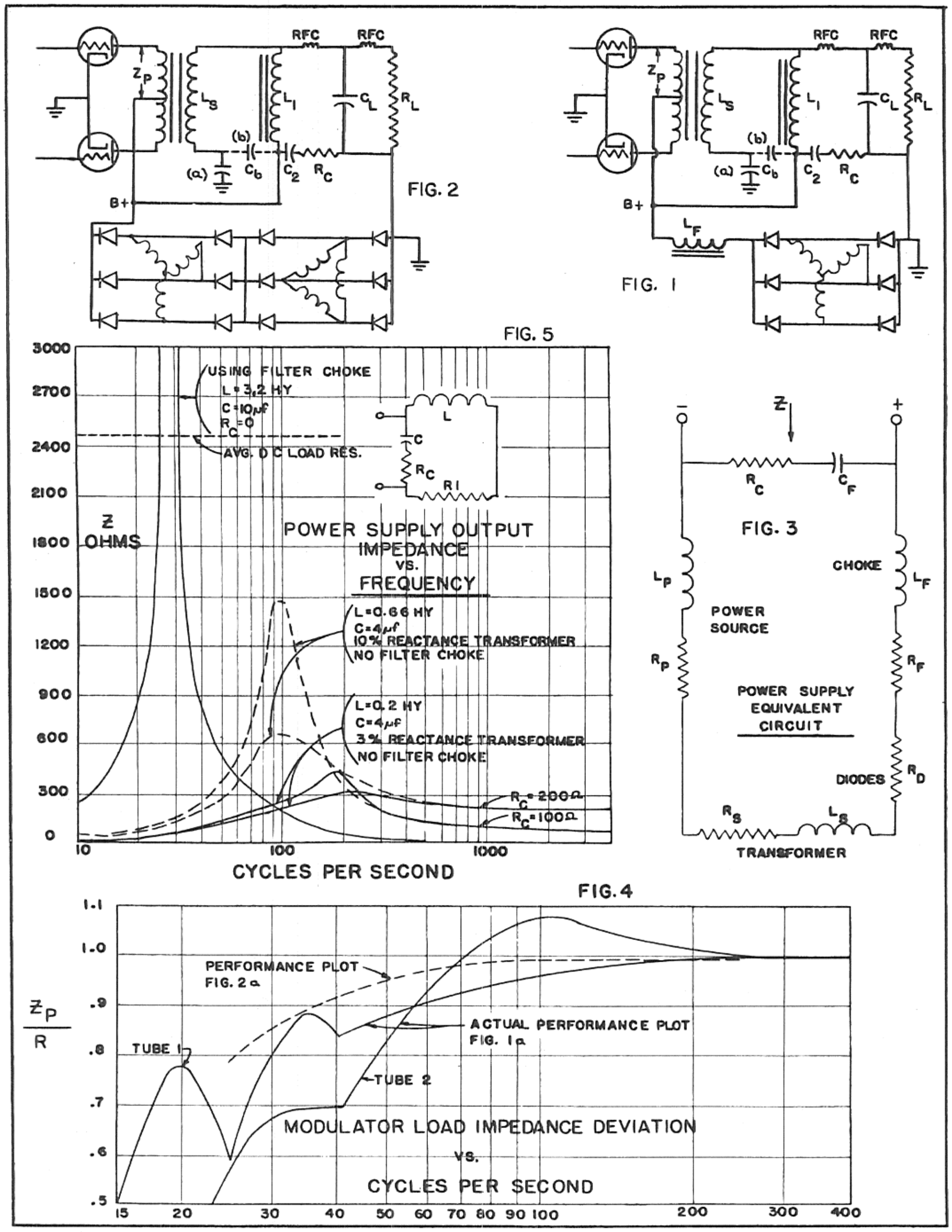
RFC
RFC
$Z_P$
$L_S$
$L_1$
$C_L$
$R_L$
(b)
(a)
$C_b$
$C_2$
$R_C$
B+
FIG. 2
$L_F$
FIG. 1
FIG. 5
USING FILTER CHOKE
L = 3.2 HY
C = 10 µf
$R_C$ = 0
AVG. D C LOAD RES.
Z OHMS
POWER SUPPLY OUTPUT IMPEDANCE VS. FREQUENCY
L = 0.66 HY
C = 4 µf
10% REACTANCE TRANSFORMER
NO FILTER CHOKE
L = 0.2 HY
C = 4 µf
3% REACTANCE TRANSFORMER
NO FILTER CHOKE
$R_C$ = 200 Ω
$R_C$ = 100 Ω
L
C
$R_C$
R1
3000
2700
2400
2100
1800
1500
1200
900
600
300
0
10
100
1000
CYCLES PER SECOND
Z
$C_F$
FIG. 3
$L_P$
POWER SOURCE
CHOKE
$R_P$
$R_F$
POWER SUPPLY EQUIVALENT CIRCUIT
DIODES
$R_D$
$R_S$
$L_S$
TRANSFORMER
FIG. 4
$\frac{Z_P}{R}$
PERFORMANCE PLOT FIG. 2a
ACTUAL PERFORMANCE PLOT FIG. 1a
TUBE 1
TUBE 2
MODULATOR LOAD IMPEDANCE DEVIATION VS. CYCLES PER SECOND
1.1
1.0
.9
.8
.7
.6
.5
15
20
30
40
50
60
70
80
90
100
200
300
400