SECTION 3

# SPECIALIZED BROADCAST RADIO ENGINEERING

## FREQUENCY-MODULATED TRANSMITTERS


Capitol Radio Engineering Institute
Washington, D. C.

312E

BROADCAST TECHNICAL ASSIGNMENT

FREQUENCY-MODULATED TRANSMITTERS

## - TABLE OF CONTENTS -

# BROADCAST TECHNICAL ASSIGNMENT

# FREQUENCY-MODULATED TRANSMITTERS

## INTRODUCTION

The previous assignments on Frequency-Modulation Receivers and on Modulation in the Advanced Practical Radio Engineering course, dealt with the fundamental principles of frequency modulation. Such matters as its superiority with respect to interference suppression, various methods of modulating the transmitter, receiver design, and the various types of f.m. detectors, were covered in that section.

In this assignment more detailed information concerning commercial transmitters will be given. It will be of advantage for the student to review the fundamental principles already studied before reading this assignment, particularly the one on modulation.

*FREQUENCY RANGE.*—The f.m. band covers the range of 88.1 to 108.1 mc, in channel steps of 200 kc. In any given area, stations are separated by at least 800 kc in order to obviate the danger of adjacent channel interference where the signal levels are high. Such separation enables less-selective and hence cheaper as well as more reliable receivers to be built; this is particularly important when a receiver requires relatively expensive components to deliver the high-fidelity reception of which f.m. is capable.

At such high frequencies as 88.1 to 108.1 mc, a band width of 200 kc is not unduly great, and receiver circuits of normal Q can cover the channel. The frequency is not so high that lumped resonant circuits are impractical, so that bulky, expensive, and mechanically complicated tuning arrangements are unnecessary. Moreover, the modern small midget tube can operate very successfully at these frequencies; transit-time and cathode-inductance loading can be kept down by the small dimensions of the tube.

In the case of the transmitter, the high power required introduces difficulties in the design of the power tubes. This is particularly true in the case of a 50-kw transmitter. An aid in obtaining such high power is the use of a grounded-grid final stage. The input power required by the stage is high, but this power is not wasted entirely in grid dissipation; a large portion of it appears as useful output power, and thus reduces the amount of power output required of the tubes in the final stage itself.

*METHODS OF MODULATION.*—Fundamentally there are two methods of frequency modulation: the phase-shift method and the frequency-shift method. In the former method, a crystal-controlled oscillator produces a fixed frequency, whose phase angle is then varied in accordance with the modulation. By making the amplitude of the latter vary inversely with its frequency, the inherent phase modulation is converted into frequency modulation.

If the phase shift is small, considerable multiplication, as well as heterodyning, is required to obtain sufficient frequency deviation at the desired carrier frequency. If greater phase shift at low distortion can be obtained, less multiplication, and no heterodyning is required. This in turn imposes less stringent requirements on the frequency stability of the initial crystal-controlled oscillator, although special circuits or tubes may be required.

One method of obtaining greater phase shift is by means of the phasitron tube, as employed by G.E.; another is by means of cascading the phase shift stages, as is done by Raytheon. In this system, the phase-shifted carrier, as produced by a suitable circuit, is then applied to a second similar stage, thereby increasing the phase shift, and so on through four stages, until sufficient total phase shift is obtained to require ordinary frequency multiplication and no heterodyning.

In the direct frequency-shift systems, as employed by Western Electric and R.C.A., an ordinary oscillator has its frequency varied by reactance tube modulators. The frequency deviation thus obtained is sufficiently great that a moderate amount of subsequent multiplication enables 75 kc final deviation to be obtained on the output carrier.

The same initial f.m. wave is also demultiplied in a separate channel to a frequency sufficiently low so that practically only the sub-carrier remains. This is then compared with a crystal-controlled oscillator, and the beat frequency resulting from any drift is then applied to a synchronous motor, which operates an auxiliary tuning capacitor in the tank circuit of the oscillator, thereby bringing it back into synchronism with the crystal-controlled oscillator.

Any of these systems are capable of meeting the F.C.C. requirements as to constancy of the carrier frequency, low distortion, frequency deviation, and audio fidelity, so that the choice of equipment depends mainly on such competitive factors as price, simplicity of the equipment, service facilities, and the like.

## THE IMPROVED ARMSTRONG SYSTEM

It will be recalled from a previous assignment that the original Armstrong system depended upon two crystal-controlled oscillators for its frequency stability: the initial oscillator that was phase-modulated, and the heterodyne oscillator required to lower the carrier frequency without affecting the deviation, used just prior to the final multiplication.

*USE OF DUAL CHANNEL F.M.*.—By a new method, Armstrong has been able to get away from the dependence of the frequency stability on the initial oscillator, and to rely solely on one crystal-controlled oscillator, that required for heterodyning. The method, as employed by R.E.L. (Radio Engineering Labs, Inc.), uses a balanced modulator which shifts the phase of the same carrier in one channel opposite to the direction of shift in a second channel. Any frequency drift of the carrier, however, *is in the same direction in both channels.*

When the f.m. carrier in one of the channels is beat with a heterodyne oscillator, and the beat note in turn heterodyned with the f.m. carrier in the other channel, the initial frequency drift is cancelled out, but the frequency modulation of the two waves is added, resulting in a final carrier having the drift of only the heterodyne oscillator, and having the frequency deviation of 75 kc maximum, as desired.

The action can be better understood with reference to Fig. 1. The audio modulating signal enters at the point marked "PROGRAM LINE INPUT," is converted into a balanced or push-pull signal by the phase inverter, and is then modified by the "AUDIO CORRECTOR" circuit so as to receive preemphasis and a drooping high-frequency response necessary to convert p.m. into f.m.

It is then fed into the balanced modulator directly below. At the same time a crystal-controlled 7C7 oscillator tube and 7C5 buffer amplifier stage feed the carrier frequency (anywhere from 192 to 205 kc, depending upon the final carrier frequency desired) into the balanced modulator. The side-band output is fed via two 7C7 tripler stages into two channels, #1 and #2, such that the instantaneous frequency deviation in the one channel is opposite to that in the other; i.e., if the frequency of Channel #1 rises by 2 kc, that of Channel #2 drops by this same amount. (The balanced modulator will be analyzed in detail farther on.)

Subsequent multiplication of the frequency in each channel takes place in the third and fourth panels from the bottom. A total of four tripler stages gives a multiplication in each channel of $3^4 = 81$ times.

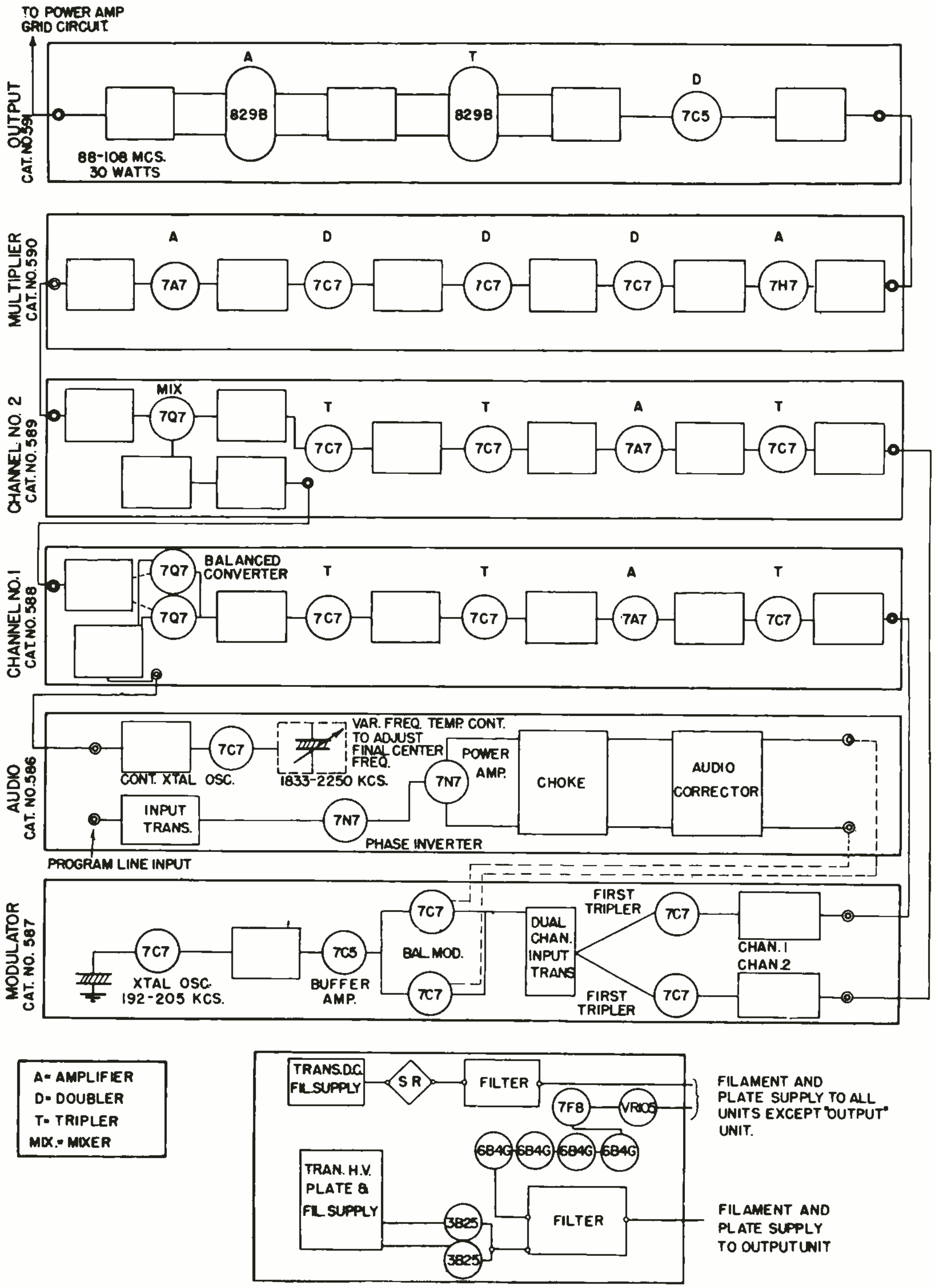
At the same time there is to be noted in the audio panel, another crystal-controlled oscillator using a 7C7 tube, and capable of oscillating at any desired frequency in the range from 1,833 to 2,250 kc. Its output is beat with the signal from Channel #1 in a balanced converter. The output of the latter is in turn beat with the signal from Channel #2 in the mixer stage shown in the third panel from the top. The resulting beat frequency output is then multiplied the required number of times in the second panel from the top to furnish the desired carrier frequency, and is then fed into a series of buffer and driver stages in the top panel. The output from the latter panel contains the desired carrier frequency and maximum deviation, and is used to drive the grids of a power amplifier stage.

Consider now the operation from a numerical standpoint. Suppose the initial crystal oscillator develops a 200-kc wave. This is multiplied 81 times to produce a frequency of 16.2 mc in each channel. Suppose the balanced modulator produced an initial phase shift of +5° (leading) in Channel #1, and simultaneously −5° (lagging) in Channel #2. The multiplication of 81 times produces an output phase shift of 81 × 5 = 405° for Channel #1, and −405° for Channel #2.

Suppose, in addition, the initial crystal oscillator drifts 1 c.p.s. from its nominal 200 kc value. This drift is also multiplied 81 times to produce a frequency drift of 81 c.p.s., *in the same direction* in both channels. One can write, therefore, as follows:

**CHANNEL #1: 16.2 mc + 81 c.p.s. + 405°**

**CHANNEL #2: 16.2 mc + 81 c.p.s. + 405°**




(Courtesy R.E.L.)

Fig. 1.—Block diagram of the dual-channel modulator.

Channel #1 is then heterodyned with the 2,250-kc crystal oscillator to produce a difference beat of (16.2 mc + 81 c.p.s. + 405°) - 2,250 kc = 13,950 kc + 81 c.p.s. + 405°.

This beat frequency is then heterodyned with the signal from Channel #2. The resulting difference beat is

$$(16.2 \text{ mc} + 81 \text{ c.p.s.} - 405°)$$

$$- (13{,}950 \text{ kc} + 81 \text{ c.p.s.} + 405°)$$

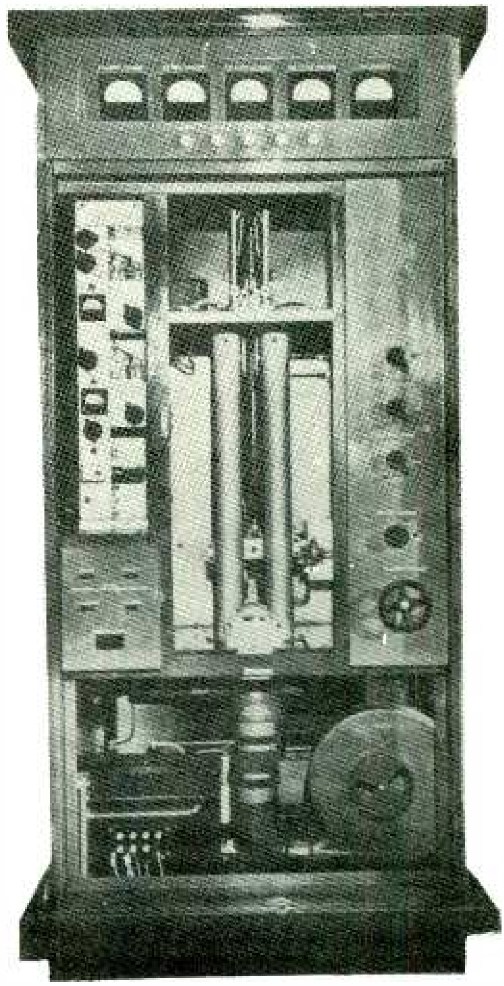
$$= (2{,}250 \text{ kc} + 0 - 2 \times 405°)$$

$$= 2.25 \text{ mc} - 810°$$

Note that the 81 c.p.s. drift of the initial crystal oscillator *has cancelled out*, whereas the 405° phase modulation *has doubled* by this double heterodyning process. The cancelling out of the drift of the initial crystal oscillator means that the only drift that may now occur is that owing to the heterodyne crystal oscillator; i.e., only this oscillator really has to be held to the desired constancy of frequency. The frequency stability is guaranteed to within ±2,000 c.p.s. and therefore meets the F.C.C. requirements,

*MECHANICAL FEATURES.*—The equipment comes in five sizes: 250 watts, 1,000 watts, 3,000 watts, 10,000 watts, and 50,000 watts. In Fig. 2 is shown a photograph of the 1,000-watt transmitter. The doors and removable panels are fitted with switches that form part of the safety interlock system. The cabinets are pressurized, and openings are sealed with special gaskets. The air outlets and inlets have filters to exclude dust, and it is claimed that they will last indefinitely.

Each chassis in the modulator section of the cabinet is mounted on hinges, so that it can be swung



*(Courtesy R.E.L.)*

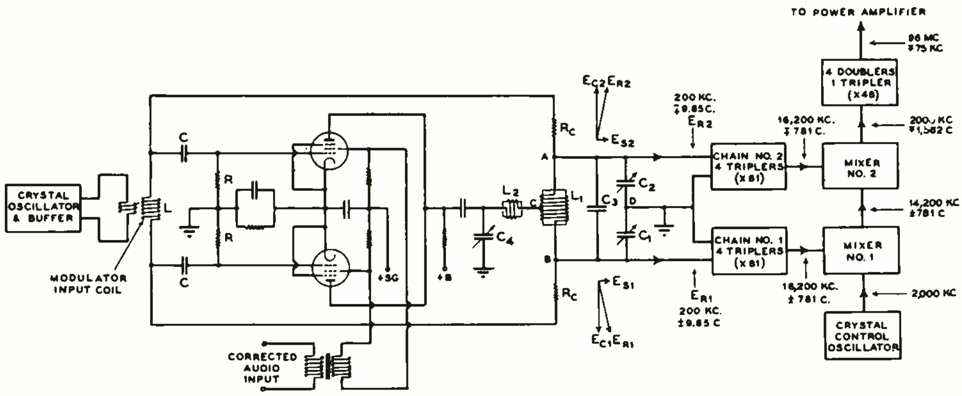
Fig. 2.—Front of the 1-kw transmitter, with panels removed, showing the physical arrangement of the output circuit.

out, thus permitting access to its components. Meters on the front panel furnish readings of the primary voltage and high-power circuits of the transmitter; the principal circuits of the modulator chassis have built-in meters, and tip jacks permit measurements to be made on all other circuits. In addition, an r.f. voltmeter enables the power level on the antenna transmission line to be measured.

*ELECTRICAL FEATURES.*—In discussing the electrical features, it will be of value first to analyze the action of the dual modulator. This, it will be recalled, modulates the two channels in opposite directions. In Fig. 3 is shown a typical modulator stage. The crystal oscillator and buffer stage supply carrier voltage to the modulator stage through C and R, and also carrier voltage to coil $L_1$ via $R_c$ for recombination with the side-band output of the modulator stage, in order to obtain phase (and frequency) modulation.

It will be recalled from an earlier assignment that phase modulation is obtained by combining the carrier with the side bands after these have been shifted through 90° from their normal phase with respect to the carrier. In Fig. 3 such 90° shift is obtained by shifting the carrier voltage applied to the modulator input by 90°. This is accomplished by each C and R. These are relatively small in value; hence the reactance of C is very large compared to R at the carrier frequency.

As a result, the current through R and C leads the carrier voltage induced in L by practically 90°. This current produces a voltage drop in R in phase with itself; hence this voltage also leads that induced in L by 90°. Since the voltage across each R is the input to each modulator tube, it is clear that the carrier input to the modulator stage leads the voltage recombined with it at points AB by 90°.




(*Courtesy R.E.L.*)

**Fig. 3.—Simplified diagram of Armstrong dual channel modulator.**

The action of the balanced modulator is exactly the same as that described in the assignment on Modulation in Section II of the course. The only difference is that here the plates, rather than the grids, are connected in parallel, and the control grids are connected in push-pull for the r.f.

carrier. A further point is that the screen grids are connected in push-pull for the audio modulating voltage.

The operation is as follows: In the absence of audio modulation, the opposite polarity of grid voltage for each tube, causes one plate current to increase when the other decreases; under balanced conditions these variations cancel in the common plate load, and no r.f. output is obtained.

When modulation is applied to the screen grids, the tube whose screen is driven in a positive direction experiences an increase in its $G_m$, and its r.f. output increases; the tube whose screen is driven in a negative direction experiences a decrease in its $G_m$, and its r.f. output decreases. Hence, the first tube overbalances the second tube, and a net polarity of r.f. output is obtained.

On the other half of the audio cycle, conditions are reversed, and the polarity of the r.f. output is reversed. The output wave therefore appears as in Fig. 4.

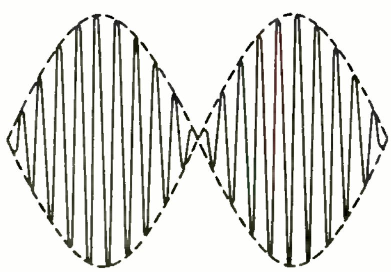


Fig. 4.—**Appearance of output wave of the balanced modulator.**

As has been explained in a previous assignment, such a wave contains only the side bands, and not the carrier, of the amplitude-modulated wave. This can be recognized by the fact that the upper and lower envelopes cross one another. If sufficient carrier voltage were present, the two envelopes would be separated sufficiently so that instead of crossing, they would at most be tangent to one another at the points of maximum inward modulation.

Since the output of the balanced modulator contains only the side bands, and moreover these have been given the necessary 90° phase shift by the input RC circuit, they are now ready to be combined with the carrier to furnish phase modulation. It is necessary that they be combined in 90° phase for one channel, and in (90° + 180°) = 270° phase to produce the opposite phase modulation in the other channel. The circuit that accomplishes this involves $L_2$, $L_1$, $C_3$, $C_2$, and $C_1$.

Consider first coil $L_1$. It has a center tap C to which $L_2$ is

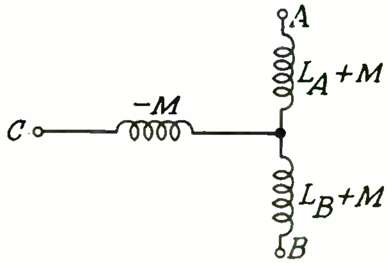



Fig. 5.—**Representation of a center-tapped coil by three inductances.**

connected. As has been explained in a previous assignment, such a coil having mutual inductance M between its two half-windings can be represented by three inductances in a Tee configuration, as shown in Fig. 5. The *self-inductances* of the two half windings are $L_A$ and $L_B$. Inspection of Fig. 5 proves this: the inductance between terminals A and C is $(L_A + M) - M = L_A$; the inductance between C and B is $(L_B + M) - M = L_B$. The inductance between A and B, however, is $(L_A + M) + (L_B + M) = L_A + L_B + 2M$, which is correct: the total inductance involves not only the self-inductances of each half winding, but also twice the mutual inductance between them.

The three capacitors $C_1$, $C_2$, and $C_3$ act essentially like two capacitors in series: one ($C_A$) connected between A and ground, and the other ($C_B$) between ground and B. The reason for using $C_3$ in the actual circuit is that $C_1$ and $C_2$ can be thereby made smaller and less bulky variable capacitors: in action the three behave (as just stated) as if there were only two larger capacitors $C_B$ and $C_A$ connected as $C_2$ and $C_1$.

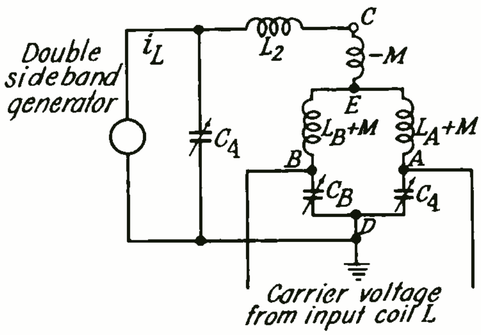



**Fig. 6.—Equivalent circuit showing elements of the output load of the balanced modulator of Fig. 3.**

Hence finally, the various circuit components can be drawn as in Fig. 6. The two parallel branches involving $(L_B + M) + C_B$, and $(L_A + M) + C_A$, are each made series resonant to the carrier frequency and therefore resistive in nature, so that the net impedance is a resistance. The negative inductance -M is more than balanced by the positive inductance of coil $L_2$, hence the total impedance paralleling $C_4$ is a net inductive reactance and resistance in series; i.e., $L_2$, $-M$, $(L_B + M)$, $C_B$, $(L_A + M)$, and $C_A$ act like an L-R series circuit.

The net inductance tunes with $C_4$ to parallel resonance. As a result, the double side band generator (balanced modulator tubes) feed a resistive load, and the line current $i_L$ is in phase with the impressed voltage. But the branch current in $C_4$ leads the line current by 90°, and the branch current in $L_2$, etc., lags $i_L$ by 90°. At point E this current divides equally between the two parallel paths, and produces equal voltages across $C_B$ and $C_A$, in phase with one another, and both lagging the current by 90°. Hence, these two voltages lag the line current $i_L$ by 90° + 90° = 180°, and thus also lag the impressed voltage of the side band generator by 180°.

Since a 90° phase shift was initially introduced into the input voltage of the balanced modulator by R and C of Fig. 3, the output side-band voltage across $C_B$ and $C_A$ lag the input carrier voltage by 180° + 90° = 270°. The same phase angle exists with respect to the carrier voltage applied to points A and B of Figs. 3 and 6 from input coil L.

However, the carrier voltage is applied in push-pull or balanced-to-ground fashion. This means that the carrier voltage from ground to terminal B is 180° out of phase with that from ground to terminal A. Hence, the two carrier voltages combine with the side-band voltages as indicated in Fig. 3, and repeated in Fig. 7. In Fig. 7, the carrier voltage applied between terminal A and ground is denoted by $E_{c2}$; that between terminal B and ground is denoted by $E_{c1}$. It will be observed that they are 180° out of phase, *as measured from ground to either terminal.*

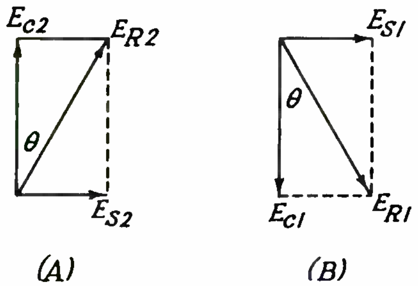



**Fig. 7.—Vector diagram for carrier and side-band voltages in balanced modulator.**

The two side-band voltages $E_{S2}$ and $E_{S1}$ are in phase. Voltage $E_{S1}$ is developed across $C_B$, and voltage $E_{S2}$ is developed across $C_4$, in Fig. 6. Therefore, the respective resultants, $E_{R1}$ and $E_{R2}$ represent phase deviations of $\theta$ degrees in opposite directions: $E_{R1}$ is a phase advance, and corresponds to a rise in frequency, whereas $E_{R2}$ is a phase delay and corresponds to a decrease in frequency. These two voltages are fed to the two channels, and give rise to the equal and opposite frequency deviations in the channels.

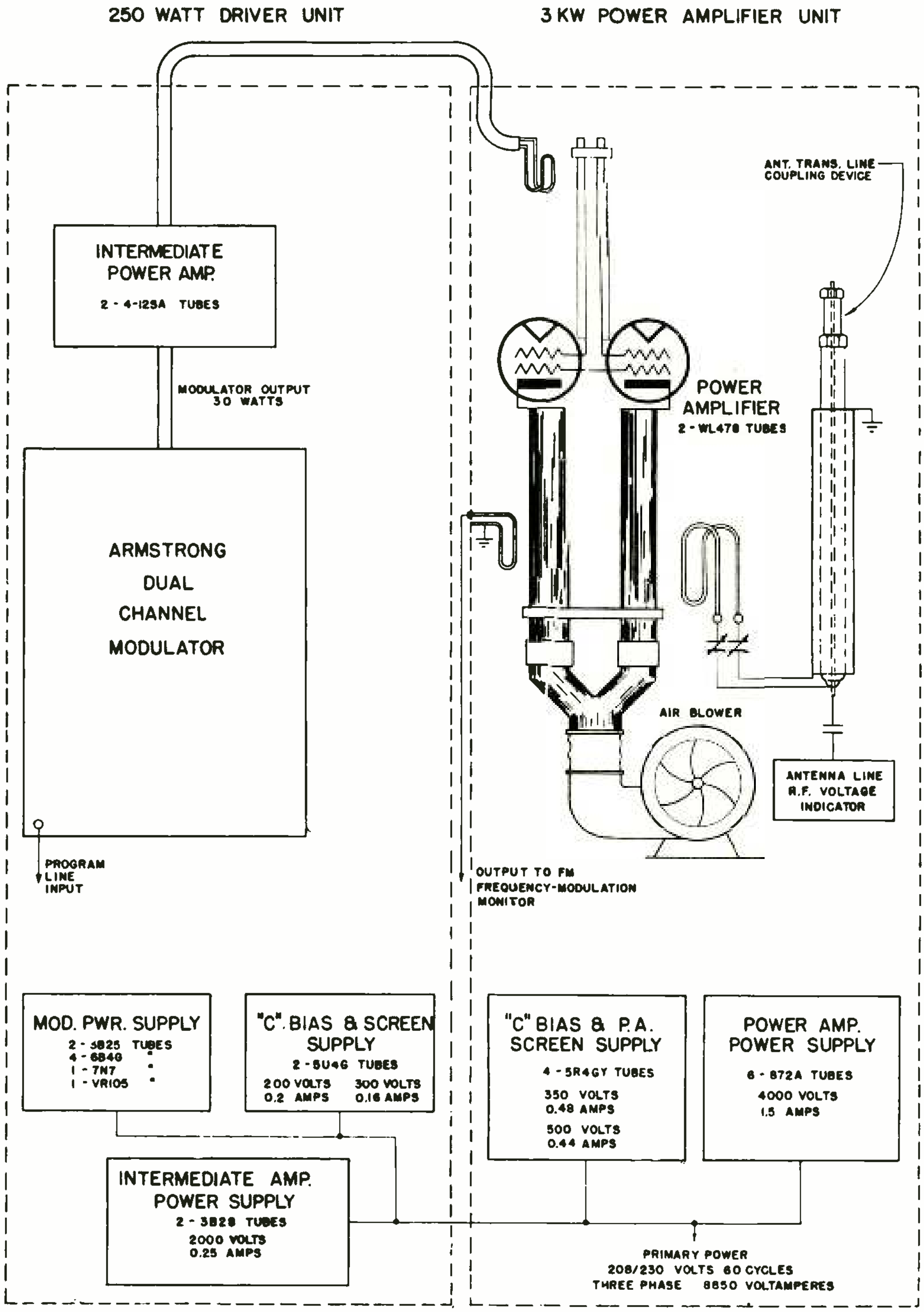
The output of the dual-channel modulator is 30 watts. This feeds an intermediate power amplifier consisting of two 4-125A tubes in push-pull, whose output is 250 watts. These units are mounted in a vertical metal cabinet, and constitute the 250-watt transmitter.

In the case of the 3-kw transmitter, another similar vertical cabinet is mounted alongside of the 250-watt cabinet, and contains the power amplifier stage and associated plate, screen and bias supplies. The arrangement is shown in block diagram form in Fig. 8.

It will be observed that the intermediate 250-watt power amplifier feeds the 3-kw power amplifier in the adjacent cubicle through a coaxial line. This is loop coupled to the two-wire grid transmission line of the power amplifier. The latter consists of a pair of WL478 tubes in push-pull. The shorted quarter-wave grid line furnishes the excitation signal, the shorted quarter-wave plate line furnishes the output power.

Note that the bottom of the plate line is at r.f. ground, although at a high d.c. potential to ground. Hence, an ordinary insulating section of the ventilating pipe is required between the blower and the plate line; a low-loss section is not necessary. However, the insulating section must also be a good acoustic insulator in order to prevent the transmission of mechanical vibrations from the blower and motor to the power tube elements. The construction is such that the tubes are inserted into the plate lines in order that adequate cooling, particularly of the plates, be obtained.

The antenna line is coupled to the plate line by means of a tuned coupling loop. Since the




(Courtesy R.E.L.)

Fig. 8.—Simplified diagram of the 3-kw f.m. transmitter.

plate line is of the two-wire form to operate balanced-to-ground in view of the push-pull arrangement of the tubes, and since the antenna line is of the unbalanced-to-ground or coaxial form, a special "bazooka" or quarter-wave sleeve is required.

This sleeve can assume at its bottom end a certain r.f. potential to ground and the coaxial sheath it surrounds, the inner conductor can assume an equal and opposite r.f. potential to ground and the coaxial sheath, and yet one-quarter wave-length above the bottom end, the sleeve can be conductively connected to the coaxial sheath and to ground. The theory of operation has been discussed in a previous assignment in the u.h.f. section of the course. In passing it is to be noted that an output power from 1,000 to 3,000 watts, can be obtained in the $1^5/_8$ inch diameter antenna line when the latter is terminated in its characteristic impedance. A further point is that the plate efficiency has a value in excess of 70 per cent, which is excellent at the ultra-high frequency operation.

At the left-hand side of the plate line will also be observed another coupling loop with one side grounded. This is for the purpose of furnishing output for the f.m. monitor. The output power is 2 watts r.f.

Some further points are that the audio input level for 100 per cent modulation is only +12 ±2 db; the input impedance is 600 ohms balanced to ground. The overall response is within ±1 db from 50 to 15,000 c.p.s. at 25%, 50%, and 100% modulation. The measured r.m.s. harmonic distortion is less then $1^1/_2$% for all signal frequencies between 50 and 15,000 c.p.s. at 25%, 50%, and 100% modulation, and the f.m. signal-to-noise ratio is at least 70 db below 100% modulation as measured at the output of a monitor receiver within a band of 50 to 15,000 c.p.s. The a.m. signal-to-noise ratio is at least 60 db below 100 per cent modulation measured over the same band.

The maximum frequency swing possible is ±100 kc, or is in excess of the ±75 kc specified by the F.C.C. indeed, the transmitter exceeds in practically every respect the requirements as laid down by the F.C.C.

## RCA F.M. TRANSMITTERS

*BASIC FEATURES.*—This transmitter operates on essentially the same basic principles as the Western Electric transmitter: the use of a balanced reactance tube stage to frequency-modulate an oscillator tube, with subsequent multiplication to obtain the desired carrier frequency and frequency deviation, together with a demultiplying system to obtain a sub-carrier frequency, which is then compared with a crystal-controlled oscillator, and the beat frequency employed to operate a synchronous motor to retune the initial tank circuit.

Aside from constructional details, the main difference between the two is the use of the different types of frequency-divider circuit, This will be discussed farther on. A simplified schematic of the transmitter for the Type BTF-3B 3-kw transmitter is shown in Fig. 9. The 6V6 modulated oscillator is essentially a Hartley oscillator, with B+ (r.f. ground) tapped up on the tank coil; the grid connected through a grid leak and condenser

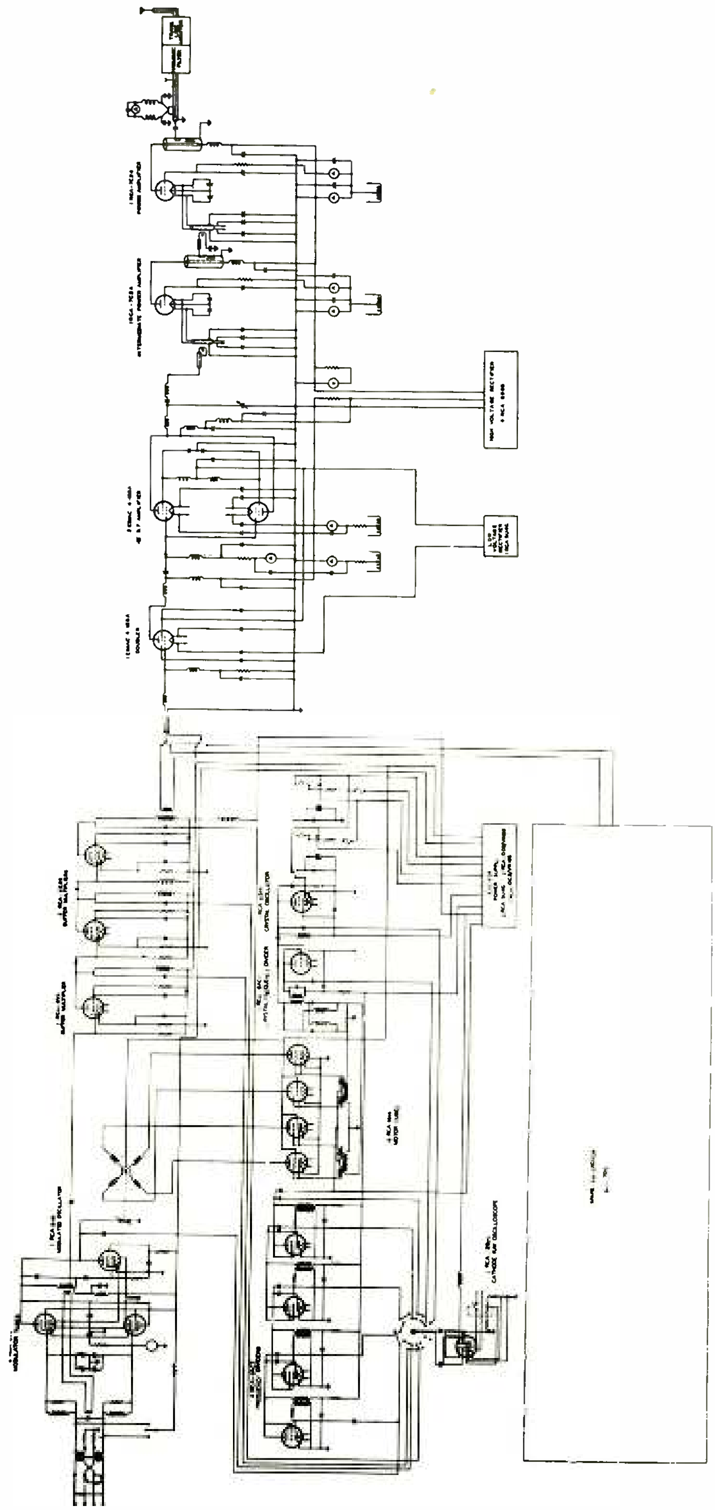


Fig. 9.—Simplified schematic diagram of RCA Type BTF-3B 3-kw transmitter. (Courtesy RCA.)

to the bottom end of the coil; and the plate connected to the top end.

A small coil directly to the left of the tank coil picks up some of the signal and transfers it via a link circuit to the two coils in the grid circuits of the 6V6 modulator tubes. The latter two coils are tuned by the two variable capacitors, so that the voltage induced in each sets up a grid excitation voltage 90° out of phase with that across the oscillator tank circuit.

Since the plates of the modulator tubes are connected in parallel to the oscillator tank circuit, their currents flow through this circuit. Since the currents are essentially in phase with the excitation voltages on the grids, they are 90° out of phase with the tank voltage. Thus, the modulator tubes act as reactances across the tank circuit.

However, the voltages induced in the two modulator grid coils by the link circuit are 180° out of phase. This means that if the plate current of one modulator tube leads the tank voltage by 90°, the plate current of the other modulator tube *lags* the tank voltage by 90°. Since the two plates are connected in parallel to the tank circuit, their reactive currents are equal and opposite and hence cancel in the absence of modulating signal.

When a modulating signal is applied through the two audio transformers in the modulator grid circuits, the grids are modulated by this signal also 180° out-of-phase. This means that as one grid swings more negative the other swings more positive, so that the transconductance of one decreases as that of the other increases. Hence one reactive plate current decreases while the other reactive plate current increases, and the two no longer cancel.

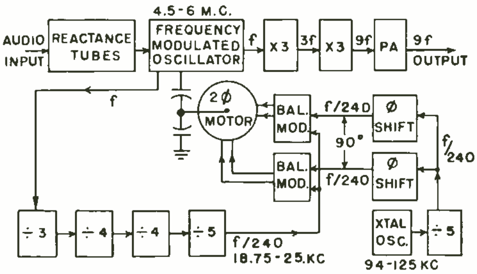
As a result, during one-half of the audio cycle, a net leading current flows through the tank circuit from the two modulator tubes, and during the next half of the audio cycle a net lagging current flows through the tank circuit. Therefore, during the first half-cycle the oscillator frequency decreases, because it is as if the tank capacity had increased, and during the next half-cycle the frequency increases, because it is as if the tank inductance had decreased. In this way frequency modulation is obtained.

The use of a balanced modulator circuit cancels, at least in part, the nonlinearity in variation of the transconductance of each tube with audio signal. As a result, a frequency deviation as high as ±10 kc on an oscillator frequency of 5 mc is possible, so that the frequency multiplication required for ±75 kc deviation is relatively small. In the actual equipment a multiplication of nine times is employed before driving a power amplifier. The output of the latter is in the 40.5 to 54 mc range.

This output is fed through a transmission line to the main transmitter, where it is once more multiplied in frequency by a factor of two, resulting in a final output frequency in the 81 to 108 mc range. The circuit in block diagram form is shown in Fig. 10, which should be consulted in conjunction with Fig. 9.

The doubler stage feeds the first r.f. amplifier, consisting of two Eimac 4-125A tubes in parallel.

Note that the screens are series tuned to ground by adjustable tuning capacitors, and isolated from one another and the power supply by individual r.f. chokes and resistors. The first r.f. amplifier feeds a coaxial line through an impedance-matching tee network.




*(Courtesy RCA Review)*
**Fig. 10.—Block diagram of RCA f.m. exciter unit.**

The coaxial line is then loop-coupled to another transmission line in the cathode circuit of the intermediate power amplifier. This is of the grounded-grid type, hence the signal is injected in the cathode circuit as just mentioned. Since the latter is of the filament type, means must be provided to keep the signal out of the 60-cycle a.c. secondary supply, and also to prevent the capacity of the latter to ground from shorting out the r.f. signal.

This is accomplished by surrounding the two filament leads with a cylinder as shown in Fig. 9. The cylinder is suitably grounded, and the portion of the cylinder above the ground is one-quarter wave in length. This makes the top end of the two filament leads have a high impedance with respect to the grounded cylinder owing to the properties of a $\lambda/4$ line.

Hence the top end of the filament leads and the filament itself has a high impedance to ground so far as the r.f. signal is concerned, and the filament can therefore support the high r.f. voltage to ground induced in it by the coupling loop. At the same time the bottom ends of the filament leads are at ground potential, so that no r.f. signal leaks into the filament supply secondary.

The output load of the intermediate power amplifier is a quarter-wave resonant coaxial line (r.f.) grounded at the bottom end. This is the tank circuit for the tube. Another coaxial line taps off this tank circuit at an appropriate impedance point on the inside of the outer conductor, and feeds a coupling loop as shown. This loop induces an r.f. signal voltage in the filament of the power amplifier stage in exactly the same manner as described previously.

The output stage also has a quarter-wave coaxial tank circuit, which is suitably coupled to the antenna line in the same way as that employed for the intermediate power amplifier. Note that this method of coupling from the inside of the outer sheath keeps the d.c. potential away from the loop coupling and antenna line circuits.

The antenna meter is coupled to the interior of the outer conductor of the antenna coaxial line at two points. The resistance unit of the thermogalvanometer circuit then actuates the thermocouple, which in turn energizes the d.c. meter movement.

A harmonic filter of the fixed tuned type reduces all harmonics by 30 db or better, and is provided as standard equipment. In addition, a transmission line monitor guards

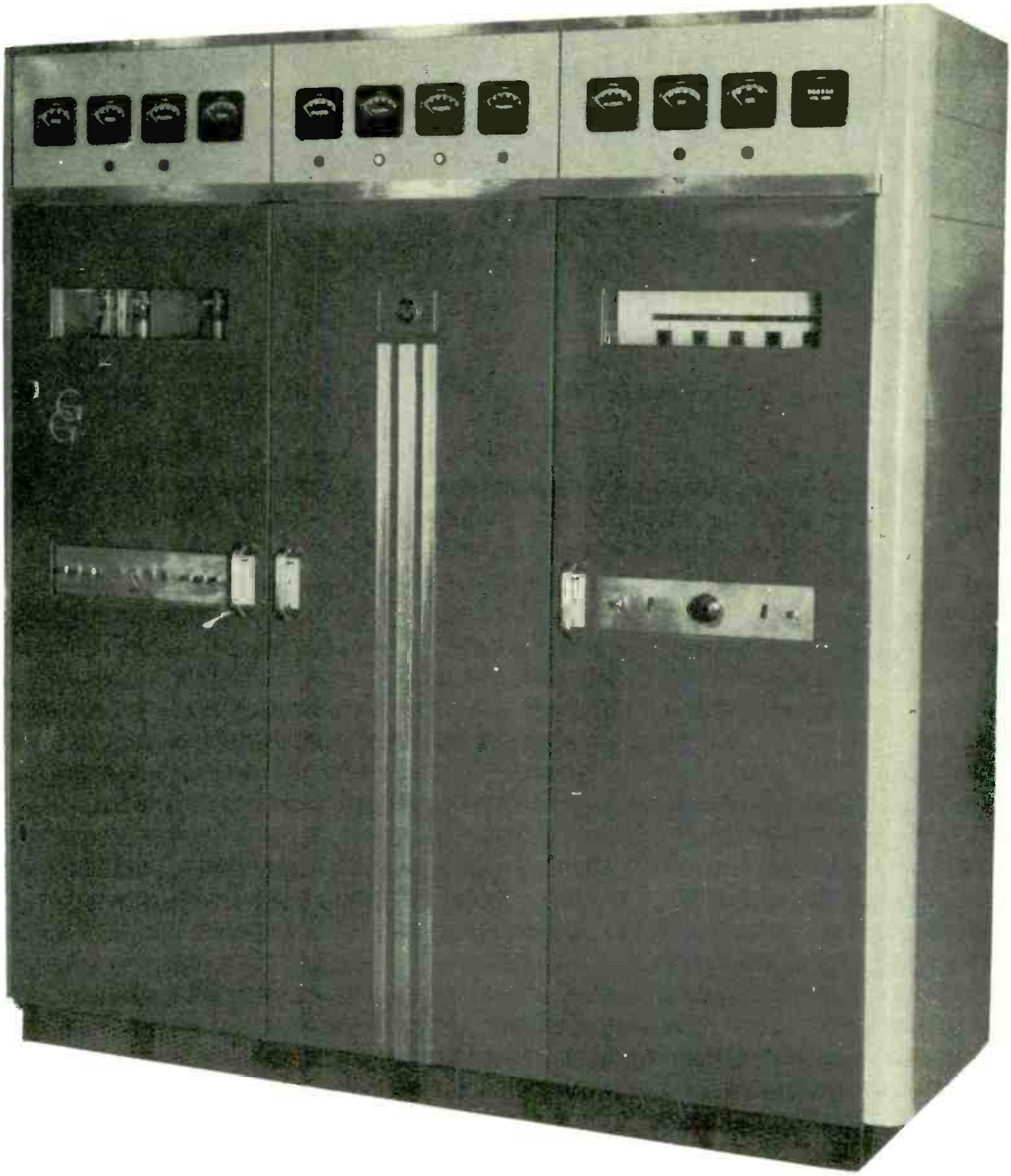


Fig. 11.—Front view of RCA Type BTF-3B 3-kw f.m. broadcast transmitter.

against any change in the signal intensity such as might be produced by an arc in the transmission line or a fault in the antenna itself; any trouble actuating the monitor causes the transmitter to be shut down.

In Fig. 11, is shown a photograph of the transmitter. It consists of three fabricated steel frames bolted to a base frame. Each compartment has a front and rear door; these doors are provided with windows for observing the transmitter while in operation.

A filtered air supply is supplied through individual air inlet openings and removable filters (in the base frame) for each compartment. The warm air is expelled through the roof, and dust collectors collect any dirt which may settle while the transmitter is idle.

Reading from left to right, the compartments house the r.f., f.m. exciter, and control and main rectifier units. The exciter compartment has room for two units (including their power supplies), so that either unit may be selected by the flick of a switch if the other becomes defective. Normally, however, only one is furnished.

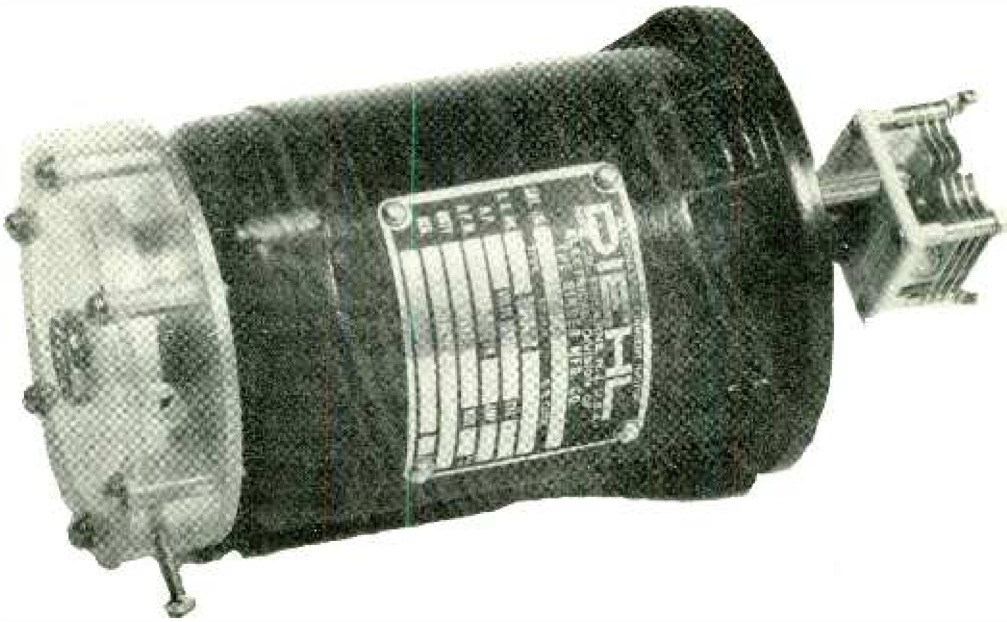
Safety features for the operating personnel include automatic interlock switches for all doors giving access to high-voltage compartments, and grounding bars, mechanically operated, which are automatically released when the doors are opened. Thus, opening a door immediately de-energized all rectifiers and grounds the circuits so that no dangerous charge can remain in any capacitor.

Both manual and automatic sequence starting are provided. In the automatic position, a 3 shot recycling sequence is provided by the control "brain center." This automatically returns the transmitter to the air up to 3 times in case of repeated overloads. If the overload condition still persists, the transmitter is automatically shut down. The carrier monitor mentioned previously is also connected into this circuit so as to cause the transmitter momentarily to shut down in the event of an arc on the transmission line, thereby permitting the arc to clear.

In the event of momentary power-line failure, a special hold-in circuit permits the transmitter instantly to return to the air before a 30-second time delay relay in the plate circuit can function, thereby avoiding this 30-second delay. Means are also provided to operate at reduced power for tuning-up and emergency operation through the use of a power reduction switch in the primary circuit of the main rectifier.

*FREQUENCY CONTROL.*—As indicated previously, the frequency control is similar in principle to that employed in the Western Electric transmitter. The frequency of the modulated oscillator is divided in steps of 3, 4, 4, and 5 down to about 20 kc. A crystal oscillator in the range of 94-125 kc is divided by 5. The two subdivided frequencies are beat with one another in a balanced modulator (refer to Figs. 8 and 10) to provide a difference beat frequency two-phase supply. This operates a motor on whose shaft is mounted the rotor of an air capacitor having a split stator; so that no additional bearings are required in the capacitor (see Fig. 12).

The capacitor is connected in the tank circuit of the modulated oscillator, and retunes it if it drifts relative to the crystal oscillator. Note the oil dash-pot pair of balanced modulator stages has been redrawn. Observe that the output coil L of the frequency divider for the crystal-oscillato feeds a pair of coils and capacitors



(Courtesy RCA Review)

Fig. 12.—Frequency control induction motor, showing capacitor at one end and oil dash pot at other end.

arrangement on the opposite end of the motor shaft. This dash pot offers no friction to continuous rotation in either direction because it rotates with the shaft as a whole.

If, however, the motor overshoots the mark, and begins to rotate in the opposite direction, the inertia of the oil-coupled flywheel tends to prevent the latter from immediately following the reverse rotation, and thus momentarily puts a drag on the motor until it is up to the motor speed. In this way hunting of the control is prevented.

In Fig. 13 the circuit of the as shown redrawn in the right-hand figure. The phase-shift in each branch is such that the voltage across $R_1$ is 90° out of phase with respect to that in $R_2$.

As a result the voltage across $L_2$ is 90° out of phase with respect to that of $R_1$. On the other hand, the output of the subdivided modulated oscillator frequency is fed to the center taps of the associated secondary coils in phase. The two resultant beat frequencies produced by each pair of modulator tubes are 90° out of phase with one another, and since these two beat frequencies are fed to the two phase windings of the motor, a rotating field will be produced of a velocity depending upon the value of the beat frequency.

As explained in the assignment on modulation, the velocity of the rotating field is independent of the magnitude of the motor currents, although the latter determine the motor torque and hence the sensitivity of the frequency control. Moreover, if the carrier frequency of the modulated oscillator varies from above to below that of the crystal oscillator, the phase of one of the motor currents reverses, whereas that of the other remains unchanged, whereupon the motor reverses its direction of rotation. In this way both positive and negative drifts can be corrected.'

The motor can respond to beat frequencies from as low as 0.02 c.p.s. or 1.2 cycles per minute to as high as 1,000 c.p.s. This is a result of the low frictional drag on the motor. The low value of 0.02 c.p.s. corresponds to a final frequency drift of

$$.02 \times 5 \times 4 \times 4 \times 3 \times 3 \times 3 \times 2 = 86.4 \text{ cycles}$$

from that of the crystal oscillator. Since at the final carrier frequency a total drift of 2,000 c.p.s. is permitted, this leaves 2,000 - 86.4 = 1913.6 c.p.s. for the crystal oscillator to control. 1913.6 c.p.s. at 100 mc, for example, corresponds to 19.13 parts in a million, which is a constancy easily attained by a crystal oscillator.

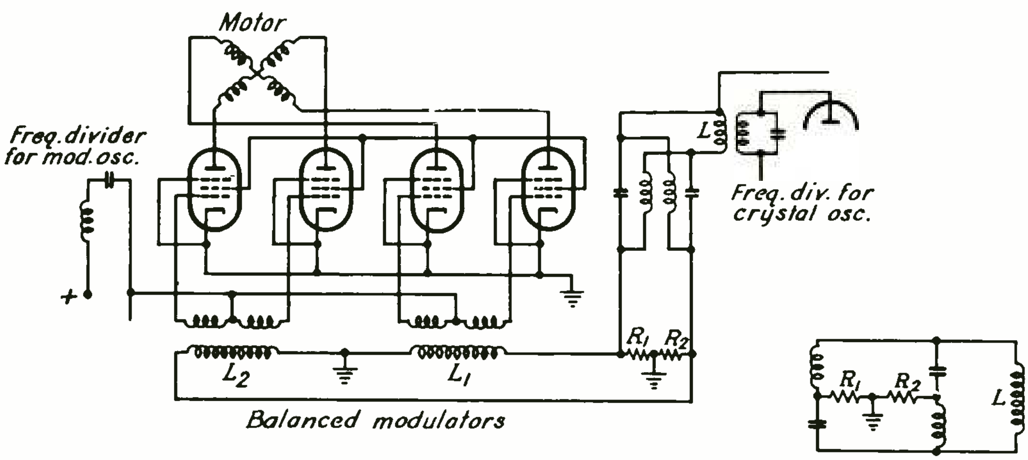



Fig. 13.—Portion of Fig. 9 showing detail of two balanced modulator stages employed for motor control.

The maximum value of 1,000 c.p.s. beat frequency means that if for any reason a large drift should occur—say—when the equipment is first turned on, the motor will be able to respond and reduce it to within the value specified by the F.C.C. It is stated that the control is so sensitive that the slight variation in crystal frequency produced by changes in its thermostat-controlled oven temperature, are followed by the control system. These variations correspond to only ±10 cycles at 100 mc.

*ELECTRICAL CHARACTERISTICS.*—An

important feature to note is the use of grounded grid circuits. These have been discussed fully in the assignments on U.H.F. Techniques. Suffice it to say here that one grounded grid circuit is not only particularly well suited for u.h.f. operation in that triode tubes can be employed and yet neutralization avoided owing to the shielding action of the grounded grid between the cathode and plate, but also the appreciable input power required to drive the cathode is not wasted, and appears in large part as useful power in the plate circuit. This is particularly important in the higher-powered transmitters, since large power output tubes in the u.h.f. range are difficult to build.

Some other electrical characteristics of interest are the following:

| | | |
|---|---|---|
| 1. | R.F. output impedance | 35 to 75 ohms |
| 2. | Modulation capability | ±100 kc |
| 3. | Audio input impedance | 600 ohms |
| 4. | Average program level | +4 ±2 vu |
| 5. | 100% modulation level | +12 ±2 dbm |
| 6. | Audio-frequency response, 1,000-cycle reference, uniform within | ±1 db |
| 7. | Audio-frequency distortion | |
| | 30 - 100 c.p.s. | 1.5% |
| | 100 - 7,500 c.p.s. | 1.0% |
| | 7,500 - 15,000 c.p.s. | 1.5% |
| | including all harmonics up to 30 kc/s at 75-kc swing | |
| 8. | F.M. noise level, below ±75-kc swing | 65 db |
| 9. | A.M. noise level, below 100% amplitube modulation (this refers to residual amplitude modulation of the f.m. carrier). | 50 db |

Three-phase power is required at a frequency between 50 and 60 cycles, and at a voltage between 208 and 230 volts. The approximate power consumption is 8,300 watts, at a power factor of 85%. Another important point to note is the use of the same kind of tube in as many different stages as possible, thus reducing the total number of tube types required to be kept in stock.

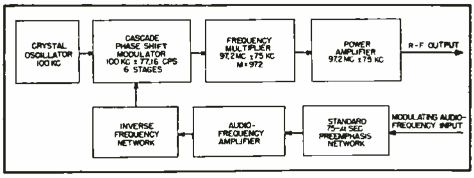
## THE RAYTHEON F.M. TRANSMITTER

*BASIC THEORY.*—The Raytheon f.m. transmitter operates on the phase-modulation rather than direct frequency-modulation principle. It obtains a large initial phase shift by a series of cascaded phase shifters. Thus, the first phase-shift stage may produce a phase shift of 23.9°; the output is then fed into a second stage, which produces an additional phase shift of 23.9°, or a total of 47.8°, and so on through six stages, which produce a total phase shift of $6 \times 23.9° = 143.4°$. For a 30-cycle note this corresponds to a frequency deviation of

$$\frac{143.4}{57.3} \times 30 = 75 \text{ c.p.s.}$$

The resulting multiplication in order to obtain a deviation of 75 kc for a 30-cycle note is 75,000/75 = 1,000. Such a value of multiplication of an initial 100-kc carrier frequency results in a carrier frequency of 100 mc, which is in the f.m. range, *and does not require a heterodyne oscillator.* The elimination of a heterodyne oscillator obviates spurious beat frequencies that can appear in its output unless tuned circuits and adequate shielding are employed.

A block diagram of the system is shown in Fig. 14. The 100-kc crystal




*(Courtesy Electronics)*

**Fig. 14.—Block diagram of Raytheon transmitter employing cascade phase-shift modulator.**

oscillator feeds the 6-stage cascaded phase-shift modulator, wherein a total phase of 147.5°, or 24.6° per stage, is obtained. (This is slightly more than the value of 23.9° employed in the example cited above, in order that the final carrier frequency come out to be 97.2 mc.)

A frequency multiplier of 972 times then brings this frequency up to 97.2 mc, and the frequency deviation up to $\pm 77.16 \times 972 = \pm 75$ kc. Suitable power amplifier stages then bring the power level up to the desired value. The ratings for the transmitters are 250 watts, 1 kw, 3 kw, and 10 kw.

*PHASE SHIFT MODULATOR.*—The phase-shift modulator employs a pentode tube in each stage as a constant-current source, which feeds a special plate load. This load includes a vacuum tube that functions as a variable r.f. resistor, the variations being controlled by the modulating voltage applied to its grid. As a result, the phase of the voltage output varies, but not its amplitude, at the modulating frequency, and thus phase modulation is produced.

The basic circuit which accomplishes this is shown in Fig. 15. The plate load consists of L, R, and C, as shown, with R, actually a vacuum tube (to be discussed subsequently). The driver tube is the

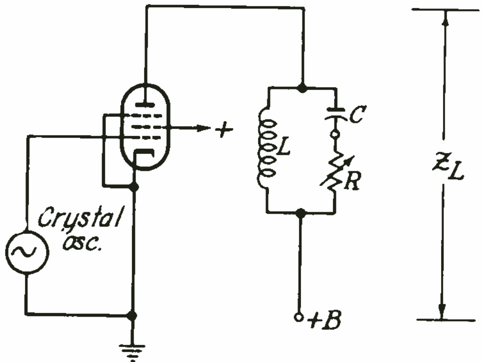



**Fig. 15.—Basic phase-shift modulator circuit.**

initial crystal-controlled oscillator, or an amplifier tube following the oscillator and acting as a constant-current source.

The absolute value of the impedance $|Z|$ of the load is given by

$$|Z| = (2\pi fL)\sqrt{\frac{1+\left(\frac{1}{b}\right)^2}{1+\left(\frac{1}{b}\right)^2\left(\frac{1}{a}-1\right)^2}} \qquad (1)$$

where $a = 1/(2\pi f)^2LC$, and $b = 2\pi fCR$. An examination of Eq. (1) shows that the numerator would be equal to the denominator if $(1/a - 1)^2$ were equal to unity, and Z would equal simply $2\pi fL$ regardless of the value of b; i.e., independent of $2\pi fCR$. Thus

$$\left(\frac{1}{a}-1\right)^2 = 1$$

or

$$\frac{1}{a} - 1 = 1$$

$$\frac{1}{a} = 2$$

$$a = 1/2 \qquad (2)$$

If this value is employed, then

$$a = 1(2\pi f)^2LC = 1/2$$

$$= 1/(2\pi fL)(2\pi fC)$$

or

$$1/2 = X_C/X_L$$

from which

$$X_L = 2X_C$$

or

$$2\pi fL = 2/2\pi fC \qquad (3)$$

In other words, at the carrier frequency of the crystal oscillator the inductive reactance must be twice the capacitive reactance. Then, as R is varied, the impedance Z remains constant in amplitude, *but its phase angle changes.*

The significance of this is that the output voltage of this stage is given by

$$e_o = e_i G_m Z \qquad (4)$$

where $e_i$ is the input voltage to the grid of the constant-current generator. (This has been developed in a previous assignment.) In Eq. (4), if the magnitude of Z is constant, then $e_o$ is constant in magnitude for constant magnitude of the input voltage $e_i$.

However, the phase angle of Z is also the phase angle between $e_i$ and $e_o$. As this angle changes (by varying R), the phase angle between $e_o$ and $e_i$ varies accordingly. Hence, if $e_i$ is the r.f. excitation voltage applied to this unit, and R is varied at the modulation frequency, then $e_o$ will be constant in amplitude, *but will vary in phase relative to* $e_i$ and will therefore acquire phase modulation with its attendant frequency deviation. If the audio control voltage for R has a response that varies inversely with the audio frequency, then true frequency modulation will result.

The method of obtaining a resistance R that varies at the audio frequency is an electronic one, and is illustrated in Fig. 16. Coil L and C are the same components as

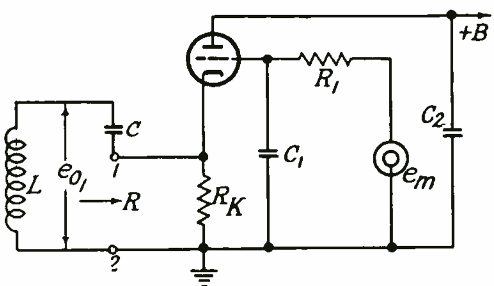



**Fig. 16.—Phase-shift stage showing use of vacuum tube as a variable r.f. resistor R.**

shown in Fig. 15. Resistor R, however, is the r.f. resistance looking into the cathode circuit of the vacuum tube.

The impedance looking into terminals 1-2 is the same as looking into a grounded-grid amplifier, since $C_1$ essentially shorts the grid to ground so far as r.f. is concerned. This impedance has been derived in the assignment on grounded-grid amplifiers; and (for Class A operation and disregarding for the moment the shunt resistor $R_k$) is

$$R = (Z_L + R_p)/(1 + \mu) \qquad (5)$$

where $Z_L$ is the plate load impedance. But $Z_L = 0$ in Fig. 12 because $C_2$ effectively shorts the plate to ground for r.f. currents. Hence

$$R = R_p/(1 + \mu) \qquad (6)$$

In the circuit shown, the r.f. signal voltage, however, is sufficiently strong to drive the cathode beyond cutoff for some portion of the cycle. Hence the above value of $R_p/(1 + \mu)$ varies very markedly during the r.f. cycle, with the result that the plate current flows in a series of pulses instead of as a true sine wave (Class B or C operation instead of Class A).

One can still find the fundamental component of these pulses, and divide its amplitude into the input voltage between terminals 1-2, thereby obtaining the resistance of the tube to the fundamental component.

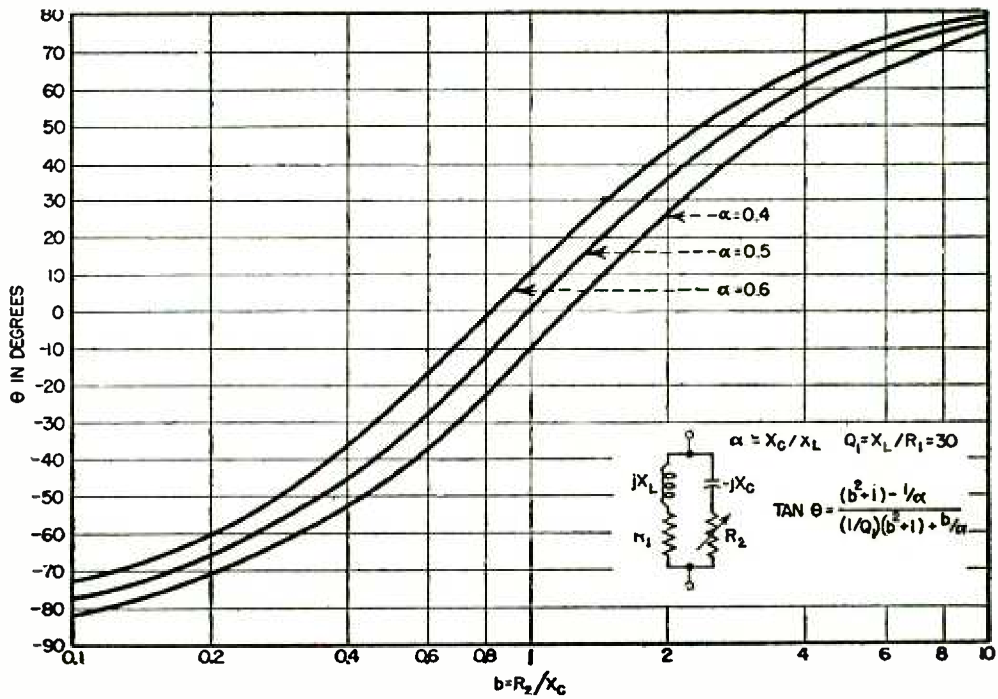
This value constitutes the magnitude R that is presented to the input circuit of L and C. It will be considerably higher than the value for Class A operation, the more so as the grid is biased more and more negative, and the current flows in smaller and smaller pulses of narrower and narrower widths.

Therefore, if a modulating (audio) voltage $e_m$ is applied to the grid, it will vary the bias at a low-frequency audio rate, and thus vary the r.f. resistance appearing between terminals 1-2. Note that if the instantaneous grid bias at the negative peak of the audio wave were such that the superimposed r.f. signal between terminals 1-2 just drove the cathode up to cutoff, then essentially no plate current would flow at all during the r.f. cycle, and the apparent resistance between terminals 1-2, due to the tube, would be infinite. Owing to the presence of $R_k$, however, the actual resistance between 1-2 would be in this case $R_k$.

At the positive peak of the grid audio swing, the bias is a minimum, current flows over most of the r.f. cycle, and the cathode input resistance drops to a relatively low value, thereby markedly shunting $R_k$, and thus reducing the net resistance between terminals 1-2 to a value much below $R_k$.

In Fig. 17 are shown the phase characteristics for three values of $a = 1/(2\pi f)^2LC$; namely, $a = 0.4$, $a = 0.5$ (the theoretical value), and $a = 0.6$. It will be observed that the phase-shift angle shows an S-shaped characteristic versus $\omega RC$. Hence R must vary in inverse manner with the audio grid voltage $e_m$ to make the phase-shift angle vary linearly with $e_m$. The circuit of Fig. 16 produces just the right variation in R with $e_m$ to provide the desired linear relation between the phase-shift angle and $e_m$ over the range of values of about ±25°. This is the required range to provide 75 kc. deviation in the final carrier wave.

When the audio signal $e_m$ varies the grid bias through a range of values,

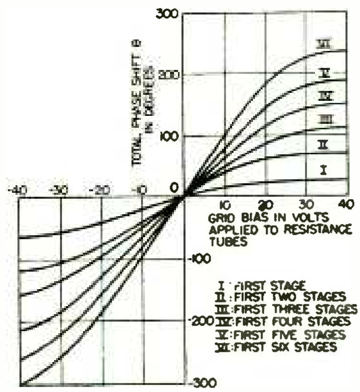



(*Courtesy Electronics*)

**Fig. 17.—Phase characteristics of constant-impedance network.**

A further point to note is the combination $R_1$ and $C_1$. The time constance $R_1C_1$ is sufficiently high compared to the period of a 30-cycle wave that the audio current is determined principally by $R_1$. Hence, if $e_m$ is constant in amplitude with frequency, the current $i_m$ through $R_1$ and $C_1$ will also be constant with frequency. The signal applied to the grid is $e_g = i_m(1/\omega C_1)$, and if $i_m$ is constant as well as $C_1$, $e_g$ varies inversely with $\omega$. This is the desired relation to convert phase into frequency modulation.

In Fig. 18 is shown the relationship between the total phase shift in degrees and the grid bias applied to the resistance tubes. the phase shift varies accordingly. It will be observed that over the




(*Courtesy Electronics*)

**Fig. 18.—Operating characteristics of cascade phase-shift modulator.**

range ±150°, the relationship is fairly linear, and indicates that a minimum of distortion will be encountered in modulating in this manner even at 30 c.p.s., which requires the above maximum of phase shift.

Measurements indicate a distortion of only 1.25 per cent for this range of phase shift (required for 30 c.p.s.). For 50 c.p.s., the required phase shift is correspondingly less, and as a result, the distortion is less than 0.6 per cent. Of course, for higher frequencies, the distortion is still less.

The f.m. noise has been found to be 72 db below a frequency deviation of 75 kc corresponding to 100 per cent modulation. An important point about the cascade phase-shift modulator is that the total signal phase shift is the sum of the phase shifts for the individual stages. Thus 6 stages produce six times the phase shift of one stage.

The noise produced by each stage, on the other hand, does not add in similar manner because of its random nature. Instead, it varies as the square root of the number of stages, or for 6 stages, is $\sqrt{6} = 2.45$ times as great as for one stage. On the other hand, if only a single phase-shift stage is employed, and the phase shift is therefore small, considerable frequency multiplying, including heterodyning, is required. Such multiplication multiplies the noise produced to exactly the same degree as the signal, so that greater care is required to obtain a final signal-to-noise ratio as high as that for the cascade phase-shift modulator.

To illustrate the noise problem, refer first to Fig. 19. Here a noise voltage component $E_n$ = 15,000 c.p.s. above or below the carrier

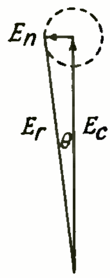



**Fig. 19.—Relation between noise voltage and signal for noise phase modulation.**

voltage $E_c$ in frequency, is superimposed on the latter. In the position shown, the resultant $E_r$ of $E_n$ and $E_c$ has the greatest angular deviation $\theta$ with respect to the carrier $E_c$.

If $E_n$ is small compared to $E_c$, the angle $\theta$ (in radians) and its tangent are practically equal, or

$$\theta_n \cong \tan\theta = E_n/E_c \qquad (7)$$

In degrees, this is $57.3\, E_n/E_c$. If $E_n$ = 10 $\mu$volts, and $E_c$ = 10 volts, $\theta_n$ comes out to be 0.0000573°. Next suppose that $E_c$ is modulated 100 per cent (75 kc at the final carrier frequency) at an audio rate of 15,000 c.p.s. If one stage of phase modulation is employed, a frequency multiplication of 6,000 is required for an initial phase shift of 24° at 30 c.p.s. to give a final 75 kc deviation at that low audio frequency. The same value of multiplication takes place, of course, for the 15,000-cycle signal.

Hence the signal frequency deviation at the modulator is

$$f_D = 75{,}000/6{,}000 = 12.5 \text{ c.p.s.}$$

This corresponds to a signal phase shift in degrees of

$$\theta_s = \frac{57.3\ f_D}{f_m} \qquad (8)$$

where $f_m$ is the modulating frequency. For $f_m$ = 15,000 c.p.s.,

$$\theta_s = \frac{57.3 \times 12.5}{15,000} = .0477^\circ$$

The signal-to-noise ratio is

$$S/N = \frac{.0477}{.0000573} = 833$$

In db this is

$$20 \log S/N = 20 \log (833)$$

$$= 20(2.9206) = 58.4 \text{ db}$$

Suppose, however, six cascaded phase-shift stages are employed, as in the Raytheon system. Then a frequency multiplication of but 6,000/6 = 1,000 is required. The frequency deviation for the signal is now six times as great, or $6 \times 12.5^\circ = 75^\circ$. The noise contributed by the six cascaded stages is $\sqrt{6} = 2.45$ times as great. Hence the signal-to-noise ratio is increased

$$6/2.45 = 2.45 \text{ times}$$

or in db

$$20 \log 2.45 = 7.8 \text{ db}$$

additional, giving a total of 58.4 + 7.8 = 66.2 db below 100 per cent modulation. However, it is to be noted that noise can originate in other parts of the transmitter, so that a reduction in noise at the modulator unit does not produce a proportionate reduction in the total noise.

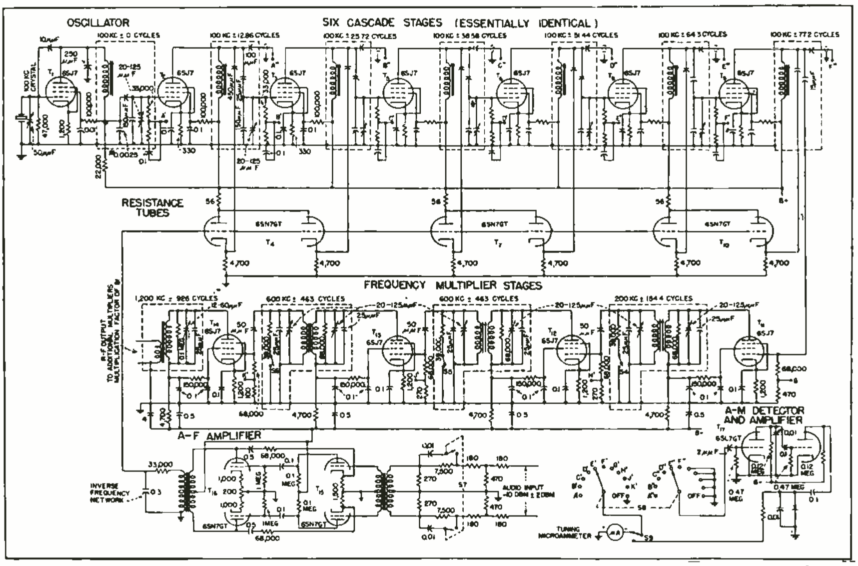
*CIRCUIT DETAILS.*—In Fig. 20 is shown a schematic diagram of the modulator unit. The crystal oscillator (top left) is of the conventional tuned-plate tuned-grid type, with the crystal acting as the tuned circuit for the grid. The first modulator stage (to the right of the oscillator) is fed the oscillator voltage, developed across the 0.0025-$\mu$f capacitor below the plate inductance. This voltage acts on a capacity voltage divider composed of a 150 $\mu\mu$f capacitor in series with two capacitors, 150 $\mu\mu$f and 20-125 $\mu\mu$f (variable) in parallel. The variable capacitor enables the grid excitation voltage to be set to the proper level for the 6SN7GT tube acting as a variable resistor to have the proper resistance variation for linearity between the phase shift and the modulating signal.

It will be observed from the figure that there are six 6SJ7 tubes functioning as constant-current sources for the six phase-shift networks, and three 6SN7GT double-triodes acting as resistance tubes. (The 56-ohm resistors in the plate circuits of the latter tubes are to prevent parasitic oscillations between them.)

In adjusting the modulator, the first step is to adjust the voltage input level of each tube ($T_1$ to $T_9$). To do this, the microammeter (lower right) is arranged by an associated switch to connect to any one of the terminals marked A′, B′, C′, etc. It now reads the grid current of the tube and thereby functions as a crest vacuum-tube voltmeter. The various trimmer capacitors are set so that the voltage readings are adjusted to the desired level settings.

The next step is to adjust the inductance in the plate circuit of each modulator stage so that its reactance is twice that of the associated capacitor. It will be recalled that this condition—Eq. (3)—makes the plate load constant and independent of variations in the resistance of the resistance tube.

If amplitude modulation is present, the r.f. signal detected by the first grid of the 6SL7GT A.M. DETECTOR produces a pulsating d.c. in the plate circuit; this is passed through the 0.01-µf coupling capacitor to the second grid, and causes




*(Courtesy Electronics)*

Fig. 20.—Schematic diagram of complete cascade phase-shift type of modulator.

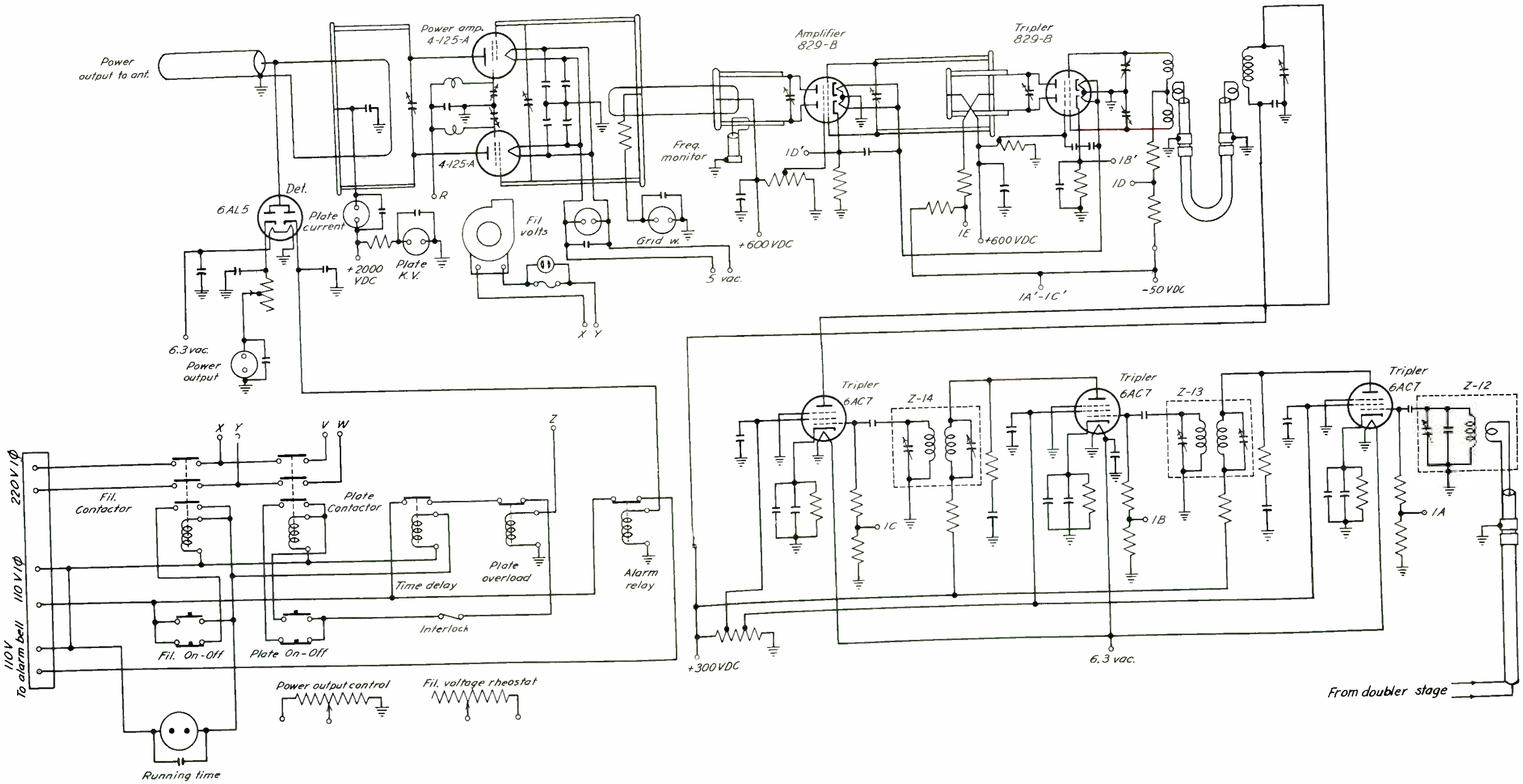
This in turn means that as the grids of the latter tubes are fed the modulating signal, no variation in the output of the modulator stage occurs; i.e.—no amplitude modulation.

The presence of amplitude modulation in the output therefore indicates improper adjustment of the inductance. To check, the microammeter is switched to the output half of the 6SL7GT A.M. DETECTOR AND AMPLIFIER, and the grid of the other half of this double diode is connected to the terminal under test, such as A″, B″, etc.

An a.f. modulating signal is applied through the A.F. AMPLIFIER channel, consisting of the 6SN7GT double-diode push-pull amplifier.

a pulsation in the output plate current. The pulsation or a.c. component can then pass through the 0.1-µf capacitor to the rectifier element shunted by the 0.01-µf capacitor, and the resulting d.c. read on the microammeter.

The inductance tuning slug is then adjusted until the reading on the microammeter is a minimum. The adjustment is made, starting with the first stage, and proceeding until finally the sixth and last stage is adjusted. It is first necessary, however, that the r.f. signal level at each modulator grid be of the proper level before the inductances are tuned, so that the procedure for setting the signal level, as described previously,




(Courtesy of Raytheon)

Fig. 21. —Frequency-multiplying and power-amplifier stages for Raytheon RF-250 f.m. transmitters.

must first be made.

The frequency multiplier stages are band-pass coupled in order to pass the side bands produced by frequency modulation. It will be observed that there is first a frequency doubler, then a tripler, followed by a buffer stage, and then another doubler. This gives a total multiplication of 12. A further multiplication of 81 then takes place in the power amplifier unit to give a total multiplication of 972 times.

The band-pass double-tuned transformers are coupled somewhat more than the critical value, so that they exhibit a slight double-peaked resonance curve. It will be recalled from a previous assignment that the value for critical coupling is given by

$$K_c = \frac{1}{\sqrt{Q_p Q_s}} \qquad (9)$$

where $Q_p$ is the Q of the primary circuit, and $Q_s$ is the Q of the secondary circuit.

If either or both Q's are reduced, the critical coupling value is increased, and can thereupon exceed the actual value of coupling employed. In this case the double-peaked curve reverts to a single-peaked curve. In checking the circuit for alignment, the capacitors can be adjusted so that the peak occurs at the carrier frequency, thus obviating an asymmetrical selectivity curve with respect to the frequency-modulated signal. (An asymmetrical curve increases the harmonic distortion and may even reduce the frequency deviation for the higher modulating frequencies.)

To reduce the Q of the stage a push-button switch is arranged in the secondary of each band-pass transformer, so as to connect a 39,000-ohm resistor from the high side of the secondary to ground. This additional load on the secondary reduces $Q_s$, thereby increasing $K_c$ above the actual value of coupling employed, and thus producing a single-peaked curve. Note that by locating the switch in the grounded side of the circuit no change occurs in the secondary capacity owing to the switching action. It will be observed that the adjustments are simple and are readily made; a total adjustment time of less than ten minutes is claimed for this system.

In Fig. 21 is shown the subsequent multiplying and amplifying stages for the 250-watt transmitter. It will be observed that three 6AC7 tubes act as frequency triplers, followed by an 829-B tripler stage. Of interest is the coaxial coupling circuit, with a loop at either end, that transfers power from the last 6AC7 tank to the input tank of the 829-B stage.

The output tank of the latter is of the transmission line type, with a shunt capacitor to tune the line to resonance. Note that two such lines are employed, one for the plates, and one for the grids of the succeeding 829-B amplifier stage. Thus, while coupling between the two serves to transfer r.f. power from one line to the other each can be biased as desired: the plate tank has a d.c. potential of 600 volts, and the grid tank of -50 volts.

The plates of the 829-B amplifier stage are connected in turn to a tuned transmission line, which is link-coupled to a tuned transmission line feeding the power amplifier grids. The two power amplifier tubes are of the type 4-125-A. The plates are connected to a tuned transmission line, which is loop coupled to the antenna coaxial cable.

Observe the coaxial feed to the frequency monitor from the 829-B amplifier stage. Also observe the 6AL5 rectifier stage employed, in conjunction with a d.c. meter, to read the power output of the 4-125-A stage. In addition various relay switches are to be noted for closing filament and plate circuits, in conjunction with time delay and overload relays for the latter circuits.

A further point is the use of capacitors in the screen-grid circuits of the output stage to tune out the lead inductances, together with r.f. chokes in series with the d.c. screen supply to isolate the two screens from one another and from the supply. These two measures respectively place the screens at ground potential, and prevent parasitic oscillations from being set up.

*CONSTRUCTIONAL DETAILS.*—In Fig. 22 is shown the 1-kw transmitter. The 3 kw is similar in appearance, but has a small cubicle containing the 3-kw amplifier power supply inserted between the two cubicles shown. In either case the left-hand cubicle contains the cascade-shift modulator and a 250-watt exciter unit which, incidentally, forms the power amplifier for the 250-watt transmitter.

In the case of the 10-kw unit, the 3-kw transmitter is employed, together with an additional cubicle containing the 10-kw amplifier. It will be observed that any size equipment can be readily increased to the next size by the addition of the appropriate cubicle; this makes for flexibility and ease in expansion of the station.

The construction involves vertical type chassis and full height front and rear doors to facilitate servicing of the equipment. Meters are mounted on the front so that readings can be made at a glance. All output stages employ push-pull tetrodes operating in conventional linear tank circuits, except the 10-kw transmitter which employs push-pull triodes, and all stages are completely shielded to eliminate power losses by radiation, interaction, and parasitic oscillation. A mechanical feature is that unit dimensions have been held to convenient cubicle sizes for moving through standard doorways in elevators and rooms.

*(Courtesy of Raytheon)*

Fig. 22.—Raytheon RF-1 1-kw transmitter.

The 1-kw unit requires 220-volt, single-phase 60-cycle power of approximately $5^1/_2$ kva; the 3-kw unit requires 220-volt three-phase 60-cycle power of approximately 8 kva, and the 10-kw unit requires 220-volt 3-phase, 60-cycle power of approximately 22 kva.

Some of the electrical characteristics will be of interest. The f.m. noise level is at least 65 db below 100 per cent modulation, and the a.m. noise level is at least 50 db below 100 per cent modulation. The distortion is less than $1^1/_2$ per cent for frequencies from 50 to 100 c.p.s., and from 7,500 to 15,000 c.p.s. It is less than 1 per cent for frequencies from 100 to 7,500 c.p.s. These figures indicate upper limits which will not be exceeded; the actual measured noise and distortion values have been given previously, and are appreciably below these limits.

The frequency stability is given as better than ±1,000 c.p.s., which is half of the F.C.C. requirements of ±2,000 c.p.s. The speech input level is +10 db ±2 db for 100 per cent modulation (±75 kc swing) for single frequency modulation. (A maximum frequency deviation of ±100 kc can be produced without appreciable distortion.)

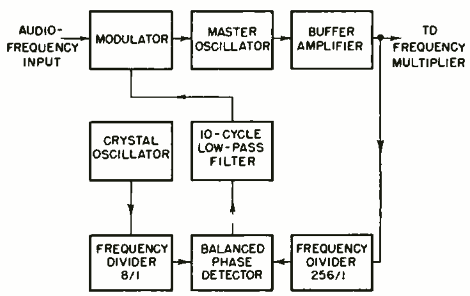
## THE FEDERAL F.M. TRANSMITTER

*BASIC THEORY.*—The basic principle of the Federal (I.T. &T.) f.m. transmitter is the use of a master oscillator, whose frequency is controlled by a variable capacitor. This capacitor is actually an electronic vacuum tube circuit, so arranged as to take advantage of the so-called Miller effect in producing an adjustable capacitance at its input terminals, which are connected across the tank circuit of the master oscillator.

The capacitance is varied by the audio modulating voltage, so that direct frequency modulation is produced. At the same time the demultiplied frequency of the master oscillator is compared with a crystal oscillator; any drift is used to apply a corrective bias to the same capacitance tube, and thus bring the master oscillator back to the proper frequency. In short, the same control tube produces both frequency modulation and carrier frequency control.

A block diagram of the system is shown in Fig. 23. The audio frequency varies the capacitive reactance of the vacuum tube circuit in the modulator, which in turn varies the frequency of the master oscillator. The frequency-modulated output is fed to a buffer amplifier, which then feeds the frequency multiplier circuits that raise both the carrier frequency and the frequency deviation to the desired values.

At the same time the buffer




*(Courtesy Elect. Comm.)*

Fig. 23.—Block diagram of frequematic modulator and center-frequency-stabilization system.

amplifier feeds frequency-divider circuits that yield an output frequency 1/256th of the oscillator frequency. Similarly, a crystal oscillator feeds another 8 to 1 frequency-divider stage whose output is that of the nominal frequency of the other frequency-divider. The two are compared in a balanced phase modulator, and any difference or drift between the two appears in its output.

The beat frequency is fed through a 10-cycle low-pass filter which eliminates the audio frequency modulation but transmits the slower drift variations. These are then applied as a bias to the modulator tube, changing its capacitive reactance in the appropriate direction to collect the drift in frequency.

*THE FREQUEMATIC MODULATOR.*—The capacitance tube forms the heart of the modulator unit, which is called the "frequematic" modulator. As stated previously, the capacitive reactance is a result of the "Miller" effect. This, in turn, is a special example of feedback, normally produced by the grid-to-plate capacitance.

Owing to this feedback, the capacitor $C_{gp}$, connected between the grid and the plate, appears to be a much larger capacitor connected between the grid and cathode terminals. Its apparent value is

$$C = C_{gp} + \frac{\mu R_L C_{gp}}{R_p + R_L} \tag{10}$$

where $R_L$ is the plate load resistance, $R_p$ is the tube's plate resistance, and $\mu$ is its amplification factor.

If the tube employed is of the pentode type, $R_p$ is normally much greater than $R_L$, so that $(R_p + R_L)$ is practically equal to $R_p$. In this case Eq. (10) reduces to

$$C' \cong C_{gp} + \frac{\mu R_L C_{gp}}{R_p}$$

$$= C_{gp} + G_m R_L C_{gp} \tag{11}$$

since $\mu/R_p = G_m$, the transconductance of the tube.

Eq. (11) indicates that the apparent input capacitance is composed of a fixed part $C_{gp}$ plus a part $G_m R_L C_{gp}$ which is in general very much larger, and moreover, *can be varied electronically* by varying $G_m$, such as by changing the control grid voltage. Thus, an audio voltage applied in series with the r.f. voltage to the grid can vary the $G_m$ and hence the apparent capacitance that the tube presents to the r.f. tank circuit.

As an example, suppose $C_{gp} = 2\ \mu\mu f$, $G_m = 2{,}000\ \mu$mhos and $R_L = 50{,}000$ ohms. Then, from Eq. (11), the capacitance looking into the input terminals is

$$C' = 2 + (2{,}000 \times 10^{-6})(50{,}000)(2)$$

$$= 2 + 200 = 202\ \mu\mu f$$

If, upon the extreme negative swing of the grid, $G_m$ decreases to 5 $\mu$mhos, $C'$ will decrease to

$$2 + (5 \times 10^{-6})(50{,}000)(2) = 2 + .5$$

$$= 2.5\ \mu\mu f$$

The total change is 202 - 2.5 = 199.5 $\mu\mu f$. This is sufficient to cause a considerable frequency deviation even at relatively low radio frequencies.

Eq. (11) assumes a plate load resistance $R_L$. In general, the load is equivalent to a parallel tuned circuit, which appears as a resistance only at the resonant frequency. Below the resonant frequency it appears as an inductive reactance; above the resonant frequency, it appears as a capacitive reactance.

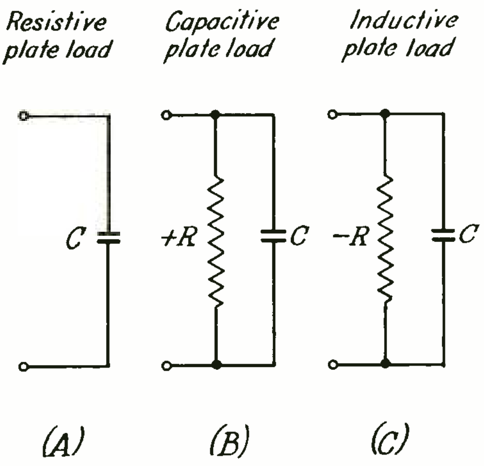



Fig. 24.—Apparent input impedance for different kinds of plate loads.

It is shown in Appendix II that if the plate load is a capacitive reactance, the input admittance is not that of a pure capacitive, but corresponds to a capacitor and resistor in parallel, as is indicated in Fig. 24 (B). The capacitor will alter the tuning of a resonant circuit connected to the input terminals, and the resistance will increase the damping (lower the Q) of the tuned circuit. In the case of the modulator, which is connected to the tank circuit of the master oscillator, the capacitance will lower the oscillator frequency, and the resistance will decrease the amplitude of oscillation.

If, on the other hand, the plate load is an inductive reactance, the input admittance is that of a capacitor and a *negative* resistance in parallel [see Fig. 24 (C)]. In that case the frequency of the connected master oscillator will decrease, but the amplitude of oscillation will *increase*, owing to the negative resistance effect. The negative resistance represents regenerative feedback from the plate to the grid circuit; the positive resistance for an capacitive plate load represents degenerative feedback.*

To summarize:

1. Owing to the presence of a capacitance between the plate and control grid of the vacuum tube, feedback takes place.

2. The nature of the feedback, such as whether it is regenerative or degenerative, depends upon the type of plate load employed:

a. If the plate load is resistive at the frequency of operation, the feedback is such as to make the grid input circuit appear to have a pure capacitive reactance; i.e., as if only a capacitor were connected across the terminals.

b. If the plate load is inductive, regenerative feedback takes place, and the input impedance

*In the early days, when triode tubes only were available, regenerative feedback of the above type in r.f. amplifier stages was a stumbling block to their successful operation, and was obviated by the use of neutralization. The advent of the screen-grid tube reduced $C_{gp}$ to a negligible value owing to the shielding action of the screen between the plate and control grid, and permitted stable r.f. amplifiers without the need for neutralization, which was effective only over a relatively narrow range. On the other hand, transmitter tubes which operate in general at one fixed frequency and in the larger sizes are of triode construction, are readily neutralized, at least at the lower frequencies. At the higher frequencies, such as in the f.m. range, triode tubes are often operated as grounded-grid amplifiers to obviate the need for neutralization, which becomes more difficult at the higher frequencies.

appears to be that of a capacitor paralleled by a negative conductance, which is in other words a negative resistance. If the grid signal source is a tuned circuit of low damping, oscillation can readily take place owing to the negative resistance.

c. If the plate load is capacitive, degenerative feedback takes place, and the input impedance appears to be that of a capacitor and positive conductance (resistance) in parallel. A tuned input source will then be more damped owing to this positive resistance component.

3. Note that in all cases the input impedance involves a capacitance. This is the circuit element effect desired; the presence of a positive or negative resistance is undesirable, particularly in the f.m. modulator unit.

The reason that the resistive component is undesirable is that it tends to vary the amplitude of the oscillator output at the same time that the capacitive component varies its frequency. Thus amplitude as well as frequency modulation will take place, and the amplitude modulation can in turn react to vary the frequency of the oscillator, thus introducing distortion in the frequency modulation. Hence, the plate load of the modulator tube should be a pure resistance over the frequency range of modulation; the modulator then presents a pure capacitive reactance to the oscillator tank circuit, and essentially only frequency modulation takes place.

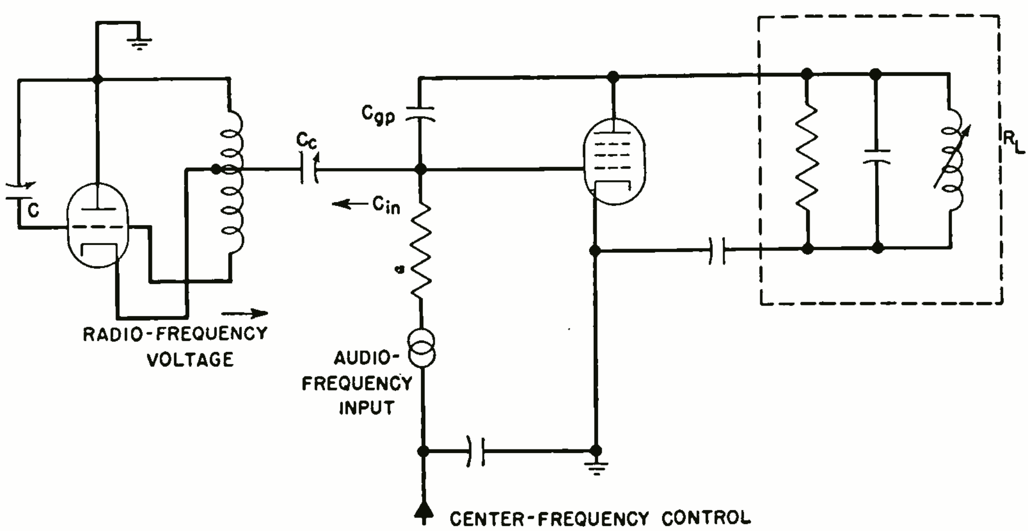
The circuit is shown in simplified form in Fig. 25. A Hartley oscillator is employed; the only feature being that the plate is grounded. The tank circuit is temperature compensated to reduce frequency drift to a minimum. A powdered-iron core in the tank coil permits the frequency to be adjusted over a range from 3.4 to 4.7 mc.

The modulator tube is a pentode of the super-control type, in order that its transconductance can be varied by varying the electrode voltages, such as the control grid bias. The change in transconductance in turn alters the capacitive reactance which its grid circuit presents to the oscillator tank.

The modulator plate load is a tuned circuit, as shown. As stated previously, a tuned circuit normally appears as a pure resistance at the resonant frequency; as an inductive reactance below the resonant frequency; and as a capacitive reactance above the resonant frequency. As was also just explained, such a circuit would exhibit an additional negative resistance component below the resonant frequency, and an additional positive resistive component above the resonant frequency, so that amplitude modulation would take place in the oscillator as it is frequency modulated.

However, by damping the tuned plate load with a low shunt resistor, the impedance remains essentially resistive over the range of frequency deviations encountered, so that the tube presents practically a pure capacitive reactance to the oscillator tank. Moreover, the low Q of the modulator plate load makes its tuning less critical and its adjustment relatively easy to perform.

The use of a super-control tube of the pentode type results in a very low plate-to-grid inter-electrode capacitance, since this is the normal function of a pentode or tetrode tube. The Miller effect or feedback will therefore be very grid as shown. By the proper selection of tube, circuit components,




(*Courtesy Electrical Communications*)

Fig. 25.—Basic circuit of modulator.

small. To enhance this effect, an external capacitor $C_{gp}$ is connected between the two electrodes, as shown in the figure; the pentode construction is used because it lends itself more readily to supercontrol properties. Capacitor $C_c$ couples the input capacitance of the modulator tube to the tank circuit and at the same time isolates the d.c. difference in potential between these two points. In order to permit the same frequency deviation to be obtained throughout the f.m. range, it is made adjustable to fit the operating frequency assigned. It is tapped down on the tank coil to prevent excessive r.f. excitation from being applied to the modulator grid.

The audio signal, in series with the low-frequency (essentially d.c.) center-frequency control voltages, are applied to the modulator

and electrode voltages, it is possible to obtain a capacitance variation such that the oscillator frequency deviation is practically directly proportional to the audio and center-frequency control voltages.

To obtain this effect, it is to be noted that the frequency of a resonant circuit changes by an amount $\Delta f$ such that

$$\Delta f = -\frac{f}{2C}\Delta C \qquad (12)$$

where f is the initial frequency determined by the inductance L and capacity C of the resonant circuit, and $\Delta C$ is the subsequent change in capacity C. Note from Eq. (12) that $\Delta f$, the frequency deviation, is directly proportional to $\Delta C$, as well as incidentally to f, and inversely proportional to C. This is true for small change $\Delta C$ relative to the initial value C.

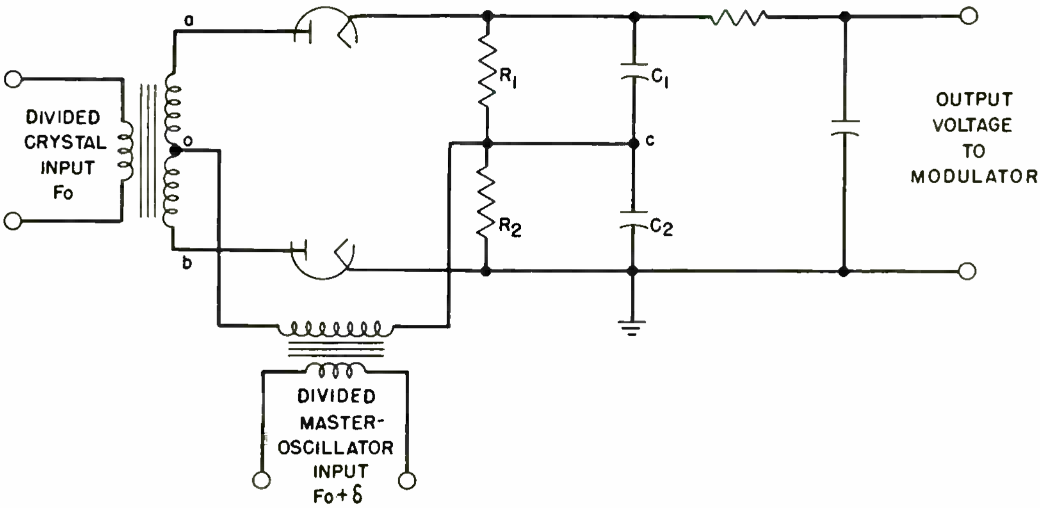
The significance of this is that if the modulator tube is adjusted so that its $G_m$ is directly proportional to the modulating voltage, then—by Eq. (11)—the change in capacitance $\Delta C$ it produces by the Miller effect will also be proportional to the modulating voltage, whereupon the frequency deviation $\Delta f$ will in turn be linearly related to the modulating voltage. This, of course, is the desired result that yields distortionless frequency modulation.

*CENTER-FREQUENCY CONTROL.*—The center-frequency control circuits have been described in block-diagram form. The frequency dividers are of the multivibrator type. For the frequency-modulated oscillator the division is 256:1; for the crystal oscillator, it is 8:1, which means that the crystal oscillator has a frequency 1/32 that of the f.m. oscillator. The reason that the latter has such a higher frequency (in the range from 3.4 to 4.7 mc), is that the capacitance variation in the modulator tube is more pronounced at this higher frequency, and hence produces a greater frequency deviation.

The two subdivided frequencies are applied to a balanced phase detector. This is shown in Fig. 26. It will be recognized as practically identical with the balanced discriminator used in an f.m. receiver and described in a previous assignment.

There a voltage between the center tap of the secondary and ground was introduced by connecting the high side of the primary winding through a coupling capacitor to the secondary center tap. In addition the secondary was tuned to the center frequency.

Thus the primary voltage and the two half-secondary voltages were all of the same frequency; their phase relationships and hence vector combination with respect to the two




(Courtesy Electrical Communications,

Fig. 26.—Basic circuit of balanced phase detector.

diodes depended upon whether they were higher, lower, or equal to in frequency to that of the tuned circuit. Upon this relation depended the magnitude and polarity of the output voltage.

In Fig. 26 the two half-secondary voltages are of one frequency $f_o$: that of the divided crystal input. The voltage from the secondary center tap to ground is that of the divided master oscillator input, $f_o + \delta$, where $\delta$, is a small frequency drift (positive or negative) of the master oscillator. (Note that the connection of the secondary of the divided master oscillator transformer to the junction of $R_1$ and $R_2$ is essentially that to ground owing to the by-passing action of $C_2$.)

The vector conditions are therefore essentially the same as for a receiver discriminator and are shown in Fig. 27. In (A), the divider master oscillator is assumed to be of the same frequency as that of the divided crystal oscillator. The voltage of the divided master oscillator is OA; that of the divided crystal oscillator is AB for the top half-secondary, and AC for the bottom half-secondary. The voltage across the top diode is the vector sum of OA + AB or OB; the voltage across the bottom diode is OC, and is equal to OB. The d.c. voltage across $R_1$ is equal and opposite to that across $R_2$, and there is no output voltage.

Suppose that the divided master oscillator frequency begins to increase above that of the crystal. Vector OA begins to rotate counterclockwise with respect to AB and AC, as is indicated in Fig. 27(B). The voltage OB now exceeds OC, thus the d.c. voltage across $R_1$ exceeds that across $R_2$, and there is a positive voltage developed at the output of the detector. This is applied as a bias to the modulator tube, increasing its capacity effect across the master oscillator tank circuit and thus decreasing the frequency, or rather, preventing the frequency from increasing except momentarily in the nature of a phase advance of vector OA from its perpendicular position in Fig. 27(A) to its oblique position in (B).

If the frequency tends to decrease, vector OA starts to rotate clockwise as in (C), thereby making OC exceed OB. The corrective effect on the master oscillator is now in the opposite direction, and holds OA in the phase shown in (C). In short, any attempt to change the

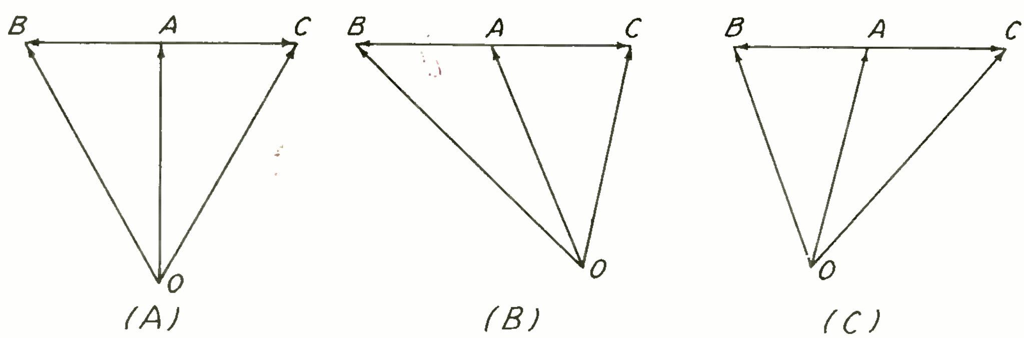



Fig. 27.—Vector relations for the balanced phase detector.

frequency of the master oscillator results merely in a shift in phase of the master oscillator relative to the crystal-controlled oscillator, without any actual relative frequency drift. The master oscillator is therefore as stable in frequency as the crystal oscillator.

The maximum amount of correction possible is when OA is parallel to AB or AC. This means that the most OA can shift in phase from its perpendicular position in (A) is ±90°. If the constants of the master oscillator change by a sufficient amount, synchronism will be broken and control lost. However, frequency-modulation of the master oscillator does not exceed about ±24° phase shift after demultiplication, which is less than the ±90° available for corrective purposes, so that there is sufficient reserve for correction. But since even the ±24° phase shift would tend to be counteracted by the stabilizing circuit, so that a large reduction in the frequency modulation would occur, the 10-cycle cutoff filter is connected to the output of the phase detector to permit only corrections to low drift variations to be applied to the modulator tube.

In starting up the equipment, there is a period during which the master oscillator is running free. So long as its divided frequency does not exceed the divided oscillator by more than 10 cycles/sec., the output of the phase detector will pass through the 10-cycle filter and apply a corrective effect. Owing to the large amount of total multiplication produced, this corresponds to 61-kc drift in the final carrier, and to 2,560 c.p.s. in the master oscillator. The latter is so designed (temperature compensation of the tuned circuits) that it does not depart from its normal value by 2,560 c.p.s., so that it is locked very readily even when first turned on: synchronism occurs before the final stages are operative.

*CONSTRUCTIONAL DETAILS.*—Federal transmitters are rated in powers of 1, 3, 10, and 50 kw. In Fig. 28 is shown a photograph of the 1-kw and 3-kw transmitters. They are identical in tubes and appearance; the only difference is that the 1 kw operates from a 220-volt single-phase supply, and furnishes 2,000 volts to the power amplifier, whereas the 3 kw operates from a 230-volt three-phase supply, and furnishes 3,000 volts to the power amplifier, thereby obtaining the increased power output.

The photograph shows the unit type of construction employed in all transmitters. Two units, the modulator-exciter unit and the power-amplifier unit, are bolted together, then side panels, a top, doors, and decorative trim added to give it the appearance of being housed in a a single cabinet.

The inner doors giving access to the individual compartments involving high voltages are interlocked with the high-voltage power supplies for safety precautions. All vacuum tubes can be replaced from the front; in addition, doors in the rear provide access to components mounted in the rear of the chassis. All controls, switches, etc., are mounted alongside the front inner doors, so that all adjustments may be made from the front.

All tubes are air-cooled by air supplied from blowers through ducts and filters. The power-amplifier unit in particular, has its trans-

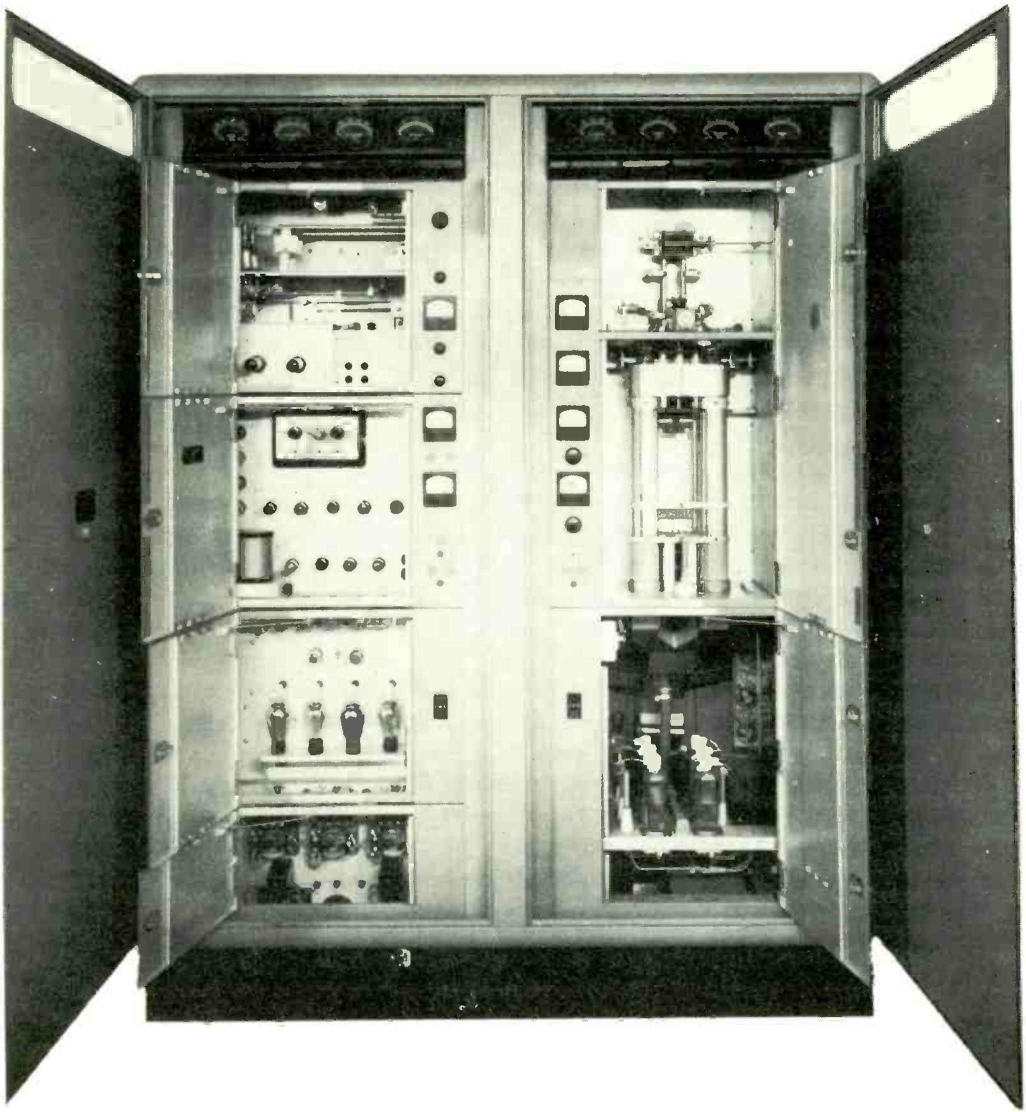


Fig. 28.—Front view of the Federal 1- and 3-kw transmitters.

mission-line-type plate inductor cooled as well as the amplifier tubes and the bases of the rectifier tubes. The air is then discharged through louvers at the upper rear and top portion of each assembly.

Although a single "on-off" switch normally controls the transmitter individual plate-, bias-, and control-voltage, switches are provided for maintenance and adjustment of the transmitter. Sequencing and time-delay cycling are completely automatic, and a set of supervisory lights are arranged in the order of sequencing and cycling of the transmitter to enable faulty operation or failure to operate at any point to be detected instantly. Control switches and indicating lamps can be located, if desired, at a remote operating point. Finally, a buzzer is employed to give audible warning of plate-current overload and of excessive temperatures in the transmitter.

The modulator unit employs a 12J5 oscillator, a 6AB7 modulator, and a 6AB7 buffer. In the frequency multiplying section, three type 1614 tubes act as doubler stages, and one type 815 dual beam power tube acts as the tripler. The first two doublers, operating at lower frequencies, require damping resistors in their tank circuits to lower the Q sufficiently to pass the required side bands, but the third doubler is at a sufficiently high frequency to have sufficient band width for the normal Q. The two halves of the 815 tripler are connected in push-pull Class C. The plate tank has a variable capacitor and a transmission-line-type inductor, with a movable shorting bar to afford means of adjusting the effective length of the inductor.

The tripler stage is capacitively coupled to the grids of a buffer stage. The latter also employs an 815 tube in push-pull Class C arrangement, with a similar variable capacitor and transmission-line-type inductor. Neutralization is accomplished by mounting externally small studs adjacent to the plate and grid leads, with the studs cross-connected.

The buffer then feeds a 250-watt amplifier employing two Eimac type 4-250 A tetrodes in a push-pull Class C arrangement. Transmission-line-type inductors are used both in the plate and grid circuits, and the latter is inductively coupled to the plate inductor of the buffer amplifier stage. The plate lines in the 250-watt amplifier are in turn inductively coupled by an adjustable loop to the following stage. The 250-watt stage is neutralized by grounding the screens through series-resonant circuits tuned to the frequency of operation. The series-resonant circuit consists of a split-stator capacitor; one set of stator plates connects to each screen grid, and the rotor to ground. The inductance between the screens and ground is tuned to series resonance by adjusting the capacitor, and the screens thereupon operate more effectively as shields between the respective control grids and plates of the tubes.

The power amplifier stage employs two Federal type 7C26 tubes in a neutralized push-pull circuit. The 250-watt amplifier feeds a balanced transmission line, which is loop-coupled to a transmission-line-type grid inductor. This inductor, in conjunction with a variable capacitor, forms the power amplifier's grid tank circuit.

A similar tank network is used in the plate circuit, together with a dual capacitor cross-connected between the plates and opposite grids to afford neutralization. A variable loop then couples the amplifier to an output matching network, which transforms the balanced impedance of the tank circuit into the single-ended 51-ohm antenna transmission-line impedance, and also makes the power amplifier tuning independent of the coupled antenna load. Coupling loops in the network are used to operate a thermocouple-type meter and a program monitor.

In the 10-kw transmitter, the two units of the 3-kw transmitter are used to drive the 10-kw power amplifier, and a 4,000-volt supply is added. Similarly, the 10-kw power amplifier is used to drive the 50-kw transmitter. This makes for flexibility in design and planning of an f.m. station; expansion can take place without necessarily requiring the older units to be discarded.

The center frequency drift is given as 1,000 c.p.s., or half of that permitted by the F.C.C. Since the master oscillator is locked to the crystal oscillator, the above drift represents that of the crystal oscillator itself. The frequency response is flat within ±1 db from 30 to 15,000 c.p.s., and the distortion is below 1.5 per cent between 50 and 15,000 c.p.s., and below 1 per cent from 100 to 7,500 c.p.s. The f.m. and a.m. noise levels are 65 and 50 db below full carrier modulation, and the modulation capabilities are considerably in excess of ±75 kc with very little distortion.

## THE WESTINGHOUSE F.M. TRANSMITTER

*BASIC THEORY*.—The Westinghouse transmitter is also of the direct frequency-modulated type, and employs a master oscillator and reactance tube circuit. This circuit also acts to correct for any drift in the frequency of the master oscillator.

The principle involved in the carrier frequency stabilizing device is quite different from the methods employed by other companies, and is an interesting example of pulse techniques. It will be described in detail in a following section.

An important characteristic of this control is that it is not necessary to demultiply the frequency-modulated master-oscillator wave before comparing it with the output of the crystal oscillator. Instead, the device compares the number of cycles by which the master oscillator frequency exceeds that of the crystal oscillator during one time interval with the number of cycles by which it is less then that of the crystal oscillator in the next time interval. If the excess is not equal to the deficit, a corrective voltage is developed which brings the two into equality. This therefore prevents any appreciable relative drift in either direction, and makes the frequency stability that of the crystal oscillator.

*MODULATOR CIRCUIT*.—The modulator circuit is shown in schematic form in Fig. 29.

The audio input can be applied through a preemphasis network consisting of a parallel R-C combination that passes the "highs" better than the "lows," or through an ordinary resistor, to the 6SJ7 audio amplifier tube. This is resistance coupled to a 1614 modulator control tube, whose plate (and screen) are by-passed to ground by capacitor $C_2$, and are also connected to the direction, the modulator control plate voltage drops, and the diode cathodes connected to the plate thereupon also go down in potential.

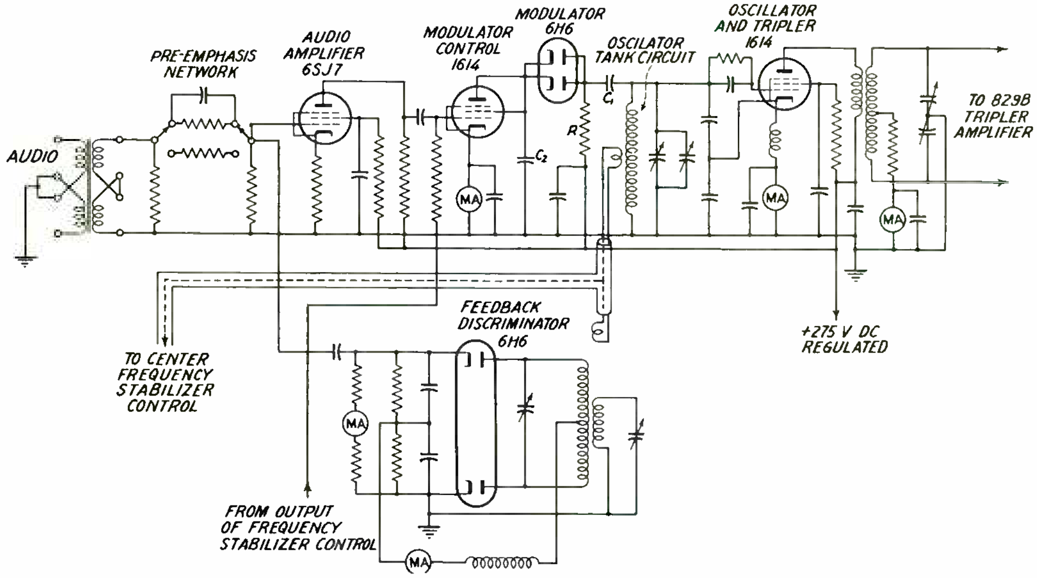



Fig. 29.—Schematic diagram of Westinghouse frequency modulator.

6H6 modulator tube. The combination of the two tubes is essentially connected across the oscillator tank circuit, and functions as a variable capacitor under control of the audio signal.

The principle involved is that the diode modulator tube is biased by the modulator control tube in a variable manner depending upon the input audio signal. The oscillator develops an r.f. voltage across its tank circuit. This voltage is also impressed across the modulator combination. Current can flow through the diode only for a portion of the r.f. cycle; the angle of flow depends upon the audio bias of the diode cathodes relative to the diode plates.

Specifically, if the modulator control grid is driven in a positive The diode plates are connected through resistor R to the fixed potential of +275 volts. Hence the diode is made more conductive since its plates are now relatively positive to its cathodes. As a result, r.f. current can flow through $C_1$, the diode, and by-pass capacitor $C_2$ over a large portion of its cycle. The combination therefore acts as a relatively large capacitor, and lowers the oscillator frequency.

On the other hand, when the modulator control grid is driven in a negative direction by the audio signal, the modulator control plate and diode cathodes go up in potential relative to the diode plates, and the diode conducts for a small portion of the r.f. cycle. The combination therefore appears as a small capacitor, little leading current is drawn from the tank, and the oscillator frequency goes up. The action is very similar in many respects to the electronically controlled resistor in the Raytheon phase modulator.

Note that the bottom end of the modulator control grid resistor is connected to the output of the frequency stabilizer control. The latter produces a negative bias which varies in accordance with the relative drift between the master and crystal oscillators, and thus superimposes on the audio frequency bias of the diodes a correction bias that corrects the relative drift.

An additional feature to note is the inverse feedback circuit. Some of the frequency-modulated output of the oscillator is fed into a balanced receiver-type discriminator through a link circuit. The output of the discriminator is then fed degeneratively into the grid of the 6SJ7 audio amplifier, and tends to eliminate any distortion in the f.m. wave that may be caused by nonlinearity in the modulator between audio input and capacitive variation across the tank circuit. It is claimed for the diode-type modulator, however, that it is freer from noise than the ordinary reactance type modulators.

*FREQUENCY STABILIZING CONTROL.*—Mention was made that the frequency stabilizing control circuit operates so as to insure that the number of cycles by which the master oscillator is high compared to the crystal oscillator during one period of time equals the number of cycles by which it is low during a following period of time.

The cycles are converted into narrow pulses of one polarity if they represent an excess, and in opposite polarity if they represent a deficit. The two sets of pulses are fed through diodes (unilateral conductors) to a capacitor. Pulses of one polarity tend to charge it up; pulses of opposite polarity tend to discharge it. Ordinary frequency modulation produces equal numbers of pulses of each polarity, hence the average capacitor charge and voltage remains constant. Frequency drift, however, produces unequal numbers of positive and negative pulses; both the capacitor charge and voltage change and effect a correction by acting on the modulator control grid.

In Fig. 30(A) is shown a block diagram of the circuit for producing first a two-phase source of beat-frequency voltage. The master-oscillator frequency is denoted by $f_s$; the crystal-oscillator frequency, by $f_c$. The latter is passed through two oppositely connected phase-shift networks that produce a +45° and -45° phase shift respectively. The two shifted crystal-oscillator voltages are then mixed in two mixer tubes (A and B) to produce two beat frequencies.

The output of mixer A is shown as passing through zero at the start. Then the output of mixer B starts at its crest value (90° out of phase with that from mixer A). If $f_s$ is greater than $f_c$, the output of mixer B leads that of mixer A [Fig. 30(B)], if $f_s$ is less than $f_c$, the output of mixer B lags that of mixer A [Fig. 30(C)]. This is exactly the same condition as exists in the RCA motor-control system: as $f_s$ varies from above to below $f_c$, one beat frequency reverses in phase while the other does not.

The output of mixer A is then

used to trigger a multivibrator of the direct-coupled type, as illustrated in Fig. 31. This type of multivibrator is aperiodic; i.e., it has no natural frequency, since no coupling capacitors and consequently R-C time constants are involved. The device therefore remains in one condition (plate current of one tube at a maximum and that of the other tube at a minimum) until a grid pulse is applied to bring the cutoff tube into operative condition, whereupon there is a reversal in conditions, with the other tube at cutoff. This state of affairs continues until the next pulse triggers the next grid, whereupon the device reverts to its initial condition.

The triggering is done by the sine wave output of mixer A. The

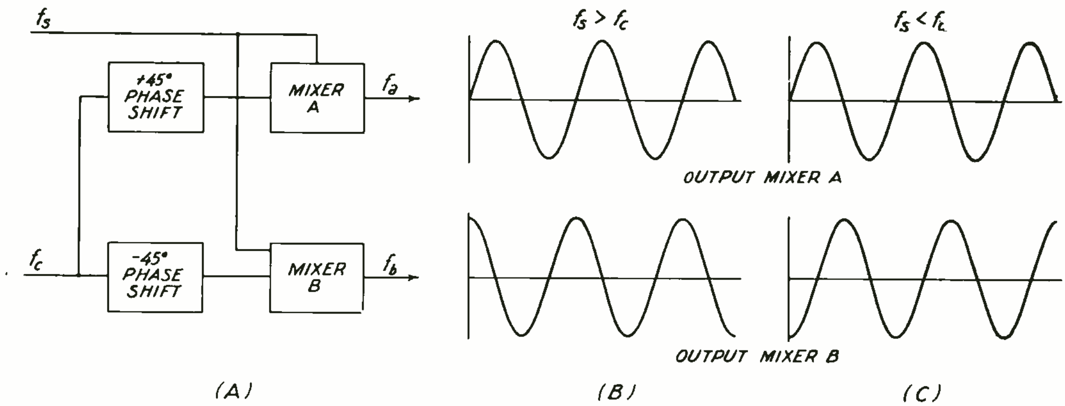



Fig. 30.—Block diagram of mixer circuit for master and crystal oscillators, together with wave shapes obtained at output.

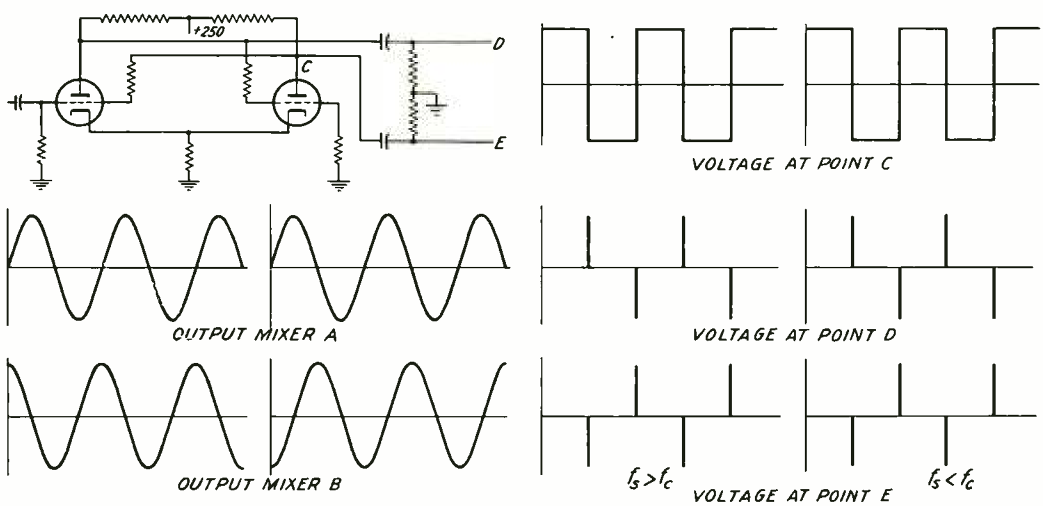



Fig. 31.—Multivibrator action and output voltages.

multivibrator stays in one state until the sine wave reverses, whereupon it triggers and goes into the second half cycle, and so on. The large amplitude of the sine wave insures the reversal taking place at the moment when the sine wave passes through zero.

The result is that the multivibrator generates a square wave output whose reversals occur at the zero points in the sine-wave output of mixer A. Since the phase of the sine wave does not change when $f_s$ decreases from above to below $f_c$, the square wave output at one of the plates (point C) is the same whether $f_s > f_c$ or $f_s < f_c$, as is indicated in Fig. 31. (The square wave output at the other multivibrator plate is of the opposite polarity to that at C.)

The output at points D and E are differentiated square waves or impulses owing to the small series capacitors coupling the plates to these terminals. The square wave at terminal C is converted into the impulses shown for terminal E. Since the other plate developes a square wave of opposite polarity, its differentiated impulses appearing at terminal D are opposite in polarity to those at E.

Note, however, that the impulses, which occur at the moments when both the sinusoidal output of mixer A and the square-wave output of the multivibrator go through zero, occur at the moments when the sinusoidal output of mixer B is at a crest value. Hence, if these impulses are mixed with the output of mixer B, they will combine to give one or the other of the resultant waves shown in Fig. 32.

It will be observed from Fig. 32(A) that if the signal frequency $f_s$ of the master oscillator exceeds that of the reference crystal oscillator, $f_c$, the impulses from D subtract from the sine wave, and the impulses from E add to the sine wave. On the other hand, if $f_s$ is less than $f_c$, the reverse is true, as is indicated in Fig. 32(B).

The outputs of points D and E are readily mixed with the sine-wave output of mixer B and constitute the signals of two channels. Each channel is connected to a biased diode so arranged as to pass the impulses when they "ride" outwardly from the sine-wave crests. In this way the output of one diode contains impulses corresponding in number to the number of cycles by

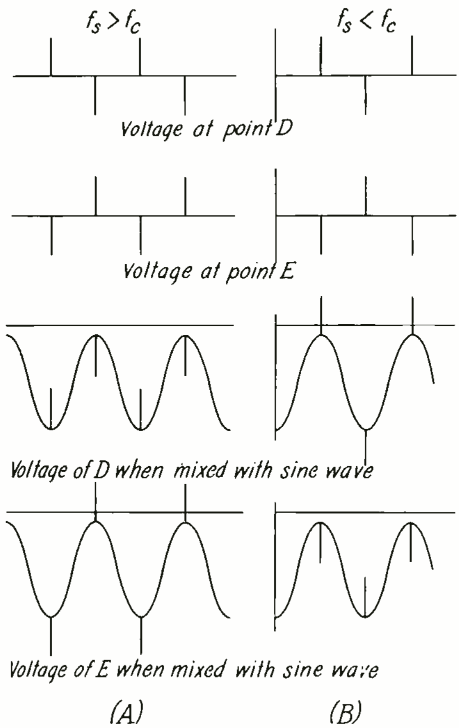



Fig. 32.—Waves shapes resulting from the combination of impulse voltages and sine-wave voltage.

which $f_s$ exceeds $f_c$; and the output of the other diode contains impulses corresponding in number to the number of cycles by which $f_s$ is less than $f_c$. The two sets of impulses can then be added algebraically; if one set exceeds the other in number, a corrective voltage can thereupon be applied to the master oscillator to restore the two numbers to equality.

The combining of the impulses from terminals D and E of Fig. 31, and the sine-wave output of mixer B, together with the biased diode selector effect, is accomplished very simply by the circuit shown in fundamental form in Fig. 33. The impulses from terminals D and E of Fig. 31 are fed to the diode plates through coupling capacitors, while the sinusoidal output of mixer B is

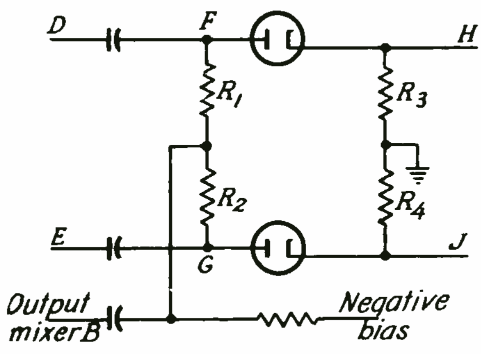



Fig. 33.—**Method of mixing and selecting waves.**

fed to the junction of the two resistors $R_1$ and $R_2$, so that both diodes receive the sinusoidal input, as well as a negative bias, whereas the top diode receives the impulses from terminal D only, and the bottom diode receives the impulses from terminal E only.

The bias is set just greater than the peaks of the sinusoidal output from mixer B. Hence only if the impulses extend outward from these peaks, do they overcome the

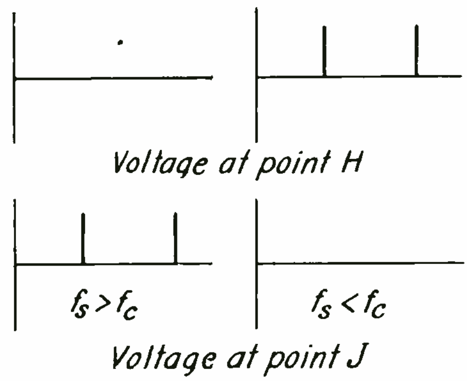



Fig. 34.—**Output voltages of the biased diode discriminators.**

negative bias of the diode plate relative to its cathode and register across the resistors $R_3$ and $R_4$, that is—at terminals H and J.

If $f_s$ exceeds $f_c$, the impulses at terminal D are inward from the sine-wave peaks, and hence do not register at H. On the other hand, the impulses at E are outward from the peaks, and therefore register at J. The reverse is true if $f_s$ is less than $f_c$: the impulses register at D and not at H. This is all clearly shown in Fig. 34.

The impulses are essentially narrow rectangular pulses. They are passed through a combination of diodes into a capacitor so as to charge it first in one direction and then the other. The circuit arrangement is shown in Fig. 35, together with the various wave shapes.

An examination of this figure reveals that capacitor $C_s$ charges when pulses arrive from H, and discharges when pulses arrive from

J. The mechanism of action is indicated by the wave shapes.

Thus, suppose a positive voltage pulse appears at H. This pulls electrons out of the left-hand terminal of capacitor $C_1$, and electrons flow into the right-hand terminal. These electrons cannot come from the top diode; they must come from $C_s$ through the diode M. Thus $C_s$ loses electrons and acquires therefore a positive charge with respect to ground.

When the positive pulse is completed, the right-hand terminal of $C_1$ has an excess of electrons, and no positive pulse voltage to hold them as a bound charge. They therefore overcome the -3 volt bias of the top diode plate and flow through this diode to ground as a reverse current pulse. This leaves $C_1$ in its initial condition ready for the next pulse from H, and this pulse will further charge $C_s$ in the same manner as the preceding pulse did.

On the other hand, when a positive pulse appears at J, it overcomes the +3-volt bias of the bottom diode, and draws electrons from ground into the right-hand terminal of $C_2$ and thereby charges it negative. When the pulse is over, the excess negative charge on $C_2$ is released and passes through diode N into $C_s$. The latter is thereupon reduced in positive charge by the acquisition of these electrons, and $C_2$ returns to its uncharged condition just prior to the appearance of a positive pulse at J. Succeeding pulses at J tend further to reduce the positive charge on $C_s$ and ultimately to charge it negative.

Hence the voltage of $C_s$ rises from zero as pulses appear at H, and drops to zero once again when an equal number of pulses appear at J. Should the number of pulses at H exceed those at J, the capacitor $C_s$ will acquire a net positive

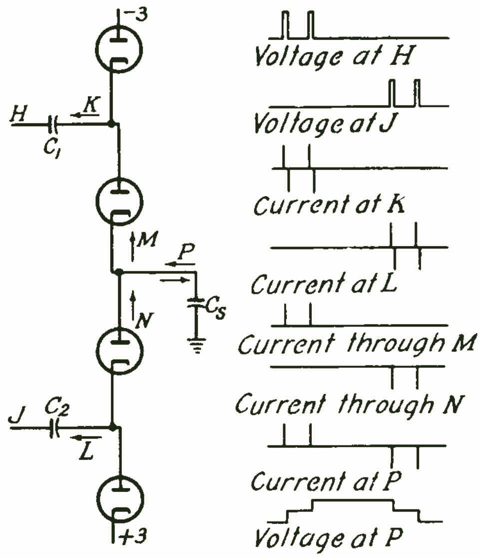



Fig. 35.—Diode arrangement for charging a capacitor with the pulses.

charge; should the number of pulses at J exceed those at H, it will acquire a net negative charge.

The grid of a cathode-follower tube (not shown in Fig. 35) is connected to the top terminal of $C_s$, and developes a d.c. output voltage that varies with the potential on $C_s$. The output voltage is applied to the grid of the modulator control tube shown in Fig. 29 and varies its frequency until it is equal to that of the crystal oscillator, whereupon the number of pulses appearing at H equals the number at J, in Fig. 35, there is no longer any further accumulation of charge of either polarity on $C_s$, and the corrective voltage remains at the value necessary to hold the master oscillator in synchronism with the crystal

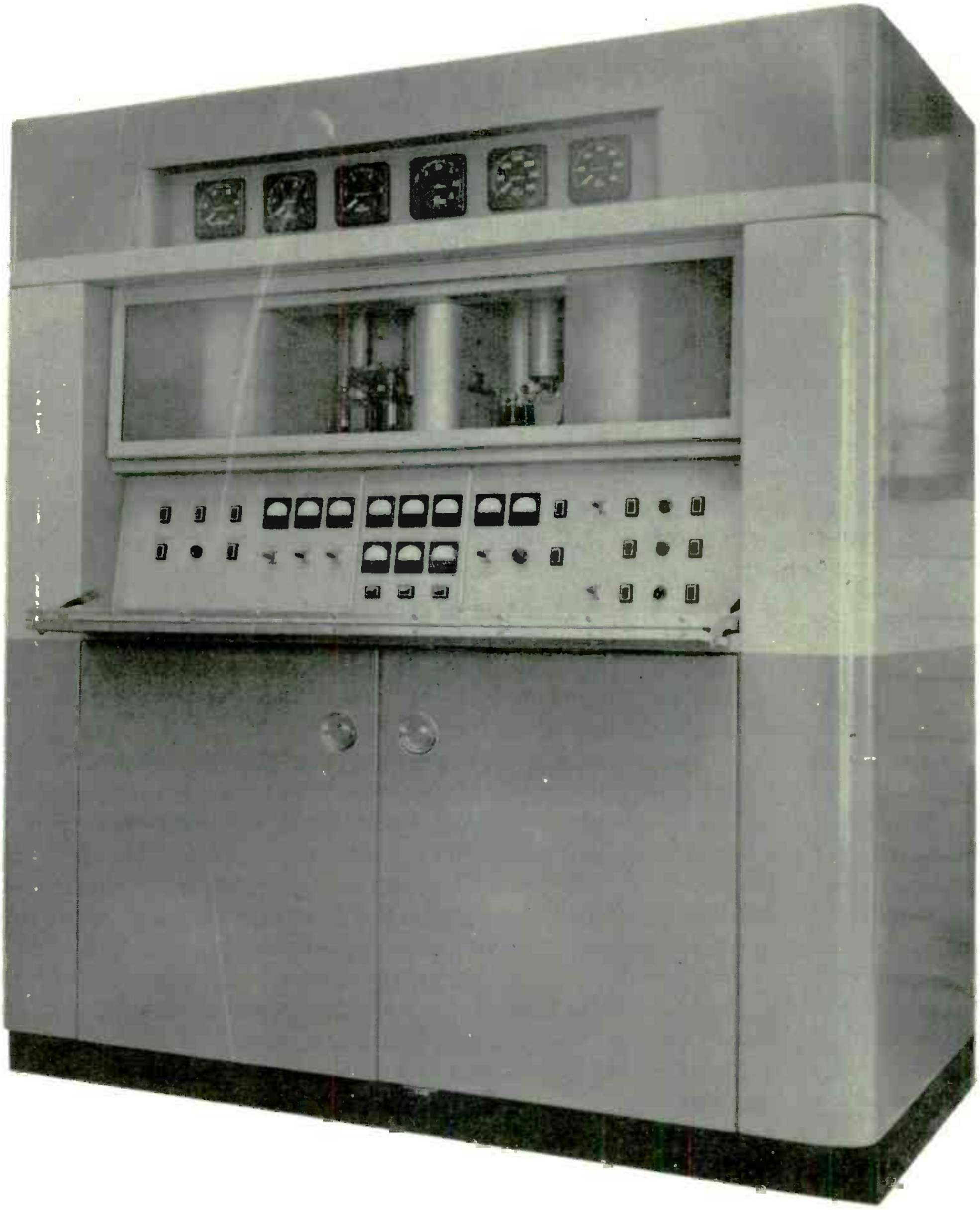


Fig. 36.—Front view of 3-kw frequency-modulated Westinghouse transmitter.

oscillator.

The preceding discussion may lead the reader to assume that the circuit is a very complicated and unreliable. Actually it is not. Although a number of tubes are required, these are operated either as "on" or "off" devices; in short, as switching elements. Hence, any normal variations in their characteristics do not materially affect the action of this frequency stabilizing unit, and tubes that ultimately become weak and defective will be discovered in sufficient time to be discarded during the normal routine check of the equipment.

Another point is that the unit does not contain any tuned circuits or critical components, nor does it employ frequency dividers or locked oscillators which may get out of adjustment. Even line voltage variations should not produce any error in functioning in that a change in line voltage affects the pulse amplitudes and bias voltages to approximately equal degree. Moreover, the power supplies are regulated to preclude any mal-functioning on this account.

*GENERAL DATA.*—In other respects the Westinghouse transmitters are similar to those of other companies. The units come in 1, 3, 10, and 50 kw sizes. The 3 kw is the basic exciter for the 10-kw and 50-kw transmitters. It is illustrated in Fig. 36. The sheet aluminum housing has been designed to be especially roomy, so that cleaning and maintenance can be easily performed around the components in the cabinet. The meters are at eye-level, thus making it more convenient to take their readings. The tubes are also visible through windows, which is an aid in operational procedure.

Vertical open construction is employed, so that it is easy to follow the signal path through the various unit from audio input to antenna output, particularly when rapid servicing during an emergency is required. Moreover, higher-powered units can be readily added, much like building blocks, when it is desired to expand the station facilities.

In Fig. 37 is shown a schematic wiring diagram of the transmitter. The frequency-modulated oscillator and the frequency stabilizer unit have been described previously. An inspection of the complete diagram indicates some features omitted in the simplified diagrams. For example, the two 45° phase-shift networks used to produce two quadrature components for mixing with the master oscillator output, are each R-C combinations.

For mixer A (lower 6SA7 tube) the capacitor is in series with the mixer grid, and the resistor shunts it to ground; for mixer B, the resistor is the series element, and the capacitor shunts the mixer grid (upper 6SA7 tube), although an additional resistor is employed to furnish a d.c. path to ground for the latter grid. In either case, the capacitive reactance equals the resistance; this is the condition for 45° phase shift, leading if the the voltage is taken across the resistor, and lagging if taken across the capacitor.

The mixers are 6SA7 tubes, as stated previously. Grids Nos. 1 and 3 receive the two signals to be mixed or heterodyned, grids Nos. 2 and 4 are connected together and grounded through a by-pass capacitor to act as a shield between the other

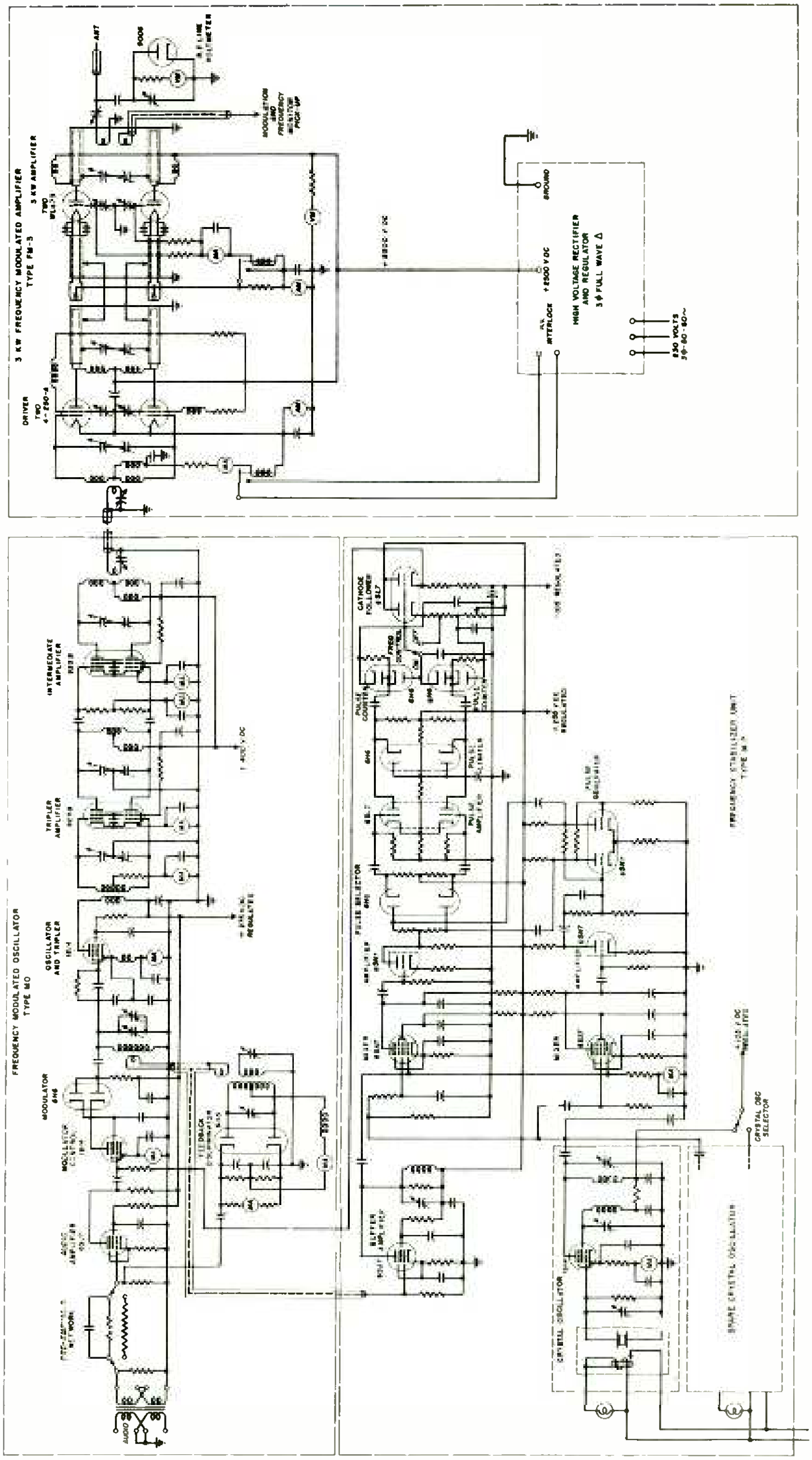


Fig. 37.—Schematic diagram of the f.m.-3 Westinghouse transmitter.

two grids. Mixing therefore occurs in the electron stream just as in the case of a superheterodyne mixer.

Note also the isolating amplifier stages containing 6SN7 tubes between the mixers and the biased diode discriminators, and also the 6SL7 pulse amplifier stages following the diode discriminators. Finally, note that the voltage developed across the pulse counter capacitor ($C_s$ in Fig. 35) is applied to the double cathode-follower stage using a 6SL7 tube. One cathode furnishes the +3 and -3 volts bias (to ground); the other furnishes the correcting voltage to the grid of the 1614 modulator control tube.

The master oscillator is a 1614 tube involving the control and screen grids as the oscillating electrodes. The plate is electron coupled to the oscillator section, and its tank circuit is tuned to three times the frequency of the oscillator, as that tripling occurs immediately. An 829-B tube acts as a second tripler, followed by an 829-B intermediate amplifier.

The latter then excites a driver stage consisting of two 4-250-A tubes in push-pull. The coupling is through tuned coupling loops and a coaxial line. The plate tank circuit consists of trough-line inductors (modified coaxial lines in which the outer sheath is in the form of a trough), tuned by motor-operated capacitors, and shunt fed through r.f. choke coils.

Note that no coupling capacitors are required to the following power amplifier stage; motor-operated taps couple the driver plate inductors to transmission-line elements through which pass the filament leads of the power amplifier stage.

Observe further that the screen grids of the driver stage are fed through resistors and r.f. chokes, and are connected directly to ground through adjustable capacitors. The latter actually form a series resonant circuit with the screen lead inductance, thereby bringing the screens down to ground potential. This, as explained in a previous example, prevents the stage from oscillating by permitting the screens to act as r.f. grounded shields between the control grids and the plates.

The power amplifier is of the grounded-grid type and consists of two WL473 tubes. As indicated previously, the driver stage furnishes some of the output power through this type of amplifier. Observe that here the *control grids* are series tuned by variable capacitors to bring them down to ground potential. The plate inductors are also of the transmission-line type; used in conjunction with motor-operated tuning capacitors.

The output coupling circuit is also motor-driven. A flexible coaxial cable runs from the coupling network to a capacity voltage divider used to excite the r.f. line voltmeter. The latter contains a 9006 tube as the rectifier. A separate coupling unit feeds another line to which is connected the modulation and frequency monitor.

Blowers draw in air at the base, pass it through filters, the power transformers, and power tubes, and then exhaust it through parts at the top of the cubicle. A thorough arrangement of interlocking, as well as overload and under-voltage protection, is employed, and relays and control lamps serve to inform the operator as to just where the trouble is. All main power switches are of the De-ion type (which can very effectively interrupt large currents at high voltages) and have thermal-overload release.

Another feature is the use of motor-driven line voltage regulators on both the auxiliary bus and the high-voltage supply. The regulator on the latter supply can be used to reduce the power output from 3 to 1 kw, whenever this is necessary, as for testing and adjustments.

## CONCLUSION

This concludes the assignment on frequency-modulated transmitters. The designs of various companies were analyzed primarily with respect to four features:

1. Method of frequency modulation.

2. Method of center frequency control.

3. Amplifier data and design.

4. Mechanical construction.

Two of the manufacturer's transmitters (General Electric and Western Electric) have been discussed in a previous assignment. In this assignment the Armstrong dual modulator, the RCA, Raytheon, Federal, and Westinghouse designs have been discussed.

The Armstrong and Raytheon (as well as G.E.) operate on the phase shift principle, with conversion to frequency modulation by the proper shape to the audio response. This method permits the use of a crystal oscillator, so that no center-frequency control is required, although the Armstrong system employs a special method of cancelling out the drift of one of the two crystal oscillators required.

The other manufacturers (including Western Electric) employ direct frequency modulation: a master oscillator is varied in frequency by a reactance tube modulator. This method requires that the frequency-modulated wave be compared with that of a crystal oscillator, and any relative drift be corrected.

The fundamental factor is that of frequency dividing the modulated wave so as to accentuate the carrier relative to the f.m. side bands. The demultiplied carrier can then be compared with the crystal oscillator output, or a submultiple of it. The comparison circuit delivers an output whose frequency is the difference between those of the two waves compared.

The difference beat can then be applied directly as a variable bias to the modulator stage, thereby causing this stage to correct drift as well as to produce frequency modulation; (Federal and Westinghouse systems) or the difference beat can be converted into a two-phase supply and employed to operate an a.c. motor, which is mechanically coupled to a capacitor in the master-oscillator tank circuit (R.C.A., and W.E. systems).

The grounded-grid amplifier appears to be a favorite for power output purposes, because it permits the use of triode tubes without the need for neutralization, and also enables some of the driver power to be transferred to the antenna as well. Thus, a 3-kw transmitter can be readily increased in power output by using the 3-kw power amplifier stage to drive a 10- or even 50-kw power amplifier stage, without the driver power being entirely wasted.

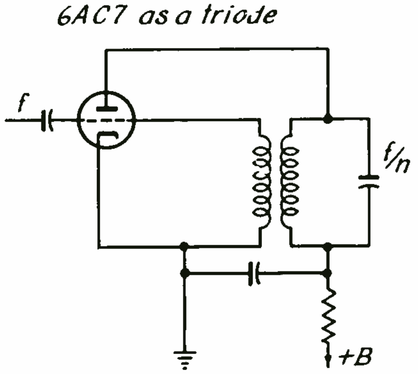
The mechanical design of the transmitters facilitates such expansion, as well as routine maintenance and emergency servicing. The unit type of construction is favored; the addition of space units, such as crystal oscillators, can be readily effected in such a design.

In spite of the variation in electrical and mechanical design features, all makes of transmitters meet the F.C.C. specifications or better, and all are built by reputable manufacturers with a sense of responsibility for their product. Excellent results can therefore be expected from any of the transmitters offered to the broadcasters.

## APPENDIX I

## RCA FREQUENCY DIVIDER

The RCA frequency divider is a different circuit from that employed in the Western Electric f.m. transmitter, which was described in a previous assignment. The RCA circuit is shown in Fig. 1. This




(*Courtesy RCA Review*)
Fig. 1.—Locked-in oscillator frequency divider.

circuit is an oscillator, which oscillates at a frequency f/n, and is locked or synchronized by a frequency f. It therefore delivers an output frequency one-nth of the incoming, synchronizing frequency. Division ratios as high as n = 12 can be obtained, but a value of n = 5 is the maximum used in this equipment.

The action of this circuit* is somewhat similar to the frequency-divider circuit employed in the Western Electric system, and which was described in the assignment on modulation. In the RCA divider, however, output is obtained from the stage regardless of whether or not the higher frequency is impressed; the action of the higher frequency is mainly to lock the lower frequency of the oscillator to a submultiple of its value.

It does this by a process of cross-modulation, very much like the action of the W.E. frequency divider. Suppose the incoming signal (from the reactance tube oscillator) is 4.8 mc, and it is desired to demultiply this frequency by a factor of 3. Then the plate tank circuit in Fig. 1 is tuned to 4.8/3 = 1.6 mc; i.e., n in Fig. 1 equals 3.

The oscillator oscillates over a large portion of its $e_g - i_p$ tube characteristic, including the non-linear curved regions. Hence it normally generates not only a 1.6 mc wave, but harmonics of this, such as 2 × 1.6 mc (second harmonic), 3 × 1.6 mc (third harmonic), and 4 × 1.6 mc (fourth harmonic).

Consider specifically the second harmonic of 2 × 1.6 = 3.2 mc. This component can cross-modulate with the incoming 4.8 mc signal to produce two frequencies:

a. A summation beat frequency of 4.8 + 3.2 = 8.0 mc, and

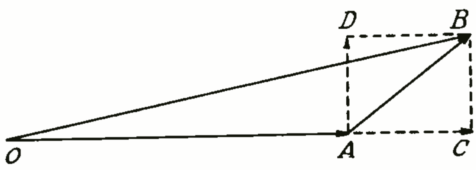
b. A difference beat frequency of 4.8 - 3.2 = 1.6 mc.

Currents of both frequencies flow in the plate circuit. The load, however, is resonant to the difference beat frequency of 1.6 mc, and acts as a by-pass or short-circuit to the 8.0 mc summation component. Hence only the 1.6 mc component *adds* to the normal 1.6 mc current developed by the tube as an oscillator to furnish somewhat increased power. The oscillator adjusts the phase of its output wave to be the

*See also "A Frequency-Dividing Locked-in Oscillator Frequency-Modulation Receiver," G. L. Beers, *Proc. I.R.E.*, Dec. 1944.

same as that of the 1.6 mc cross-modulation product, and stable operation ensues. The oscillator is thus locked to the incoming wave of three times its frequency.

Now suppose the incoming 4.8 mc signal goes up somewhat in frequency. In this case the difference-beat frequency will also increase, and will no longer be the same as the 1.6 mc output of the oscillator. Conditions are now as in Fig. 2.




*(Courtesy Proc. I.R.E.)*

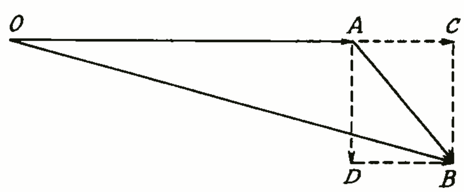
**Fig. 2.—Vector relations in locked oscillator for increase in synchronizing frequency.**

Here OA is the vector representing the 1.6 mc oscillator wave, and AB is a vector representing the somewhat higher difference-beat frequency produced by cross modulation between the second harmonic of the oscillator output and the incoming synchronizing frequency. Since AB is of somewhat higher frequency than OA, it is shown as rotated somewhat counterclockwise (leading) with respect to OA at the instant under consideration.

It can be considered as made up of two components AC and AD at this instant. Component AC is in phase with OA and adds to its output; AD is a leading component, and represents so much additional leading current fed into the oscillator tank circuit. This effect is equivalent to increasing the tank capacity; this in turn causes the oscillator frequency to decrease.

The result is to speed up the counterclockwise rotation of AB relative to OA, so that a moment later the two vectors appear as in Fig. 3. Vector AB has rotated from its position in Fig. 2 so that now it is facing the lower right-hand corner of the page. Component AC is still in phase with OA and therefore adds to its output, but component AD is now lagging, and represents the effect of an additional inductance connected in parallel with the tank circuit elements.

An inductance parallel to the normal tank circuit inductance represents a lower equivalent total inductance, so that now the oscillator increases its frequency. It can now "catch up" with the higher incoming synchronizing frequency, and lock in step with it at one-third its frequency. The beat frequency vector AB then assumes a *fixed* angle with respect to the oscillator frequency vector OA such that the apparent effective inductance shunting the tank circuit maintains the oscillator at the re-




*(Courtesy Proc. I.R.E.)*

**Fig. 3.—Vector relations in locked oscillator at a moment later than that shown in Fig. 2.**

quired higher frequency.

The above analysis has been based on the cross-modulation between the second harmonic of the oscillator current and the incoming synchronizing frequency which is three times the fundamental oscillator frequency. The same action can be obtained by the interaction of the fourth harmonic of the oscillator frequency and the incoming frequency, or any suitable combination of harmonics of the incoming and oscillator frequencies that yields by cross modulation the fundamental oscillator frequency.

The simultaneous interaction of the second and fourth harmonics of the oscillator current with the fundamental component of the incoming frequency (if the latter is not sinusoidal) is to produce a beat-frequency component that is variable both in amplitude and phase. The synchronizing mechanism is essentially the same as for a single harmonic component, particularly if one harmonic component is stronger than the other, as is usually the case.

Thus, the oscillator is capable of locking to an incoming frequency and furnishing an output frequency that is a submultiple of the incoming frequency. Moreover, the output is sinusoidal in shape, and hence does not require broad-band output circuits. It is to be noted that this circuit continues to furnish output even when the incoming frequency fails, so that there is no direct indication of such failure. How important this is from an operational viewpoint is a matter that will be decided as time goes on; one very important practical advantage of the RCA frequency divider is its relative simplicity.

## APPENDIX II

## ANALYSIS OF THE MILLER EFFECT

In Fig. 1 is shown an ordinary vacuum tube amplifier circuit. The plate load is $Z_L$; the output voltage across $Z_L$ is $\alpha$ times the input voltage $e_i$, where $\alpha$ is the stage gain. From an earlier assignment it will be recalled that

$$\alpha = \mu \frac{Z_L}{R_p + Z_L} \qquad (1)$$

where $\mu$ is the amplification factor of the tube, and $R_p$ is its plate resistance. Note that $Z_L$ may be a complex impedance, in which case $\alpha$ is complex. This means that the input voltage is both amplified and shifted in phase.

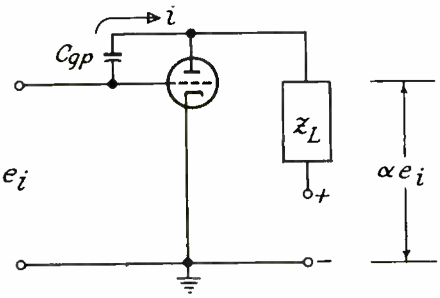



Fig. 1.—Circuit illustrating the Miller effect.

Owing to the inter-electrode capacitance $C_{gp}$ between the plate and control grid, which capacitance can be enhanced by the addition of external capacitance between the two electrodes, the output voltage $\alpha\ e_i$ and the input voltage $e_i$ act vectorially to produce a current i through $C_{gp}$, which represents current coming from or flowing into the source of $e_i$, depending upon the phase of $\alpha\ e_i$ and $e_i$.

More specifically, the voltage across $C_{gp}$ is

$$e_i + \alpha\ e_i = e_i(i + \alpha) \qquad (2)$$

The current i that flows is given by Ohm's law for a.c., namely,

$$i = e_i(1 + \alpha)(j\omega C_{gp}) \qquad (3)$$

If the value for $\alpha$ from Eq. (1) be substituted in Eq. (3), there is obtained

$$i = e_i \left[1 + \mu \frac{Z_L}{R_p + Z_L}\right]\left[j\omega C_{gp}\right] \qquad (4)$$

The apparent admittance looking into the input terminals is

$$A_i = i/e_i \qquad (5)$$

$$= j\omega C_{gp} + [\mu Z_L/(R_p + Z_L)][j\omega C_{gp}]$$

The first term on the right, $j\omega C_{gp}$, represents the susceptance (reciprocal of reactance) of a capacitance $C_{gp}$.

The second right-hand term represents an admittance which may or may not be capacitive, depending upon the nature of $Z_L$. Suppose $Z_L$ is a pure resistance $R_L$. Then the second-right-hand term becomes

$$j\omega \frac{\mu R_L C_{gp}}{R_p + R_L}$$

which is the capacitive susceptance of a capacity of magnitude

$$\frac{\mu R_L C_{gp}}{R_p + R_L}$$

since $R_L$ is a real number, so that the entire expression

$$\frac{\mu R_L C_{gp}}{R_p + R_L}$$

is a real number, and represents

therefore a pure capacitance.

Hence the admittance looking into the grid terminals is that of two capacitances in parallel: $C_{gp}$ and $\mu R_L C_{gp}/(R_p + R_L)$. Note that owing to the amplifying properties of the tube, $C_{gp}$ between the grid and plate looks like the much larger capacitance between the grid and cathode (ground). In this case the feedback effect from the plate to the grid is to draw a relatively large leading current which makes the input source think it is feeding a comparatively large capacitor connected across its terminals.

For a pentode tube $R_L$ is generally much less than $R_p$, so that Eq. (5) reduces to

$$A_1 \cong j\omega C_{gp} + \left(\frac{\mu Z_L}{R_p}\right)\left(j\omega C_{gp}\right) \tag{5}$$

$$= j\omega C_{gp} + (G_m Z_L)(j\omega C_{gp})$$

where $\mu/R_p = G_m$, the transconductance of the tube.

For a resistive load $R_L$ the added capacitance is $G_m R_L C_{gp}$ as is indicated by Eq. (10) in the text. This indicates that if it is desired to have the added capacitance vary in direct proportion to the applied modulating voltage, then the tube must be so operated on its characteristic curves as to have its $G_m$ vary in direct proportion to the modulating voltage.

On the other hand, suppose $Z_L$ is an inductive reactance $j\omega L$. Then

$$A_1 = j\omega C_{gp} + j\omega \frac{\mu Z_L}{R_p + Z_L} C_{gp}$$

$$= j\omega C_{gp} + j\omega\left(\frac{\mu j\omega L}{R_p + j\omega L}\right)C_{gp}$$

$$= j\omega C_{gp} + \frac{-\omega^2\mu L C_{gp}(R_p - j\omega L)}{(R_p + j\omega L)(R_p - j\omega L)}$$

(multiplying numberator and denominator by $R_p - j\omega L$.)

$$= j\omega C_{gp} + \frac{-\omega^2\mu L C_{gp} R_p}{R_p^2 + (\omega L)^2} + \frac{j\omega^3 L\mu C_{gp}}{R_p^2 + (\omega L)^2}$$

(multiplying out all factors and separating the terms in the numerator.)

Set

$$\frac{\omega^2\mu L C_{gp} R_p}{R_p^2 + (\omega L)^2} = G'$$

and

$$\frac{\omega^2 L\mu C_{gp}}{R_p^2 + (\omega L)^2} = C' \tag{6}$$

Then (6) becomes

$$A_1 = -G' + j\omega C' + j\omega C_{gp} \tag{7}$$

where $-G'$ is a *negative* conductance, shunted by $C'$ and $C_{gp}$. The negative sign in front of $G'$ indicates that it is a source of power to the input circuit. This power comes from the plate circuit; it represents *regenerative* feedback. It is this factor that causes oscillation of the circuit, provided $G'$ is greater than the positive conductance associated with the input circuit. Such regenerative feedback is normally compensated by neutralizing circuits to prevent oscillation.

If $Z_L$ is a net capacitive reactance, then a similar analysis yields

$$A_1 = G'' + j\omega C'' + j\omega C_{gp} \tag{8}$$

where

$$G'' = \frac{\mu\omega^2 C C_{gp} R_p}{1 + \omega^2 C^2 R_p^2}$$

and

$$C'' = \frac{\mu C_{gp}}{1 + \omega^2 C^2 R_p^2}$$

It is to be noted that for this type of plate load, the input admittance consists of a *positive* conductance paralleled by two capacitances. Thus the input circuit will be damped by the positive conductance; i.e. degenerative instead of regenerative feedback is taking place.

## *FREQUENCY-MODULATED TRANSMITTERS*

### EXAMINATION

1. What are the two basic systems of frequency modulation?

2. (A) Which system is inherently crystal-controlled?

   (B) In the other system, how is carrier frequency drift controlled?

3. Refer to 2(B). What fundamental principle is employed before such control is feasible? Describe briefly in terms of the fundamental principles of a frequency-modulated wave.

4. In the Armstrong dual modulator system, the frequency-modulated wave has a drift of +50 c.p.s. in 15.6 mc and a frequency deviation of +1.25 kc in channel A, and -1.25 kc in channel B. It is then heterodyned with another crystal controlled oscillator having a frequency of 2.672 mc and a drift of +30 c.p.s.

   (A) What is the beat frequency obtained when channel A is beat with the heterodyne oscillator, also the deviation, and the total drift?

   (B) What are the beat frequency, deviation, and total drift?

   (C) The resulting second beat frequency is then multiplied 36 times. What are the final carrier frequency, deviation, and carrier frequency drift?

5, (A) In the RCA transmitter, how is carrier-frequency drift controlled? Explain briefly.

   (B) What mechanical feature permits the motor to respond to almost zero beat frequency?

6. (A) In this same transmitter, how is hunting of the carrier-frequency control unit prevented?

EXAMINATION, Page 2.

6. (B) Assuming that only triode tubes are available in the larger power sizes, how is neutralization of the power amplifier stage avoided?

   (C) What further advantage is obtained from this method?

7. (A) How does Raytheon obtain *large* initial phase shifts?

   (B) What two advantages are obtained by such large initial phase shifts?

8. (A) What principle is employed in obtaining frequency modulation in the Federal transmitter?

   (B) Describe briefly how carrier-frequency drift is controlled?

9. (A) Describe briefly the method employed by Westinghouse in obtaining frequency modulation.

   (B) What advantage is claimed for this method?

10. Describe briefly in block diagram form how carrier-frequency drift is controlled in the Westinghouse transmitter.