CREI

SECTION 2

ADVANCED
PRACTICAL
RADIO ENGINEERING

MODULATION

242F

FOREWORD

A writer in one of the humorous magazines once re-
marked that he never questioned the source of the water when
he turned on a faucet; he took its presence for granted just
like the law of gravity, or one and one are two.

Most people are beginning to accept radio programs
in the same frame of mind; they are in the air, and all you
have to do is turn on your receiver and presto! the program
can be heard. This is a desirable state of affairs from the
entertainer's point of view: it is a cardinal maxim in the
entertainment field that the mechanism behind the entertain-
ment should never intrude itself upon the consciousness of
the spectator; he should be aware only of the play itself.

To the engineer, however, the mechanism is of
prime importance, and what goes on behind the scenes, at
least technically, is of fundamental interest to him. This
assignment takes you behind the scenes in a broadcast sta-
tion, and shows you how the audio signal is impressed upon
the radio-frequency carrier; in short—modulation.

The methods of obtaining amplitude modulation are
discussed first. The most common method is plate modula-
tion, and the text analyzes the theory and design of a modu-
lator stage. Next, grid modulation is discussed, and the
method of minimizing distortion is developed so that a suit-
able design procedure may be followed. Other methods of
modulation are discussed more briefly, as they are not used
as often as the first two methods.

The second half of the assignment deals with fre-
quency modulation. The main purpose of this text is to de-
velop the theory and practice of f.m. Hence, three repre-
sentative methods are described: the Armstrong system; the
General Electric phasitron-tube method; and the Western
Electric system. The principles involved are thoroughly

discussed so that the student will have a basic understanding of this method of modulation, with representative systems analyzed as examples of applications of the methods. It is left for the Specialized Broadcast Section to describe in more detail the equipment, as well as methods employed by other manufacturers. The present assignment therefore lays the fundamental groundwork for a proper understanding of modulation.

E. H. Rietzke,
President.

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MODULATION

GENERAL CONSIDERATIONS

The Institute of Radio Engineers, in their "Standards on Transmitters and Antennas," define modulation as "the process of producing a wave some characteristic of which varies as a function of the instantaneous value of another wave, called the modulating wave." The concept of modulation has also been discussed in other assignments. In practical systems in use today, either the amplitude or the frequency of the given wave is altered in accordance with the instantaneous value of the modulating wave, thereby giving rise either to an amplitude- or frequency-modulated wave.

This assignment will concern itself with the practical methods of modulating a wave with regard to amplitude and frequency, and specific circuits for accomplishing either type of modulation will be analyzed. Further discussion will be found in the Specialized Broadcast Section.

AMPLITUDE MODULATION

FUNDAMENTAL PRINCIPLES. — In varying the amplitude of a carrier wave, several methods may be employed:

1. The plate voltage of the amplifier tube may be varied at the modulation frequency.

2. The grid voltage (bias) may be varied at the modulation frequency.

3. Any other electrode, such as the screen grid, suppressor grid, or even the cathode, may be varied in potential at the modulation frequency.

4. A combination of such methods may be employed to afford greater linearity (proportionality) between the amplitude of the carrier wave and the modulating voltage.

As a general survey of these methods, it may be noted that plate modulation is in general more efficient—at least as regards the operation of the modulated tube, but requires a high-power modulator stage. It is also possibly the most linear of all methods, but in general requires a matching modulation transformer between the two tubes, and such a transformer, especially for high-power purposes, has a narrow band width, suited mainly for audio frequencies.

On the other hand, grid modulation requires considerably less modulation power, although the efficiency of the modulated tube is in general much lower. Nevertheless, for wide-band modulation, as in television, grid modulation is a necessity because wide-band high-power plate transformers are not as yet available, and recourse must therefore be had to grid modulation, which does not require a modulation transformer. This is discussed in the Specialized Television Section.

Screen-grid and other types of modulation are employed occassionally for special purposes, and in general have advantages and disadvantages intermediate between grid and plate modulation. They seldom, therefore, warrant

consideration in preference to the above two types.

Another consideration in amplitude modulation is whether the process should occur in an early stage of the r-f chain or in the final stage. If modulation is performed in the final stage, more modulation power is required, particularly in the case of plate modulation, but the preceding stages can be of simpler form, and therefore less critical of adjustment, and the final stage can be of the efficient Class C type. This process is known as "high-level modulation."

On the other hand, a low-power early r-f stage can be modulated, and the resulting amplitude-modulated r-f wave can then be amplified by so-called "linear" amplifiers of the Class B type. This is known as "low-level modulation." Unfortunately linear amplifiers are not as efficient as the Class C type of amplifiers that can be employed in high-level modulation systems. On the other hand, the audio (modulating) power requirements are far less, and this tends to compensate for the inefficiency of the linear amplifiers.

A compromise is possible between high- and low-level modulation. In the Doherty amplifier, developed by Doherty of the Bell Telephone Laboratories, a special design permits a linear amplifier to be built of high efficiency. This amplifier is described in detail in the Specialized Broadcast Section; suffice it to say here that low-level grid modulation is used in the r-f stage immediately preceding the final power amplifier.

In one model of such a transmitter the crystal oscillator is followed by three stages of r-f amplification, the third stage employing a Type 241-B (200-watt) tube. This stage in turn drives the modulated amplifier consisting of two Type 232-B (water-cooled, 25-kw) tubes in parallel. The modulated amplifier is grid-modulated by two Type 241-B tubes connected in parallel and operated Class A. The modulated amplifier is followed by the final Doherty power amplifier which employs two Type 298-A (100-kw) tubes. It should be noted that although relatively large tubes are used in the modulated amplifier (two 232-B's in parallel) 100 per cent grid modulation is effected by the Class A output of the 241-B tubes. Inverse feedback enables all tubes to be operated at good efficiency without sacrifice of audio quality.

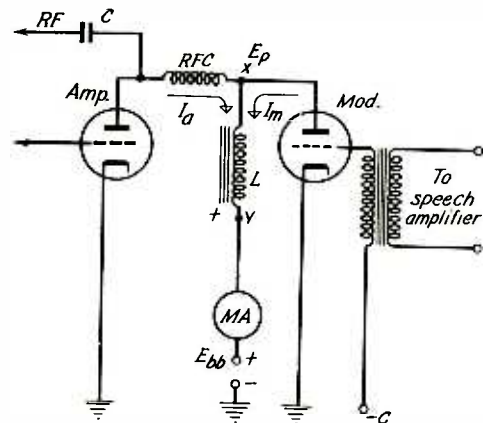*PLATE MODULATION.* — An early form of plate modulator is shown

Fig. 1. — Heising method of choke modulation.

in Fig. 1. It is known as the "constant current" or Heising system of

plate modulation after its in-
ventor. Audio choke coil L has
sufficient inductance to oppose
any appreciable change by the modu-
lator tube in the current flowing
through it even at the lowest audio
frequency. Thus, when a modulating
signal voltage is applied to the
grid of the modulator tube, and
the latter thereupon changes its
plate current, an equal and op-
posite change is produced in the
amplifier tube's plate current,
so that the sum of the two plate
currents, flowing through L, is
substantially constant.

Specifically, if the modulator
grid swings negative, the modulator
plate current I_m tends to decrease.
If the inductance of L is suffi-
ciently high, and the modulator
grid voltage and corresponding
plate current changes are not too
slow, then the actual decrease in
the choke current will induce a
voltage across L in the direction
making terminal x more positive
to terminal 6. This is equivalent
to raising the common plate voltage
E_b to a corresponding higher value
above the plate supply voltage E_{bb}.

As a result, the plate current
of the amplifier tube I_a rises to a
point nearly compensating for the
decrease in modulator current I_m,
so that the sum of the two (I_a +
I_m), which flows through L, is
nearly equal to its previous value.
The actual decrease in (I_a + I_m)
is just sufficient to induce across
L the necessary voltage to raise
the amplifier current from its
initial value to the amount re-
quired; if L is sufficiently large,
the decrease in the current (I_a +
I_m) through L will be very small.

When the modulator grid swings
positive, the modulator current

I_m increases, thereby inducing
across L an opposite voltage such
that x drops below y in potential.
This means that E_b is now less
than E_{bb} by the voltage drop across
L, and as a result I_a decreases
until (I_a + I_m) through L is just
a little greater than before;
sufficient to induce the voltage
drop across L. Thus the current
through L is practically constant;
its small variations are sufficient
to develop high induced voltages
across L because the latter's in-
ductance is so great.

The choke L can be regarded
as a 1 : 1 ratio autotransformer,
that injects a modulating voltage
E_m in series with the plate supply
voltage E_{bb} for energizing the
amplifier tube. This equivalent
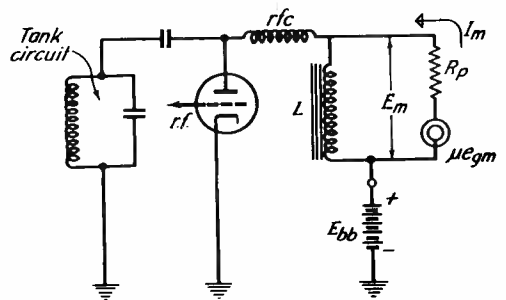circuit viewpoint is illustrated

Fig. 2. — Equivalent circuit for
Heising modulator.

in Fig. 2. The modulator tube oper-
ates as a power amplifier stage, with
the modulated amplifier tube acting
as the load resistance, and the choke
L as the means of feeding the d-c
component directly to the modu-
lator tube. Hence, the principles
of audio power amplification, as
given in a preceding assignment,
may be employed here, once the
action of the amplifier tube as a

load resistor has been analyzed.

THE AMPLIFIER TUBE AS A LOAD RESISTOR. — In order to obtain high efficiency, the amplifier tube is operated Class C. This means that the grid bias is sufficiently far beyond cutoff so that plate current flows only at the peaks of the r-f exciting voltage, and hence is in the form of narrow pulses. These are similar to those that flow in a diode detector circuit, and a similar analysis is possible.

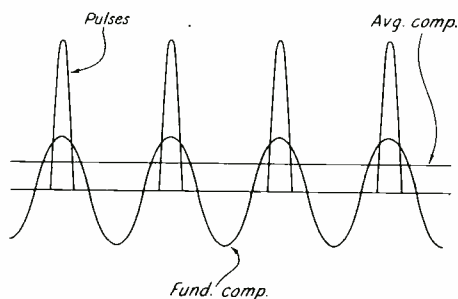In Fig. 3 are shown the pulses for the case where the plate voltage is constant at the value E_{bb};

Fig. 3. — Resolution of unmodulated plate current pulses into various components.

i.e., no modulation is taking place, and no modulation voltage E_m is being produced across the choke. The pulses are then all of equal height.

They can be resolved by the Fourier method into a series of components, two of which are shown in the figure. One is a sinusoidal fundamental component I_{rf} of radio frequency, the other is the average component of the wave, and in this particular case is direct current I_{dc}. In addition, there are a large number of harmonics,

but these are of no interest here. For such narrow pulses, the average or d-c component is very closely equal to one-half the amplitude of the fundamental component i.e.,

$$I_{dc} = \frac{1}{2} I_{rf} \qquad (1)$$

The d-c component flows through the power supply, causing the latter to deliver power to the amplifier tube of magnitude

$$W_a = I_{dc} E_{bb} \qquad (2)$$

This power is expended partly as plate dissipation in the amplifier tube, and partly as r-f power in the tank circuit and coupled load, if any. So far as the power supply is concerned, the expenditure of power w_a makes the r-f amplifier circuit appear as a resistance

$$R_{dc} = E_{bb}/I_{dc} \qquad (3)$$

since such a value of resistance would exactly produce the power W_a of Eq. (2). Thus, in spite of the fact that the amplifier grid makes the tube act as a variable resistance, which chops up the d-c current into r-f pulses and thereby produces r-f energy in the tank circuit, the amplifier tube and circuit appears to the power supply as a resistance of the value given by Eq. (3).

Usually I_{rf} is by-passed around the power supply by a by-pass capacitor. However, even if it is permitted to flow through the power supply, the product $E_{bb} I_{rf}$ has both positive and negative values with time as I_{rf} alternates, and the *average* power is zero. Hence, I_{rf} produces no resistive reaction on the power supply; as stated pre-

viously, the r-f amplifier circuit appears as a resistance R_{dc} to the power supply; R_{dc} is determined solely by the ratio E_{bb}/I_{dc}, and I_{rf} does not enter into the evaluation of this resistance.

Suppose the plate supply voltage were increased by a factor m, and I_{rf} increased by the same factor, as did also I_{dc}. Then the ratio $R_{dc} = E_{bb}/I_{dc}$ would remain unchanged. Such linearity (proportionality) between E_{bb}, I_{rf}, and I_{dc} can be obtained by proper adjustments of the amplifier circuit. For example, if the bias is adjusted to about twice cutoff, and the r-f grid drive is sufficient to saturate the plate current (somewhat flatten the peaks) for all values of E_{bb}, then the above linearity will be approximated.

The same effect can be obtained if instead of directly varying E_{bb}, a modulating voltage E_m is injected in series with it by energizing the modulator grid. For a sinusoidal modulating voltage, the pulses take on the variable amplitude shown in Fig. 4. In (A) are shown the pulses,

Fig. 4. — Current pulses and important components of modulated amplifier.

whose tips lie along the dotted-line sinusoidal envelope. In (B) is shown the r-f fundamental component I_{rf} of this wave; it will be observed that it is now variable in amplitude, that is—amplitude modulated.

It therefore consists of the original carrier component I_{rf}, plus two side band components: one of a frequency higher than the r-f grid input by the modulation frequency, and the other lower by the modulating frequency. (This was derived in a previous assignment.)

The average component is shown in (C). It is now observed to be variable in amplitude, and can be resolved into the original d-c component I_{dc}, plus an a-c component I_m of modulating frequency. From what has been said previously, the ratio E_m/I_m is approximately the same as E_{bb}/I_{dc}, and represents the same value of resistance.

However, since I_m is an a-c component, it absorbs no average power from the plate supply any more than does I_{rf}, and hence does not present a resistance effect to the power supply. On the other hand, the product $I_m E_m$ does represent power, because I_m is directly proportional to E_m and therefore in phase with it. This means that I_m draws power from the modulator tube, and hence causes the amplifier stage to appear to the modulator tube as a resistance of value

$$R_m = E_m/I_m = E_{bb}/I_{dc} = R_{dc} \quad (4)$$

In short, the amplifier stage appears as a resistance of about the same value to the modulator stage as it does to the power supply. As a consequence, if I_{dc} drawn from the power supply, and E_{bb}—its voltage—
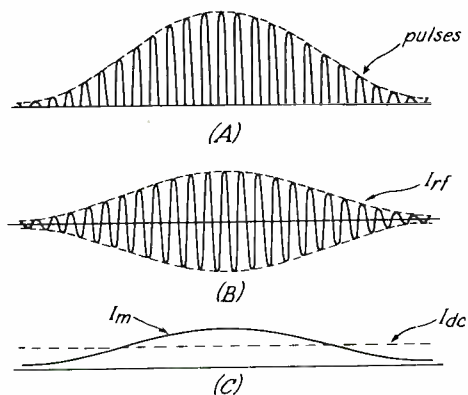
are each measured with a d-c meter, and the quotient taken, the result is essentially the resistance that the amplifier stage presents to the modulator tube.

As an example, a 203-A tube requires a power supply potential of 1,000 volts, and under normal operation as a Class C amplifier, draws 0.15 amperes d.c. It therefore appears as a resistance of

$$R = 1,000/.15 = 6,667 \text{ ohms}$$

both to the power supply and to a plate modulator tube. If a load line having a slope of 6,667 ohms is drawn on the plate characteristics of the modulator tube, a check can be made to see if the modulator tube can furnish the requisite power.

However, for 100 per cent modulation, the plate voltage of the amplifier tube must vary from twice normal to zero. This means that the peak modulation voltage E_m must be equal to E_{bb}. Since E_m is obtained from the modulator tube whose supply voltage is also E_{bb}, it is necessary for the modulator tube to furnish an output a-c voltage E_m equal to its supply voltage.

This is not possible, since E_{min} can never be quite zero, especially if the grid is not driven positive, as in a Class A amplifier tube. Hence, it is necessary for the modulator supply voltage to exceed that of the amplifier tube, and this is most simply done by inserting a by-passed resistance in series with the amplifier tank circuit, as shown in Fig. 5.

Here C_1 and L_1 represent the r-f tank circuit, and C_2 represents a blocking capacitor. Resistor R lowers the d-c voltage to the amplifier tube, whereas the modu-

lator tube receives the full plate-supply voltage E_{bb}. Capacitor C_3 is sufficiently large to by-pass R at
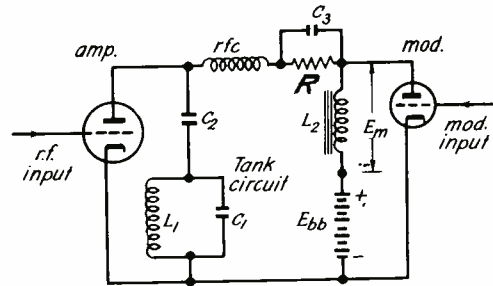

Fig. 5.—Method of reducing the plate voltage to the amplifier tube.

the modulating frequencies, permitting the full magnitude of E_m to act on the amplifier plate.

Suppose the minimum plate voltage of the modulator tube (at the peak positive swing of the modulator grid) is one-third the modulator supply voltage E_{bb}. Then the peak audio output voltage of the modulator is $(1 - 1/8)E_{bb} = 2/3\ E_{bb}$. The supply voltage of the r-f stage must then be $2/3\ E_{bb}$ in order that the modulator tube vary the plate voltage of the r-f stage from zero to $(2 \times 2/3)E_{bb} = 4/3\ E_{bb}$ in order to produce 100 per cent plate modulation. The resistor R must therefore absorb $1/3\ E_{bb}$ when the d-c component of the r-f plate current flows through it.

As an example, suppose the modulator supply voltage is 1,500 volts. Then the r-f supply voltage should be $2/3 \times 1,500 = 1,000$ volts. Suppose the r-f stage draws 100 ma d.c. when fully excited. Then

$$R = \frac{500}{0.1} = 5,000 \text{ ohms}$$

The capacitor C_3 is determined by more complex considerations. As a rough and conservative estimate, it should be large enough so that its reactance at the lowest audio frequency will be, say, one-half of R for a fairly flat response. Thus, in the example cited, for a flat response down to 50 c.p.s.,

$$\frac{1}{2\pi 50\ C_3} = \frac{5000}{2}$$

or

$$C_3 = \frac{2}{2\pi 50 \times 5000} = 1.274\ \mu f$$

Values from 1 to 8 μf are common; actually, if the resistance that the r-f stage presents to the modulator tube, plus the R_p of the modulator tube itself, is large compared to R, the response will be flat even if C_3 is rather small. This follows from a more exact analysis of the circuit.

This method is simple and permits 100 per cent modulation, but is wasteful of d-c power, although capacitors C_2 and C_3 prevent R from wasting r-f and modulation power respectively. Of course, the lowering of the d-c potential by R to the amplifier plate indirectly reduces the r-f power output of the amplifier tube, but it does not do so by directly absorbing r-f power, which is expensive, and it also reduces the amplifier plate dissipation as well as the r-f output.

USE OF A MODULATION TRANS-FORMER. — The necessity for employing R for 100 per cent modulation is one of the bad features of the original Heising circuit. Another is that the resistance presented by the amplifier tube to the modulator tube (or tubes) may not be the optimum value to obtain so-called

maximum undistorted power output from the latter.

Another factor of great practical importance is that the original Heising circuit does not permit the use of two modulator tubes in Class B push-pull operation, a mode that permits considerably more output from the tubes. This is because the Heising choke is essentially a 1 : 1 autotransformer, and as such does not afford any latitude in impedance matching.

This can be obviated by employing a tap on the choke, thereby making it other than a 1 : 1 auto-transformer and thus affording impedance matching by suitable choice of the tap. Another variation is to employ a two-winding transformer; this permits not only different supply voltages to be applied to the two tubes, if desired, but also permits two modulator tubes in push-pull to be coupled to a single-ended Class C amplifier tube.

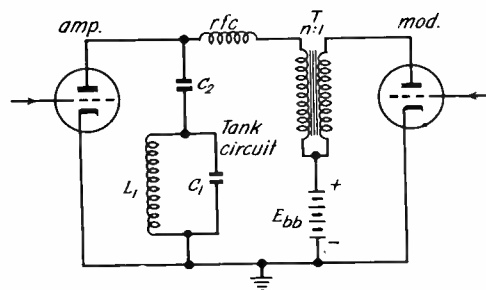The use of a two-winding transformer T is shown in Fig. 6. It has

Fig. 6. — Use of a plate modulation transformer.

a turns ratio of n : 1; this may be step-up or step-down, as determined by the characteristics of the two

tubes. An example will make this clear.

Suppose the amplifier stage employs a 203-A tube operating at 1,000 volts, and drawing 150 ma. It appears (as calculated previously) as a 6,667-ohm resistor to the modulator tube. From a Tube Manual it is found to furnish 100 watts carrier (unmodulated) power. Its d-c input is $1,000 \times .15 = 150$ watts, so that the plate dissipation is $150 - 100 = 50$ watts. This is well within the 100 watts maximum permissible. When modulated, however, the plate dissipation will rise closer to the limiting value.

The modulation voltage E_m must have a peak value equal to the supply voltage, or 1,000 volts. The required modulation power is

$$W_m = E_m^2/2R \qquad (5)$$

The factor 2 comes about from the fact that E_m is the peak rather than the r.m.s. value.

$$W_m = (1,000)^2/2 \times 6,667 = 75 \text{ watts}$$

This is the power to be supplied by the modulator tube or tubes; it is to be noted that for 100 per cent modulation, the modulator furnishes 50 per cent of the carrier power to the side bands.

An inspection of a Tube Manual shows that an 845 tube, operating Class A, at 1,000 volts on the plate, will furnish an output of 24.4 watts into a 6,667-ohm load. Hence, three such tubes in parallel will just about furnish the required power. However, if these three tubes fed the 203-A tube directly, as in a Heising modulator, they would share the amplifier load equally, and the latter would appear to each tube as

$3 \times 6,667 = 20,000$ ohms. Under these conditions a graphical construction indicates that the power output of each will be but 12.3 watts, which is inadequate.

By the use of a matching modulation transformer, the load impedance can be adjusted to appear as the optimum value of 6,667 to each tube, whereupon the required output power will be obtained. A step-down ratio is required, such that the 6,667-ohm equivalent amplifier resistance will appear as

$$6,667/3 = 2,222 \text{ ohms}$$

to the three 845 tubes in parallel, and hence as $3 \times 2,222 \overset{\sim}{=} 6,667$ ohms to each tube.

If the impedance ratio is 3 : 1, then the turns ratio should be $\sqrt{3} : 1 = 1.732 : 1$, or n should equal 1.732 in Fig. 6. A graphical construction on the 845 tube characteristics then indicates that the power output will be 24.4 watts per tube, or $3 \times 24.4 = 73.2$ watts total, and the peak output voltage will be 580 volts at the plates of the 845 tubes. This will be stepped up to $1.732 \times 580 = 1,005$ volts, which is sufficient to swing the amplifier plate voltage to zero.

Actually the voltage should not swing below the peak grid voltage. For a 203-A tube the grid bias is -135 volts, and the peak r-f grid swing is 275 volts. Hence the grid goes positive at the peak of the swing by an amount $275 - 135 = 140$ volts. The minimum plate voltage should be greater than this, say about 180 volts, in order to prevent the grid from drawing excessive current at the positive peak of its driving cycle and thus overheating and also requiring excessive driving

power. Further protection against excessive grid current is to provide a grid leak as a partial source of bias; any tendency of the grid to draw excessive current results in a greater bias drop across the grid leak, so that the grid thus pushes itself more negative and thereby holds down its grid current. Also, as will be shown farther on, the grid excitation can be modulated to prevent excessive grid current.

However, for the purposes of a conservative design, the modulating voltage E_m can be taken as equal to E_{bb} in order to calculate the modulation power, transformer turns ratio, etc., and then the modulation input adjusted in actual operation so as to meet the above conditions of reasonable r-f driving power for the grid of the amplifier tube.

CLASS B MODULATOR. — A modulator tube operated Class A furnishes relatively little power output. This is particularly true of the larger transmitter type tubes, since Class A operation requires that the grids be not driven positive nor beyond cutoff, and the larger type tubes develop relatively little power output over this restricted operating range.

Class B push-pull operation permits the grids to be driven positive and also beyond cutoff, without excessive distortion being produced. This is because the dynamic characteristics of the two tubes overlap in such manner that the combined characteristic is reasonably linear beyond cutoff of either tube. This is illustrated in Fig. 7.

The plate current of one tube is I_1; the plate current of the other tube is I_2. These vary in opposite manner for a grid excursion

e_g; i.e., as I_1 increases, I_2 decreases, and vice-versa. The output or load current is $(I_1 - I_2)$. It is obtained from the individual characteristics by subtracting the ordinates of one from the other.
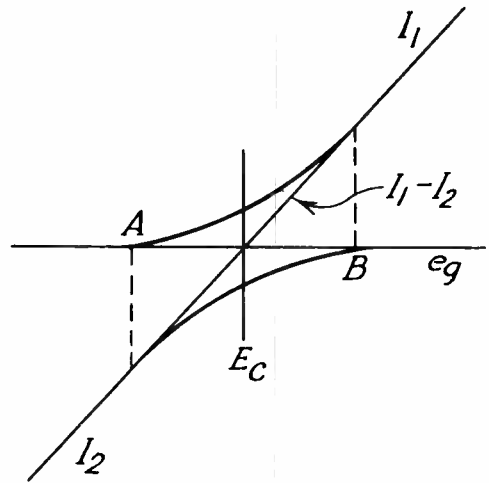

Fig. 7. — Composite characteristics of a push-pull amplifier.

Observe that although the individual characteristics are considerably curved, particularly near cutoff, the composite characteristic is fairly straight. Observe further that beyond cutoff of either tube (points A and B), the composite characteristic becomes identical with the appropriate individual characteristic, and each of these is reasonably straight for the region remote from cutoff.

Fig. 7 is the general push-pull characteristic. For Class B operation both tubes are biased just slightly above the extremely curved part of the characteristic preceding cutoff. This can be represented in Fig. 7 by moving points A and B

closer together. Essentially, one
tube furnishes power during one-half
of the modulating cycle, and the
other tube furnishes power during
the other half.

The grids of the tubes can
readily be driven positive by pro-
viding a driver tube of low enough
internal resistance, so that the
internal voltage drop in the tube is
not excessive at the positive peaks
of the grid swing, when the grids
draw current. Usually this is
accomplished by employing a push-
pull step-down transformer between
the grids and the driver tube.

The complete arrangement is
shown in Fig. 8. As in the previous
circuits, the amplifier stage acts
as a constant load resistance across
the secondary of the plate modu-
lation transformer T_m. By a suitable
choice of the turns ratio n, this
resistance can be made to present
the optimum value to the plates of
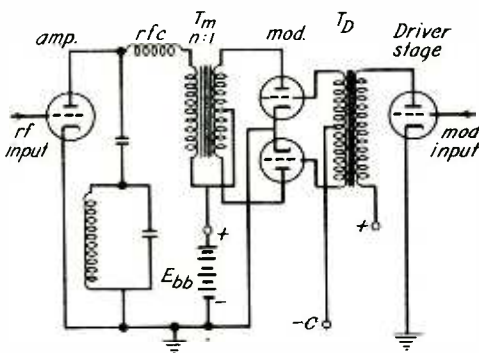the modulator tubes. Transformer T_D

to either half-secondary, thereby
making the R_p of the driver tube
appear sufficiently low to the grid
of either tube, and thus preventing
flattening of the modulating wave
when either grid is driven positive
and draws current.

The design of the Class B modu-
lator stage will be developed here
as an extension of the problem pre-
viously analyzed for the 203-A tube.
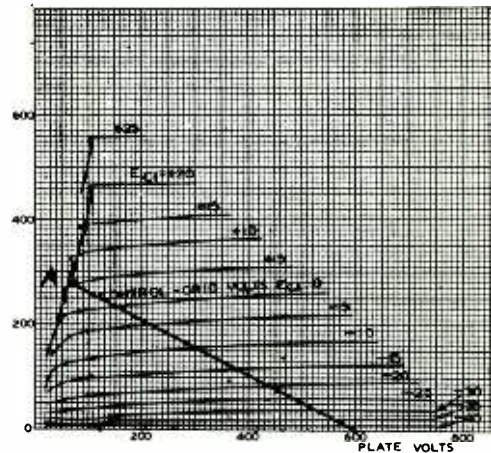In Fig. 9 are shown the plate charac-
teristics for an 807 tube. It is

Fig. 9.—Plate characteristics of an
807 beam power tube.

intended to operate two such tubes
in push-pull, and to drive the grids
alternately 5 volts positive. The
supply potential E_{bb} is to be 600
volts, and the screen potential 300
volts, so that it is advisable to
use a separate lower-voltage power
supply for the modulator stage than
for the 203-A amplifier stage (which
requires 1,000 volts).

Choose a point A on the +5-volt
curve, just to the right of the knee
of the characteristic. Join A to
the 600-volt abscissa, as shown.

Fig. 8.—Push-pull Class B modulator
stage.

is the driver transformer; it is
usually step-down from its primary

The slope of this line represents one-quarter of the optimum load resistance R_L to be presented to the extreme terminals of the primary; i.e., to both tubes. Thus, R_L is denoted as the plate-to-plate resistance.

Inspection of Fig.9 indicates a current I_{max} of 275 ma, and a value for E_{min} of 70 volts. The resistance of this load line is therefore

$$\frac{R_L}{4} = \frac{600 - 70}{.275} = 1,930 \text{ ohms}$$

so that

$$R_L = 4 \times 1,930$$

$$= 7,720 \text{ ohms plate-to-plate}$$

The power output of the two tubes will be

$$P_o = \frac{(E_{bb} - E_{min})\, I_{max}}{2} \qquad (6)$$

or

$$P_o = \frac{(600 - 70)(.275)}{2}$$

$$= 72.8 \text{ watts.}$$

This should be about sufficient to plate-modulate the 203-A tube; if the full 75 watts output is desired, R_L can be suitably reduced in the manner indicated above and the grid drive slightly increased.

A plate-to-plate resistance of 7,720 ohms is somewhat greater than the 6,667 ohms presented by the 203-A tube. Hence a step-up ratio from the secondary to the primary of

$$1 : \sqrt{7720/6667} = 1 : 1.076$$

is required. Although this is very close to unity ratio, the amplifier tube cannot be connected directly across the primary because the latter delivers a voltage balanced to ground, and the amplifier tube is single-ended. Hence a secondary coil will in general be required to perform the balanced to unbalanced transformation.

To complete the specifications, note that the maximum plate dissipation is 25 watts. For E_{bb} = 600 volts, the d-c component of each tube under no-signal conditions can be

$$I_b = 25/600 = 41.7 \text{ ma}$$

Examination of Fig.9 shows that -30 volts bias will permit 40 ma to flow at 600 volts on the plate, which is slightly less than the above permissible value of 41.7 ma. Under full grid swing, the d-c component rises, owing to self-rectification, to about 100 ma per tube. The d-c input power is now

$$W_{dc} = 600 \times .1 = 60 \text{ watts}$$

per tube. The output power is 72.8/2 = 36.4 watts. Hence the plate dissipation per tube is 60 - 36.4 = 23.6 watts, which is within the 25 watts permissible. Hence the 807 tubes should operate satisfactorily in a modulator stage.

In passing, it is of interest to note that two 845 tubes could have been employed instead of the 807 tubes, provided the supply voltage were raised to 1,000, and a load resistance of 4,600 ohms plate-to-plate were employed. The bias should be -175 volts, and the grids need not be driven positive in this case to obtain a power output of 75 watts. Unfortunately, the manufacturer does not furnish sufficient

range in the plate characteristics shown in the Tube Manual to permit a graphical construction to be made. Compare the 75 watts output of the two tubes operating Class B push-pull with the 73.2 watts output of the three tubes in parallel operating Class A.

R-F DESIGN CONSIDERATIONS. —The r-f tube that is being plate modulated requires consideration. The Class-C amplifier design given in a previous assignment can be very readily extended to cover the operation under modulation conditions. The c-w conditions discussed previously assumed the tube was keyed on to maximum output or keyed off to zero output, as in code transmission. Now the conditions for "gradual keying" or modulation will be studied, in order to see what further precautions must be taken.

In the first place, if 100 per cent plate modulation is desired, the plate voltage must be varied from twice E_b to zero. If the plate voltage is reduced to zero, and the grid bias and excitation maintained unchanged (as one would normally consider correct in plate modulation) it is evident that the grid will draw an inordinate amount of current. In other words, during the periods of inward modulation (modulation troughs) the grid current will become excessive, as will also the grid dissipation.

This indicates that the grid bias should preferably become more negative at such times or else the excitation should decrease so as to prevent excessive grid current flow. This in turn indicates that a certain amount of grid modulation should be introduced when plate modulation is employed. Such grid modulation inherently occurs when a grid-leak resistor is used for large grid currents tend to produce a large bias which in turn can be itself counteract the tendency to draw large grid currents.

Other factors, however, are also involved. The resistance that the r-f tube presents to the modulator stage is actually not quite constant over the modulation cycle, as was previously stated, and this in turn requires that the r-f excitation as well as the bias be modulated. With these further refinements, much more satisfactory, distortionless operation can be obtained.

An analysis by C. Y. Meng* indicates that for proper plate modulation, *the angle of flow should be the same at all times.* This means that as the plate voltage rises during the positive half cycle of the modulation cycle, the bias should correspondingly *increase* so as to maintain the relative position of cutoff during the r-f cycle, and the r-f excitation should correspondingly *increase* to maintain the same angle of flow.

If the angle of flow is constant, then the ratio of the r-f current to the d-c component will be essentially constant throughout the modulation cycle, and furthermore both will be directly proportional to the plate-supply potential, which is the power-supply voltage E_{bb} plus that of the output of the modulator tubes E_m.

Hence, finally, the r-f current and voltage will be proportional to the modulation voltage, and the ratio of the instantaneous d-c component to the instantaneous plate-

*"Linear Plate Modulation of Triode Radio-Frequency Amplifiers," Proc. I.R.E., Dec. 1940.

supply potential $(E_{bb} + E_m)$ will be constant. This in turn means that the load which the r-f stage presents to the modulator stage will be one of constant resistance, so that no distortion will be generated in the modulator stage. Note that the modulator stage can be considered as an *audio* driver stage, just as the preceding r-f stage is considered an (r-f) driver stage.

The requirement for constancy of angle of flow results in the need for modulating the driver excitation voltage and the bias, just as was stated previously. This means that as the plate-supply potential rises during the positive half of the modulation cycle, the bias must increase in a *negative* direction, and the r-f excitation must increase in the *same* direction as the plate-supply potential. The opposite must take place during the inward or negative portion of the modulation cycle.

This is, however, contrary to the mode of operation normally recommended and previously described.

Reference is made specifically to the use of grid-leak bias and a driver stage having a high internal resistance. When grid-leak bias is used, the bias of course increases with increase in grid current. The grid current increases when the plate-supply voltage drops, as during the negative half of the modulation cycle. Hence, grid-leak bias tends to increase rather than decrease when the plate-supply potential decreases.

It does, nevertheless, have a linearizing effect on the output because at the inward troughs of modulation, the bias automatically increases and helps reduce the plate current, thereby preventing flattening of the *troughs* in the envelope. On the outward peaks of modulation, the grid current and hence the bias decreases, thereby raising the r-f plate current and preventing flattening of the *peaks* of the envelope.

This is shown in Fig. 10 in rather exaggerated form. The grid leak, owing to the variation in grid

Flattening of peaks and troughs when fixed bias is used

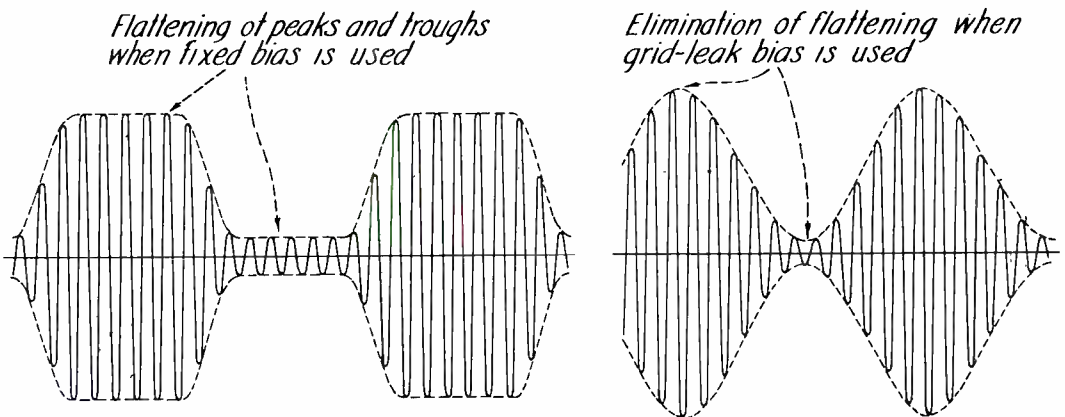Elimination of flattening when grid-leak bias is used

Fig. 10.—The use of grid-leak bias eliminates the distortion otherwise produced if fixed bias is employed.

current, automatically reduces the bias at the peaks and increases it at the troughs, thereby eliminating the flattening effects shown at the left in Fig. 10.

The action of a driver stage having a high internal resistance is also in the same direction. If a large grid current is drawn, as at the troughs of the envelope, then the internal voltage drop in the driver is high, and hence less r-f exciting voltage is available at the grid of the P. A. stage. Hence the r-f output of the P. A. stage is reduced, and the trough rounded out just the same as if the bias increases.

During the outward peak modulation the grid current decreases, so that more drive is available at the P. A. tube's grid. This raises the r-f voltage and prevents the flattening of the outward peaks. Fig. 10 therefore illustrates just as well the benefits of having a high internal driver resistance.

The combination of the two, however, do tend to offset one another. For example, during inward modulation, the grid current rise is partially counteracted both by the internal driver resistance and the increased (bias) voltage drop in the grid leak. The combined effect may be greater than that of either alone, but is not as great as if both were independently varied.

Nevertheless, this discussion seems to indicate that both the bias as well as the drive should vary in the same direction as the plate modulation voltage, whereas the analysis of Meng seems to indicate that the bias should vary in opposite manner to the drive and plate modulation voltages. What, then, is the benefit of the latter type of

operation?

The answer is in the constancy of resistive load that the r-f P. A. stage presents to the modulator stage. For example, suppose at the peaks of outward modulation, the excitation increases, but the bias decreases. Then the angle of flow increases, the d-c component tends to rise faster than the a-c (r-f) component, because it is a different kind of function of the angle of flow.

This in turn means that the resistance presented to the modulator stage at the peak of the cycle is less than at carrier conditions. In the same way, the resistance presented at the troughs is much higher than at carrier conditions. Meng reports a variation from 11,000 to 2,800 ohms for a 801 tube under constant-bias conditions, and down to 2000 ohms when grid-leak bias is used because of its contrary variation to that required. This is admittedly for Class B conditions, which is not a practical mode of operation for plate modulation, but it does indicate how variable the modulator load can be.

It would appear then desirable to modulate the bias 180° out of phase and the excitation in phase with the plate modulation, for proper operation. Some other beneficial effects claimed are elimination of hot spots on the plate owing to electron beam focusing at the peaks of modulation, and excessive grid dissipation during the troughs of modulation.

Actual practice, however, may be somewhat at variance with this mode of operation. Usually a combination of fixed and grid-leak bias is used. The fixed bias provides a safety feature in the event that

excitation voltage is lost. The grid-leak bias provides a compensating effect which may or may not be in the right direction, as will be explained next.

In addition, plate modulation is also often applied to the r-f driver stage. Thus a modulated excitation drive voltage is obtained, rather than one of fixed amplitude. This modulation is independent of the grid current, and hence can be used to make the grid-leak bias vary in the desired manner if it is of sufficient amplitude.

For example, at the peaks of modulation, the drive voltage is increased to a peak value too. It can be made high enough so as to draw more grid current than at carrier conditions, rather than less, as would be the case if its amplitude were fixed. As a result, the grid-leak bias can *increase* rather than decrease at the peaks of modulation.

Similarly, at the troughs of modulation, the excitation is reduced, and may draw less grid current in spite of the fact that the plate-supply potential is now E_b - E_m and hence at a minimum. This means that the grid-leak bias is *reduced* rather than increased at the troughs of modulation.

Whether the excitation modulation is sufficient to produce this effect to the necessary degree or not, it does tend to reverse the phase of the grid-leak bias variation. It may not be sufficient, however. In this case it may be sufficient to eliminate flattening of the peaks of modulation, but not to prevent variation in the resistance that the r-f tube presents to the modulator tube.

Since the modulator tube has appreciable internal resistance, the variable load resistance will produce variations in output voltage just as occurs in a Class AB_2 push-pull audio stage when the driver resistance is greater than zero.

The solution is inverse feedback. The distortions in the output signal owing to variations in load resistance have no proportional relation to the wave form of the input audio signal; inverse feedback can cancel or at least reduce the output voltage distortions, and thus minimize the effects of r-f tube load variations.

Inverse feedback is therefore employed because it is a simple way of obviating the distortion produced by load variations. Whether or not in addition the excitation is modulated depends upon the size of the equipment; a 1 kw transmitter may not have this feature, but instead have perhaps a driver stage of high internal resistance and regulation, whereas a 5 kw or larger transmitter may employ modulation of the excitation voltage.

PRACTICAL EXAMPLE. —As a practical example of these considerations, consider the RCA 5-KW Type BTA-5G broadcast transmitter. A simplified schematic is shown in Fig. 11. A brief discussion of the various features will not be remiss.

The two UL-4392 Oscillator Units are equipped with TMV-129B Crystal Units, which maintain the frequency within +5 cycles/sec. One is a spare in the event of failure of the other. The oscillator in turn feeds two 807 tubes in parallel acting as a buffer amplifier. Note that no neutralization is required, and furthermore than no plate tuning is required after the initial set-up is made. Observe further that the frequency monitor feed is taken from
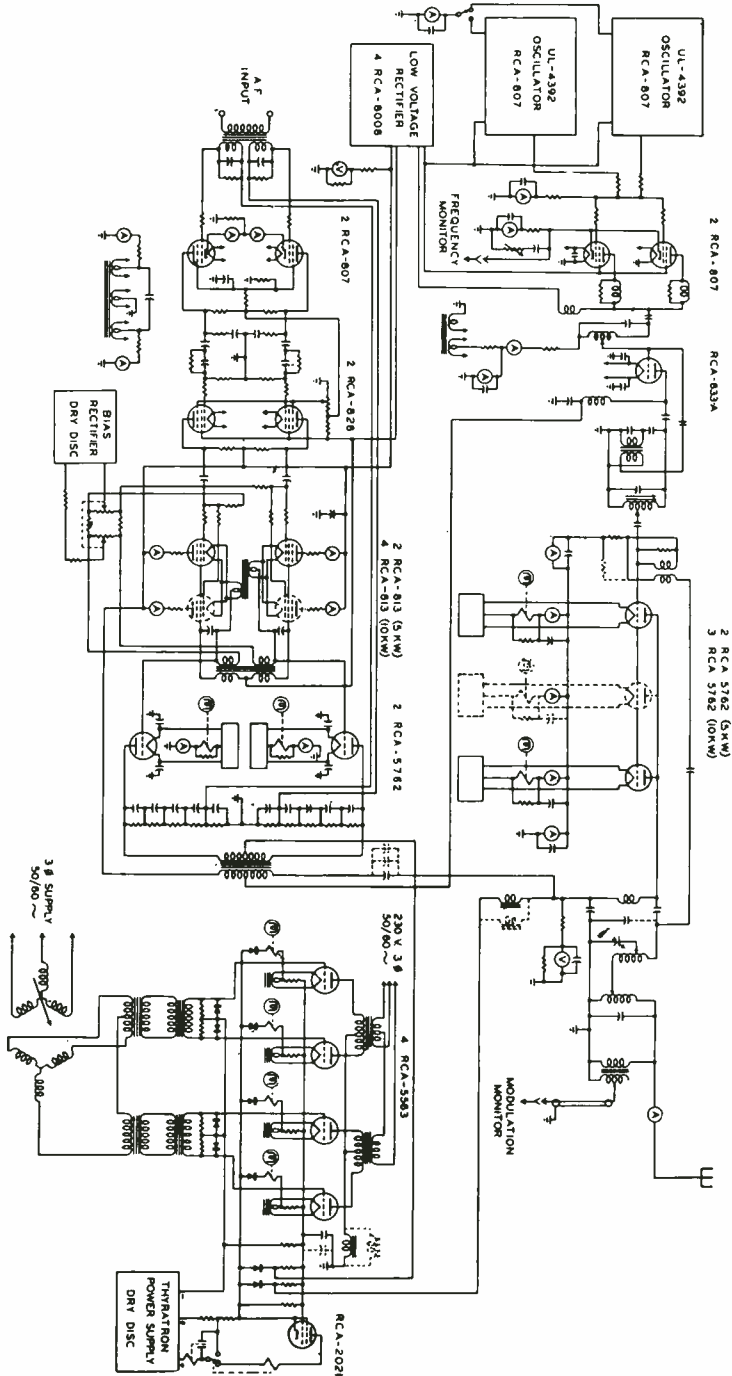
MODULATION

Fig. 11. — "Simplified Schematic of RCA Type BTA-5G 5 KW Transmitter, Broken Lines
Indicate Modification for 10 KW".

the cathode circuit of the buffer.

The latter excites the driver amplifier, which consists of one 833-A tube. This employs a slug-tuned inductor in the plate circuit and a broadband neutralization transformer. Thus a minimum of tuning adjustments are required.

The 833-A stage drives the P. A. stage, which consists of two 5762 tubes in the 5-KW BTA-5G job, and three 5762 tubes in the 10-KW BTA-10G transmitter. The plate circuit is tuned by a variable vacuum capacitor, and neutralization is here also accomplished by a broadband transformer. Observe further that the P. A. tank circuit and output matching network form a symmetrical network so the load impedance is the same to both side bands, thereby eliminating a possible source of distortion. An important feature of this transmitter is the elimination of air dielectric capacitors, and the use of but two tuning controls and one power output control.

We now come to the modulation amplifier. Audio signal is fed to two 807 tubes operating Class A push-pull. These are resistance-coupled to two 828 tubes operating in similar manner. The 828 tubes are resistance-coupled to two 813 tubes operating in push-pull Class AB_1 (in the 5-KW transmitter).

The 813 tubes are operated as cathode followers to provide a very low impedance driving source. Note that the 5762 modulator grids return directly to the cathode circuits of the 813 tubes. The 813 screen grids are maintained at a-c cathode potential by means of the screen bypass capacitors and the transformer secondary. Resistors in the plate return circuits are for the purpose of maintaining equal currents in the 813's.

No output signal is taken from the plates of the 813 tubes.

The output tubes (two 5762) operate Class B, and provide the high-level modulation power required by the plates of the r-f P. A. Note that modulator bias is obtained from a separate selenium rectifier.

The push-pull modulation output transformer has a tapped secondary winding. The full winding connects through a blocking and coupling capacitor to the top end of an audio modulation choke, whose lower end connects to the high-voltage supply. The secondary winding itself connects at its lower end to the low-voltage supply. This low voltage is applied through the tap on the secondary to the 833-A driver stage to provide the plate-supply voltage plus a suitable amount of plate modulation for this stage.

The blocking capacitor, on the other hand, blocks this low d-c voltage from the Type 5762 power tubes, and enables instead the high d-c voltage from the high-voltage rectifier to get to them. At the same time the blocking capacitor permits the full audio modulation voltage to get to the tubes in series with the high d-c voltage.

In this way the full output r-f voltage and the lower input r-f excitation voltages are correspondingly modulated by the modulator. Note that both the 5762 grids have series resistors to provide grid-leak bias in conjunction with the modulated excitation. Observe further that fixed bias is obtained by cathode filament resistors, so that should the excitation fail, there will still be a safe amount of bias on the tubes.

The feedback in the modulator unit is obtained as follows: From

the plate of each 5762 tube to ground is a series of resistors, shunted by capacitors to form an R-C voltage divider that is relatively insensitive to further capacity loading of any particular tap. As shown, the lowest tap is used to provide feedback voltage back to each half of the secondary of the A-F input transformer. Thus, voltage feedback occurs from each output plate to each input grid; since an odd number of resistance-coupled stages are involved, the feedback is negative or inverse in nature. In this way distortion due to any variation in the load impedance presented by the P.A. stage to the output modulator stage is minimized.

DESIGN CONSIDERATIONS. — If cathode emission rather than plate dissipation is the limiting factor, then the emission will determine the peak output, and the carrier voltage will have to be half that at the peak, in order to permit 100 per cent plate modulation.

However, usually plate dissipation is the limiting factor, since pulse operation is not being considered. In this case, assuming constant plate efficiency over the modulation cycle, the momentary plate dissipation at the peaks will be four times that at carrier level, since the output power is quadrupled. But the AVERAGE plate dissipation over the ENTIRE MODULATION CYCLE will be in proportion to the average output power. This is 50 per cent greater than the carrier power, hence the average plate dissipation should be 50 per cent greater than that at carrier level.

Working backwards, we see that if W is the maximum permissible plate dissipation, and occurs at 100 per cent modulation, then the permissible

plate dissipation at carrier level must be 1 : 1.5 or 2 : 3 or two-thirds of the maximum permissible. This in turn means that the plate-supply potential should be about two-thirds rather than half of that which incurs maximum permissible plate dissipation. Stated in another way, if E_{bb} is the plate voltage for ordinary Class-C C-W operation, then $2/3\ E_{bb}$ is the plate voltage required for a plate-modulated Class-C amplifier. The modulation voltage will be the same, and the peak plate voltage will therefore be $4/3\ E_{bb}$.

Now as to the bias and drive conditions. If E_c is the bias and E_g the excitation required for Class-C C-W operation, then $2/3\ E_c$ is the bias required, and $2/3\ E_g$ is the excitation required for a plate-modulated stage. This is assuming a linear tube, and at least serves as a start in the design procedure.

If the requirements given by Meng are employed, somewhere in the neighborhood of 100 per cent modulation of E_c is required, and 80 per cent of E_g*. If only E_g is to be modulated, then perhaps as much as 100 per cent modulation should be employed, but an experimental run should be made to check the linearity of the modulation and the constancy of the load resistance presented by the r-f stage to the modulator stage.

GRID MODULATION. — A typical grid-modulated stage is shown in Fig. 12. It will be observed that there are three voltages in series in the grid circuit: the r-f driving signal e_{rf}, the modulating signal e_m, and the d-c bias E_c. At such instants that e_{rf} and e_m are

*More complicated rules are given by Meng, and the interested reader can refer to his article for this information.

simultaneously at their positive peaks, the bias E_c is cancelled to the greatest degree, and the grid is driven most positive with respect to the cathode, whereupon it draws maximum current. The plate current is then also at a maximum.

current i_p be as nearly proportional to the modulating voltage e_m as possible, that the power output be as large as possible without the plate dissipation exceeding the allowable value for the tube, and the driving power, both r-f and a-f
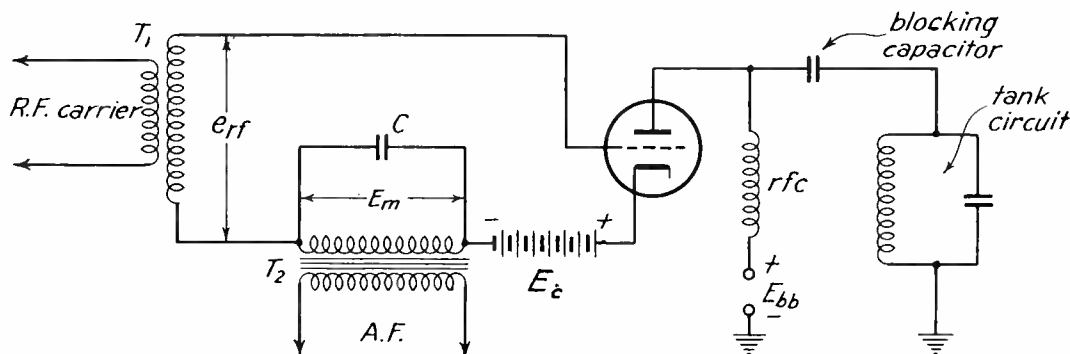

Fig. 12.—Grid-modulated stage.

On the other hand, the magnitudes of the voltages are so adjusted relative to the tube that when e_m is maximum negative and e_{rf} is maximum positive, the instantaneous total grid voltage is barely above cutoff of the tube, whereupon practically no plate current flows, and of course no grid current.

This is illustrated in Fig. 13. The dynamic plate characteristic is represented by AB; the corresponding grid characteristic by OC. A fairly large negative bias E_c is chosen. The modulating voltage e_m, of peak value E_m, varies about the bias line at a low-frequency rate, and the r-f excitation voltage e_{rf}, of peak value E_{rf}, then varies about e_m.

The operating conditions require that the r-f or fundamental component i_{rf} of the pulses of plate

(modulating power), be as small as possible.

In general these requirements are met by adjusting the bias for zero modulation or carrier conditions to about the same value as for Class C telegraphy (which is usually the maximum power output form of operation for the tube). The audio or modulating voltage is then of such magnitude, that on the positive peaks (point D in Fig. 13), it reduces the bias to a value somewhere in the neighborhood of Class B operation; i.e., to around the cutoff value. This is not a hard-and-fast rule, and considerable variations will be found from one type of tube to another.

The peak r-f grid swing E_{rf} must now be of such a value, that at the maximum negative excursion

$(E_c - E_m)$, it brings the grid just up to the cutoff value, or slightly beyond. This means that plate current will just fail to flow, or at most, only a small amount at the tips of the r-f excitation. This

As an example, Fig. 14 shows the operating conditions for an 803 tube. For grid modulation, the d-c bias is -80 volts, or within 10 volts of the Class C bias of -90 volts. Note that the maximum positive voltage
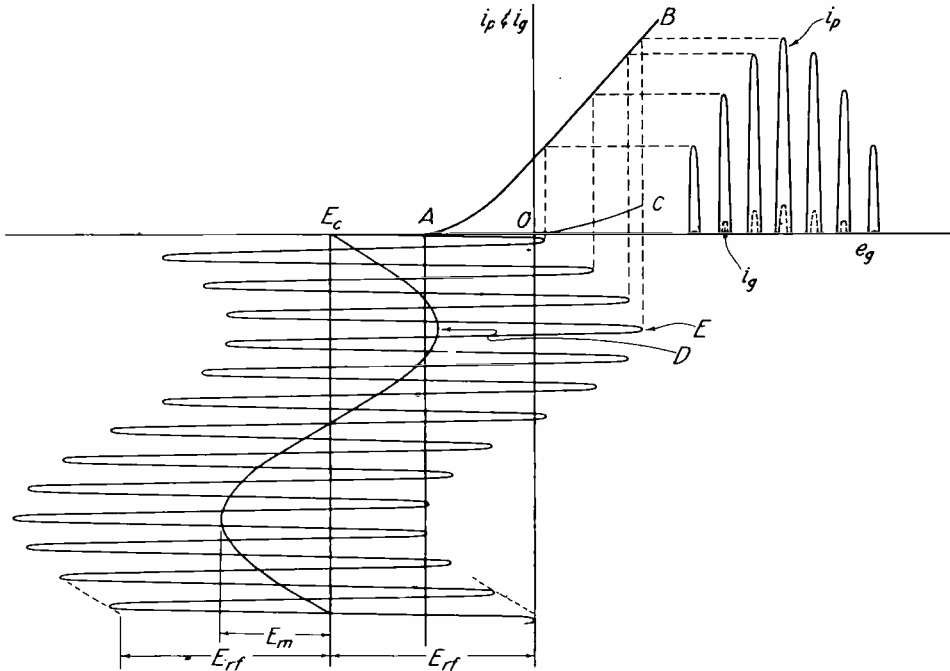

Fig. 13.—Tube characteristics involved in grid modulation.

corresponds essentially to zero carrier amplitude and 100 per cent inward modulation.

On the other hand, on the positive half-cycles of modulation, the sum of the three voltages must bring the grid to a positive value close to that attained for Class C telegraphy, in order to obtain as much output on 100 per cent outward modulation. Hence the values of E_m and E_{rf} are interrelated to meet the two conditions given in this and the preceding paragraph.

attained by the grid in the latter mode of operation is +85 volts. Both the Class B and grid-modulated stages should hit approximately the same positive voltages on 100 per cent outward modulation. The maximum positive grid potential is found to be +70 volts in either case, which is fairly close to the +85 volts for Class C operation.

Normal Class B bias is -40 volts, and the instantaneous bias attained in grid modulation on the positive half of the modulation cycle

is -30 volts, or somewhat less than the above value, but not far removed from it. Similarly, on the negative half-cycle of modulation, the grid

portion contributes very little to the modulation capabilities.

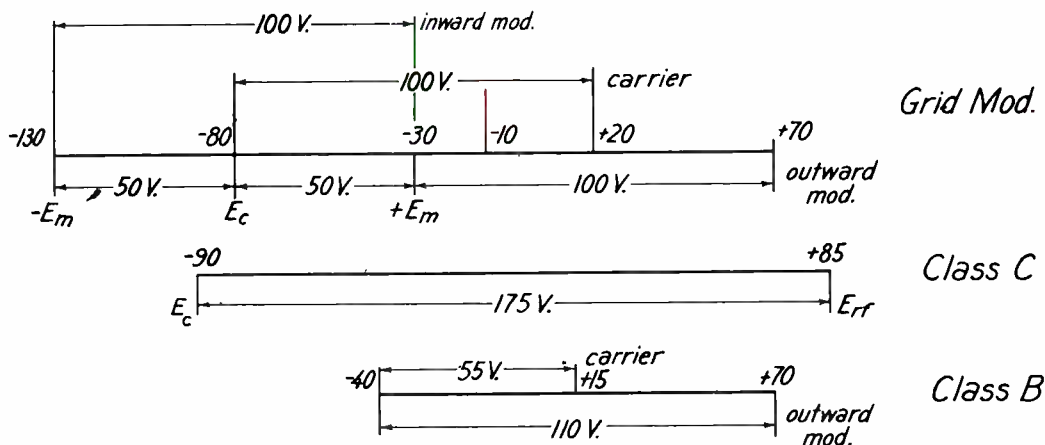Another factor is that the 803 tube's plate-current characteristics

Fig. 14.—Comparison of grid modulation, Class C, and Class B operation for an 803 tube.

swings to -130 volts, or the same as the previous peak positive modulation bias value; i.e., somewhat above cut-off, on the basis that the Class B bias of -40 volts represents cutoff.

This brings out a modification to the above rule for the amplitude of the r-f grid excitation: if the cutoff portion of the tube's dynamic characteristic is appreciably curved, then on 100 per cent inward modulation appreciable distortion of the modulation envelope will be obtained. To avoid such distortion, it is better to adjust E_{rf} and E_m so that on inward modulation the grid swing peaks are sufficiently above cutoff as to avoid the curved portion of the characteristic. Although not quite 100 per cent modulation will be obtained, the percentage will be close to this, since the curved

are such that appreciable plate current does not flow until the grid is driven positive. In short, most of the useful operating range is in the positive grid region, and -40 or -30 volts does not produce so very much difference in the operation.

The angle of flow of plate current varies during the modulating cycle. In the ideal case illustrated in Fig. 15, for 100 per cent inward modulation, the angle of flow is zero (point g); for 100 per cent outward modulation, the angle is 180°, in that of corresponds to one-half cycle. For the carrier level, the angle of flow is between 180° and 0°, since cd is less than ab = 180°.

Since in general the efficiency decreases as the angle of flow increases, the efficiency is least at

the instant of 100 per cent outward modulation, and most at the instant of 100 per cent inward modulation, when the power, however, is practically zero.
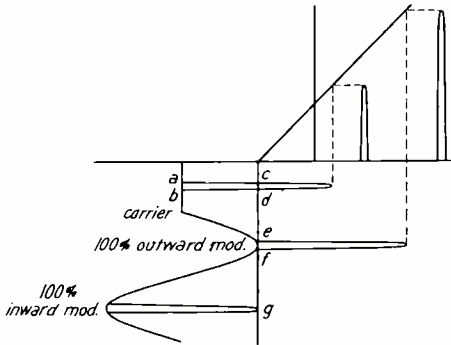

Fig. 15. — Variation in angle of flow of plate current during grid modulation.

The efficiency is less than for plate modulation, and the power output is considerably less, because the plate voltage is constant, instead of increasing on outward modulation. At the same time, since the fundamental current increases and decreases in a symmetrical manner if the grid is modulated by a symmetrical voltage, such as a sinusoid, the average component of the pulses of plate current will also vary symmetrically, since it is approximately equal to one-half of the fundamental component.

In short, the d-c component corresponding to the carrier level does not change when the tube is grid-modulated, *so that the input power to the tube is constant.* However, the r-f output power *increases under modulation* in that

side band energy is produced. Hence the *average* efficiency of operation over the modulation cycle must increase during modulation, and the plate dissipation decrease.

The peak output (at 100 per cent outward modulation) will be no more than the tube can furnish as a Class C telegraph amplifier, and since the *average* power during modulation is less than the peak power, it is apparent that the modulated output is even lower as compared to the tube operating as a Class C telegraph amplifier, and also in general, less than when operated as a plate-modulated amplifier.

Since the first cost of the tube is appreciable, it is desirable to get as much power output as possible, and hence plate modulation will be preferred to grid modulation on that account. On the other hand, grid modulation requires far less audio power than plate modulation, and audio power is at least as expensive as r-f power.

DESIGN FEATURES. — Mention has been made that on extreme inward modulation, the very curved portion near cutoff of the dynamic characteristic tends to flatten the envelope of the modulated wave, and hence produce distortion. It was also stated that this can be avoided by so adjusting the amplitudes of the modulating and r-f driving voltages that on inward modulation the tips of the r-f wave are within the operating range and reach the more linear portion of the characteristic.

Another means is to rectify some of the output voltage, thus recovering the distorted modulating voltage from the envelope, and feeding this back in inverse manner so as to obtain degenerative feedback, with its beneficial minimizing

effects on the distortion. The effect is to peak the negative halves of the modulating wave so as to drive the grid more negative. Since this causes the grid to traverse the more curved portion of the dynamic characteristic, it is advisable to have sufficient r-f drive to exceed the cutoff range, as stated above, so as to afford the feedback voltage more room on the curved portion before complete cutoff is reached.

Another factor of importance is the effect of the grid current both on the r-f driver stage and on the modulator stage. As shown in Fig. 16, when e_{rf} and e_m drive the grid positive, grid (electron) current
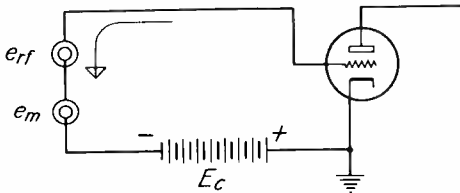

Fig. 16.—Simplified diagram of the grid circuit of a grid-modulated stage.

flows in the direction shown. This direction of current flow is such as to *charge* the bias battery, hence the latter should not be a dry battery or of the primary cell type, since these polarize when charged and tend to change their voltage. Even a storage battery will tend to overcharge, so that it is preferable to employ a rotating machine or rectifier-type bias supply of sufficiently low regulation to prevent appreciable change in the bias voltage.

The grid current not only

supplies energy to the bias source, but also energy to heat the grid (grid dissipation). This energy comes in part from the r-f driver stage, and in part from the audio stage. The grid current flows in pulses. In the absence of modulation, these pulses are of constant amplitude, and have a fundamental or r-f amplitude, and an average component that is constant in magnitude, or d.c., of approximately half the amplitude of the fundamental component.

Since this d-c component flows against the potential of the bias battery, it represents charging or input power (as explained previously), and this power must come from the driver r-f source. Hence the product of the r-f component of grid current and the driving voltage E_{rf} represents power drawn from the driver stage; in part to supply the grid dissipation, and in part to supply the input charging power.

When the tube is modulated, the pulses of grid current vary in amplitude at the modulation frequency. The r-f or fundamental component is now amplitude-modulated, and the side band power, as well as a certain fraction of the original carrier power, is dissipated in the form of heat at the grid of the tube, while the rest of the original carrier power is dissipated in the d-c bias source.

The average component of the grid current pulses now varies at the modulation frequency, and hence develops in addition to the original d-c component, an a-c component at the modulation frequency. The latter component is in phase with the modulation voltage, and hence represents power absorbed from the modulator. This power is also

dissipated at the grid, if the bias source be assumed to be adequately by-passed at the modulation frequencies.

The average component is actually considerably distorted. As shown in Fig. 17, no current flows during a portion ab of the negative half-cycle of modulation, because the net voltage of the grid is negative throughout this time. Hence the average component, as shown by the dotted line, is clipped on the negative portions. So far as the
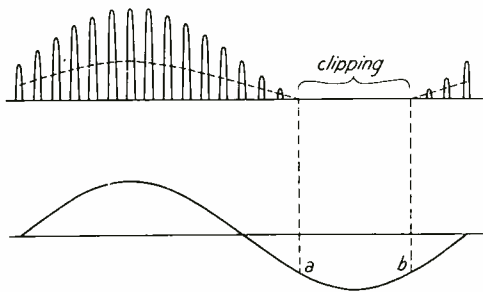

Fig. 17.—Flow of grid current during modulation cycle.

modulator is concerned, the grid circuit appears as a resistor that is switched in only during part of the modulation cycle, and is variable even during the interval it is switched in, owing to the nonlinear properties of the grid even during conduction.

This is shown in Fig. 18. Switch Sw closes only during a portion of the modulation cycle, and even then R_g, the grid-to-cathode resistance (plus any other resistance in the circuit) varies with the current flowing through it. It is a *minimum* when the current is a *maximum* through it; i.e., at the
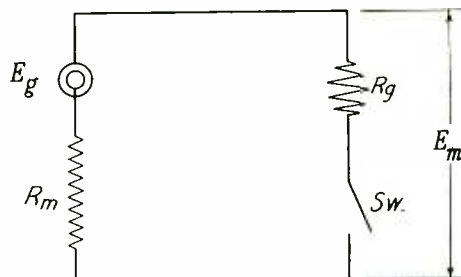
positive peak of the modulation cycle.

Fig. 18.—Grid circuit as it appears to the modulator.

The effect on the modulation voltage E_m, as it appears at the secondary terminals of the modulation output transformer, is as follows: When no grid current flows, there is no voltage drop in the modulator source, and the full generated voltage E_g appears at the output terminals; i.e., $E_m = E_g$. During the portion of the modulation cycle when grid current flows, E_m drops below E_g by an amount equal to the IR drop in the modulator internal resistance R_m.

As a result, if E_g, for example, is sinusoidal in shape, E_m will be flattened on the positive peaks, as shown in Fig. 19(A). Such an asymmetrical distortion indicates the presence of even harmonics, principally the second. Also some d-c component is generated by this non-uniform conductivity, which is opposite to the flow of the normal d-c component of grid current. This is illustrated in Fig. 19(B), where E_m is the terminal voltage, E_f is its fundamental component, E_{2f} is the second-harmonic component, and

E_{dc} is the d-c component.

Other harmonic components are present, but of small amplitude. As a reasonable approximation, the
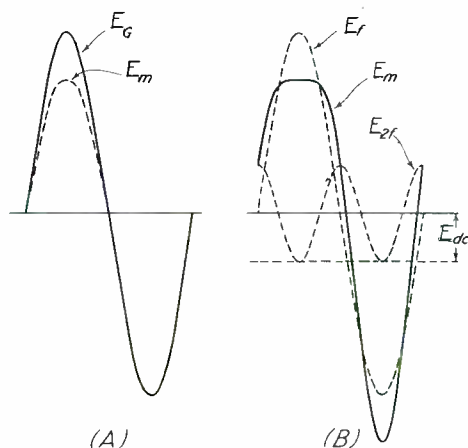

Fig. 19. —Distortion of modulation voltage by grid current flow.

distortion may be assumed all second harmonic, E_{2f}, plus an equal amount of E_{dc}. The analysis is similar to that for the per cent second-harmonic distortion of the current in a vacuum tube amplifier, as discussed in a previous assignment.

Let the permissible ratio of I_{2f} to I_f be d. Then d × 100 is the per cent second-harmonic distortion, and corresponds to a difference in the positive and negative peaks of E_m. Let n be the ratio of the absolute values of the positive and negative peaks and be a positive number less than one, for example, 0.85. Then an analysis shows that

$$n = \frac{1 - 2d}{1 + 2d}$$

For example, if the permissible fraction of second-harmonic distortion is d = .05, or 5 per cent,

then the fraction that the positive peak is of the negative peak is

$$n = \frac{1 - (2 \times .05)}{1 + (2 \times .05)} = \frac{.9}{1.1} = 0.818$$

The negative peak of E_m is equal to the generated voltage e_g; the positive peak of E_m is less than E_g by the voltage drop in the modulator internal resistance R_m. This voltage drop is produced by the modulation component of the grid current. If the latter is known, then the instantaneous resistance R_g of the grid circuit can be found, and from the desired value of n, as determined by d, the modulator internal resistance to meet these conditions can be calculated.

It is therefore first necessary to calculate the value of R at the positive peak of the modulation cycle. The maximum value of *average* current is either given by the manufacturer, or can be measured for the tube by decreasing the d-c bias voltage on the tube (no modulation) to a value corresponding to the instantaneous minimum bias occurring when modulation is applied. Then r-f excitation is applied to the grid, and a *d-c meter* in the grid circuit is used to read the d.c. flowing.

This represents the instantaneous peak value of the modulation component of the average current when modulation is normally applied. It is illustrated in Fig. 20. Here E_c is the normal bias, and E_m the normal peak modulation voltage. By changing the bias voltage from E_c to E_c' (= $E_c + E_m$), and then applying E_{rf} momentarily, a series of grid current pulses I_{gm} flow, whose average value I_{ga}, when read on a d-c meter, represents the peak value of the modulation component.

In normal operation I_{ga} flows but monentarily when the modulation component E_m reaches its peak value $+ E_m$. Hence, the instantaneous resistance is

$$R_g = E_m/I_{ga} \qquad (8)$$

As an example, a Tube Manual shows the peak d-c component of the grid current of the 803 tube to be 19 ma. The words d-c are used because the manufacturer has presumably used the test procedure illustrated by Fig. 20, in which the grid bias
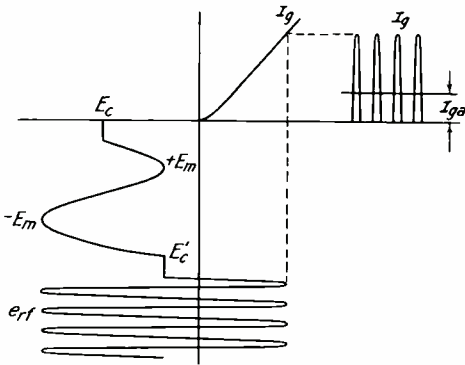

Fig. 20. —Operating characteristics of grid circuit to obtain peak modulation current.

has been changed to E_c and held at that value while the current was read, whereupon the current represented d.c. under the test conditions. From Fig. 14 it is clear that $+ E_m = 50$ volts. Hence the instantaneous (and minimum) grid resistance is, by Eq. (8),

$$R_g = 50/.019 = 2,630 \text{ ohms}$$

Therefore, at the positive peak

of the modulation cycle, the circuit appears as in Fig. 21.
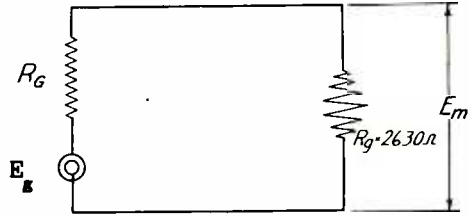

Fig. 21. —Grid circuit as it appears to the modulator at the positive peak of the modulation cycle.

From the figure it is clear that

$$\frac{E_m}{E_g} = \frac{R_g}{R_0 + R_g} \qquad (9)$$

where E_g is the peak of the generated modulation voltage, and E_m is the peak of the terminal voltage that is actually injected into the grid circuit.

On the *negative* half-cycle, there is no voltage drop in R_0 because R_g is essentially infinite, so that $E_m = E_g$. Hence the ratio of E_m/E_g in Eq. (9) corresponds to the ratio of the modulation terminal voltage on the positive and negative half cycles; that is, to n. Thus

$$n = R_g/(R_0 + R_g) \qquad (10)$$

But n is determined from Eq. (7) by specifying the permissible distortion d. This in turn means that the left-hand side of Eq. (9) is determined, and R_g, the instantaneous grid resistance, is determined by the tube employed, hence R_0, the internal resistance of the modulator, is fixed by the preceding factors.

Solving for R_G in terms of the other factors, we obtain

$$R_G = \frac{4d}{1 - 2d} R_g \qquad (11)$$

The value of d in the problem under consideration is .05, and R_g was found to be 2,630 ohms. Hence,

$$R_G = \frac{4 \times .05}{1 - 2 \times .05} \, 2,630 = 584 \text{ ohms}$$

This is obviously far less than the R_p of any available tube to be used in the modulator stage, and indicates that a matching transformer will have to be employed as shown in Fig. 22, by T. By making this of the

The peak modulation voltage E_m required by the 803 grid is 50 volts, and the required source impedance R_G is 584 ohms. Suppose a 6F6 tube is chosen, triode-connected. From a Tube Manual, its constants for Class A operation are found to be:

$$R_p = 2,600 \text{ ohms}$$

$$\mu = 6.8$$

$$E_c = -20 \text{ volts (for } E_{bb} = 250 \text{ volts)}$$

For an infinite load impedance (as when the grid is not drawing current), the maximum apparent
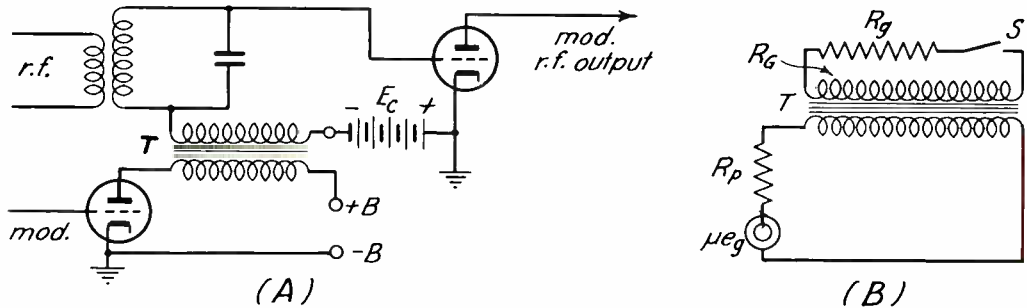

Fig. 22.—Use of a matching transformer for grid modulation, and the equivalent circuit.

proper step-down ratio n : 1, the actual R_p of the tube will be reduced to the desired value R_G. In (A) is shown the actual circuit; in (B), the equivalent circuit which makes the calculations clearer.

However, in stepping down R_p to R_G, the transformer also steps down the modulator output voltage, and unless the tube or tubes employed are large enough, insufficient modulator voltage will be obtained.

To show this, consider further the problem under consideration.

voltage generated in the plate circuit is μE_c. This is stepped down by the modulation transformer to the peak modulation voltage

$$E_m = \mu E_c / n \qquad (12)$$

where n is the step-down turns ratio.

The value of a is given by

$$n = \sqrt{R_p / R_G} \qquad (13)$$

If the value of n from Eq. (13), when substituted in Eq. (12), gives a value

for E_m sufficient for the modulation requirements, then the tube is satisfactory for use in the modulator stage.

Taking the 6F6 tube as a trial modulator, we find from Eq. (13) that

$$n = \sqrt{2,600/584} = 2.11$$

Substituting this value in Eq. (12), there is obtained

$$E_m = 6.8 \times 20/2.11 = 59.4 \text{ volts}$$

which is sufficient for the purpose, since 50 volts are required.

Had a smaller value for the distortion d been employed, R_0 would have come out smaller, too. As a result, a greater step-down ratio n would be required, and E_m might then come out too small. In such an event, two tubes in push-pull can be employed to furnish the required output. When two tubes in Class A push-pull are employed, the effective R_p is *double* that of either tube, and also μE_c must be doubled. It will be found that if a single tube is not quite sufficient for the purpose, then two tubes in push-pull will be more than adequate.

Finally, if two tubes in Class A push-pull are not sufficient, Class B operation at a HIGHER plate voltage can be used. However, the effective R_p is now FOUR times as great, but the bias and hence grid swing can be increased to more than compensate for the higher source resistance. For example, for two 6F6 tubes in Class B, E_c can be increased to -38 volts, and E_{bb} to 350 volts.

If, furthermore, the grids are driven *positive*, then the value of E_m in Eq. (12) is found by using the corresponding larger grid swing instead of $2E_c$. For example, if either

grid is driven positive to +20 volts from a bias of -38 volts, then in solving for E_m, $2 \times (38 + 20) = 116$ volts are used instead of $2 \times 38 = 76$ volts. In short, the PEAK A-F GRID-TO-GRID VOLTAGE is employed instead of $2E_c$.

In conjunction with this, $4R_p = 4 \times 2600 = 10,400$ ohms instead of $2 \times 2600 = 5,200$ ohms must be used. This will give a greater step-down value for n; nevertheless, owing to the greater apparent generated emf of 116 volts, E_m will be greater, and hence possible sufficient to allow a smaller value for d. This is left as a problem for the student in the examination at the end of this assignment.

FURTHER DISCUSSION.—In general, pentode tubes are not as well adapted for such applications as the above, where the load is variable over the cycle, because their R_p is so high that their regulation is excessive when the load—such as the grid circuit—varies markedly during the signal cycle.

Of course a fixed load resistor can be connected across the secondary, so that R_g is in parallel with it. In such a case the variations in R_g do not produce such marked variations and hence distortion in the modulation wave, but such operation requires a larger tube than where the load resistor can be dispensed with and the proper step-down transformer alone employed, as in the case of a triode tube.

Another possibility is to use inverse voltage feedback with the pentode tube. This type of feedback makes the apparent R_p of the tube appear much lower than it really is, and does not require a large tube. It does, however, require a greater modulation input voltage.

It is possible to employ grid modulation on a low-level tube, and then employ linear amplifiers to bring the modulated wave up to the desired output level. In such a case the power considerations with respect to the low-level grid-modulated tube are not of such importance, and it is therefore feasible to employ such small values of modulation and r-f drive voltage that the grid at no time goes positive.

In this case the grid-modulated tube presents essentially an infinite resistance to both the r-f driver and audio modulator tubes, so that the above distortion produced by grid current does not occur. The result is modulation with a minimum of distortion. However, as mentioned in a previous assignment, the linear amplifier tube has to be relatively large compared to the power output it develops, and also requires attention so that the proper operating voltages and settings are maintained during operation over a period of time.

OTHER METHODS OF MODULATION. — Modulation can be accomplished by impressing the modulation voltage on any other available electrode. For example, in a tetrode or pentode tube, the screen grid may be used for modulation. In a pentode tube, the suppressor grid may be used for this purpose, and in any type of tube, the modulation voltage can be injected in series with the cathode (cathode modulation). Usually these methods are employed only in special cases, since they do not possess any important advantage over the more usual plate and grid modulation.

SCREEN GRID MODULATION. — In the case of a tetrode or pentode tube, the plate current does not change appreciably when the plate voltage

is varied, as is apparent from inspection of the characteristic curves for such a tube. A large plate current change is observed when the control or screen grid potentials are varied.

Hence, when plate modulating such a tube, it is necessary to modulate the screen voltage as well, and the percentage changes in the voltages should be in proportion to their d-c voltages. If the screen voltage is obtained from a fixed d-c supply, then the proper percentage of audio voltage must be injected in series with the supply. This is shown in Fig. 23(A), where secondary S_1 of many turns supplies audio voltage to the plate, and S_2 of fewer turns supplies the proper proportion of audio voltage to the screen. An alternative arrangement is to feed the screen from a suitable tap on the plate secondary through a coupling capacitor and a-f choke coil in series with the screen supply.

If the screen is fed through a series dropping resistor from the plate supply, then for modulation purposes it is connected to the plate secondary as shown in Fig. 23(B). Dropping resistor R drops *both* the d-c and modulation voltages to the proper values for the screen, and thus insures they both bear the same ratio to the corresponding plate voltages.

However, in the case of many tetrode tubes, the bottom end of R can be returned directly to B$^+$. Modulation voltage variations will occur in the screen resistor owing to the inherent variations in the screen current when the plate is modulated. In the case of a tetrode using a fixed supply voltage, an a-f choke coil can be interposed between the screen and this supply,

and screen modulation obtained with-
out any appreciable d-c drop in
the choke coil.

plate modulation is by far the most
common method, with grid modulation
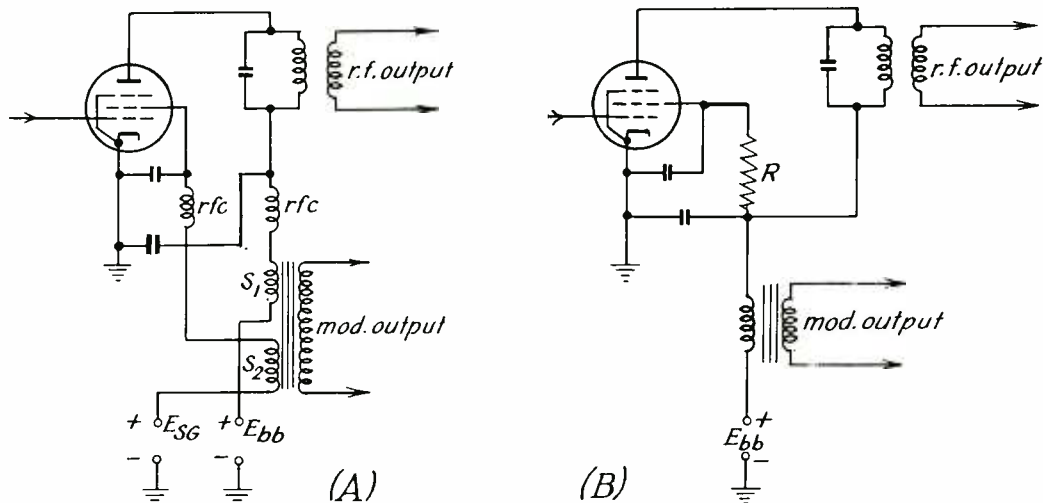next in importance.

Fig. 23. — Screen modulation for a pentode tube, used in conjunction with plate
modulation.

SUPPRESSOR MODULATION. — The
suppressor grid may be modulated in
a Class C amplifier. The circuit is
very much like that for grid-modu-
lation, except that the modulating
voltage is impressed in the suppre-
sor grid circuit, while the r.f. is
impressed in the control grid cir-
cuit. The suppressor is normally
given a negative bias, so that it
draws current only at the peaks of
the audio cycle. The design con-
sideration for the modulator stage
are exactly the same as for grid-
modulation.

There is not any particular
advantage in suppressor grid modu-
lation, except the ability to keep
the audio and r-f circuits separate,
nor is the modulation particularly
linear. In certain special cases it
may have some particular advantage;

CATHODE MODULATION. — In Fig. 24
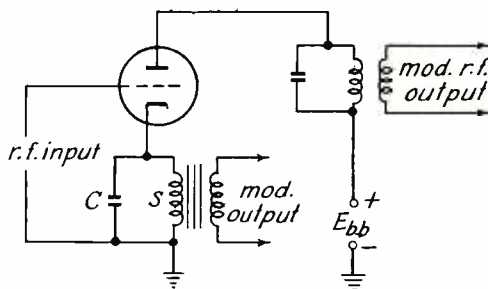is shown a cathode-modulated stage.

Fig. 24. — Cathode-modulated stage.

Secondary S injects the modulation
voltage between the cathode and
ground. (C is an r-f by-pass
capacitor.) The modulation voltage

is actually injected by this form of circuit arrangement into both the grid and plate circuits, so that the stage partakes of the features of both grid and plate modulation.

However, the amount of modulating voltage required for 100 per cent modulation is quite small because of the amplifying effect in the grid circuit, hence the amount of grid modulation is relatively small. On the other hand, the load presented to the modulator is predominantly that of plate modulation, with some variable effect from the parallel nonlinear grid current flow. Hence, the load is fairly constant during the modulation cycle, and the circuit easy to drive, although it requires appreciably more power than the grid-modulated amplifier.

MEASUREMENT OF MODULATION. — When a carrier wave is modulated, additional power is developed in the tank and antenna circuits. This is a result of the production of power in the side bands; this power is in addition to that present in the carrier wave.

To make this clear, assume sinusoidal modulation of the carrier wave. As shown in Fig. 25, when the carrier wave is modulated, the

Fig. 25. —Amplitude modulation of a carrier wave.

average amplitude I_{avg}, is equal to the normal unmodulated carrier amplitude I_c. One may therefore say that the carrier wave persists even during modulation, but that in addition side band waves are produced that combine to give the final amplitude variation shown.

For 100 per cent modulation, as shown in Fig. 25, the peak current I_{peak} is twice I_{avg}, which is equal to the carrier amplitude I_c. Let the resonant tank circuit impedance be R_T, a pure resistance. Then the peak power is

$$P_{peak} = (I_{peak})^2 R_T \qquad (14)$$

The carrier power is

$$P_c = (I_c)^2 R_T \qquad (15)$$

Since, for 100 per cent modulation, $I_{peak} = 2I_c$, Eq. (14) may be expressed as

$$P_{peak} = (2I_c)^2 R_T = 4E_c^2 R_T \quad (16)$$

$$= 4P_c$$

That is, the peak power is four times the carrier power.

However, the peak power occurs only momentarily in the modulation cycle. At the moment of maximum *inward* modulation, the power is zero. Hence, the power during 100 per cent modulation fluctuates between four times carrier power and zero power in a periodic manner. The *average* power during modulation might therefore appear to be simply P_c, the carrier power.

This, however, is not so. It is true that the current amplitude varies from I_{peak} to zero in such manner that its average is I_c, the carrier amplitude, but the power varies as the *square* of the current
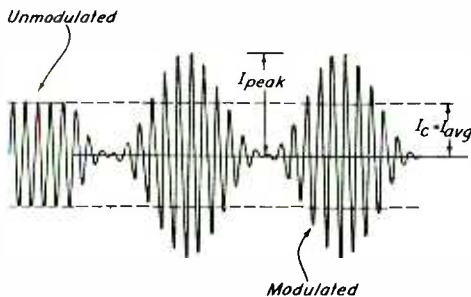
amplitude, and hence the average of the power fluctuation is not P_c, but 50 per cent more than this (for 100 per cent sinusoidal modulation). In short,

$$P_{avg} = 1.5 \ P_c \qquad (17)$$

The reason why the average power exceeds the carrier power is that the squares of the current amplitudes (to which the instantaneous power is proportional) rise to *four* times I_c at 100 per cent *outward* modulation, but drop only to zero at 100 per cent *inward* modulation. Thus the increase in power above P_c during outward modulation exceeds the decrease below P_c during inward modulation; the average comes out to be $1.5 \ P_c$.

This means that during 100 per cent modulation, .5 or 50 per cent of P_c is power developed in the side bands. This power must come from some source. In the case of plate modulation, this additional power comes from the modulator output stage, as has been indicated; in the case of grid modulation, the power is that diverted from plate dissipation and converted into useful power, a desirable feature.

Thus, during modulation, a grid-modulated tube runs cooler, and at a higher plate efficiency, while the d-c input from the power supply remains unchanged. In the case of plate modulation, the d-c input from the power supply also remains unchanged, but in this case the additional power comes from the modulator unit, and the efficiency of the r-f stage remains substantially constant throughout the modulation cycle.

The important thing to the radio engineer in a broadcast station is the reading of the r-f ammeter,

such as that in the antenna circuit. Corresponding to the carrier power P_c, there is an ammeter reading I_c. However, since the meter is an a-c meter, and reads r.m.s. rather than average values, its reading during 100 per cent modulation will not be $I_{avg} = I_c$, as shown in Fig.25, but a higher reading proportional to $\sqrt{P_{avg}}$.

Since $P_{avg} = 1.5 \ P_c$, the corresponding current reading during modulation will be

$$I_m = I_c \ \sqrt{P_{avg}/P_c} = I_c \ \frac{1.5 \ P_c}{P_c}$$

$$= I_c \ \sqrt{1.5} = 1.225 \ I_c$$

In short, the antenna meter will read 22.5 per cent higher during 100 per cent modulation than it does during periods of no modulation.

As an example, suppose the unmodulated carrier reading is 20 amperes. Then during 100 per cent modulation, the reading will rise to

$$20 \times 1.225 = 24.5 \text{ amperes}$$

a net increase of 4.5 amperes.

If the modulation is less than 100 per cent, the increase in reading will be correspondingly less. The increase will depend upon the degree of modulation m. This represents the ratio of the peak variation in current amplitude to the average or carrier amplitude. Refer to Fig.26. Here I_1 represents the increase in the current amplitude over the average or carrier (unmodulated) value I_{avg}. For a symmetrical modulating wave, the decrease in amplitude will also be I_1, and a single fraction m will be obtained. The value of m is simple given by

$$m = I_1/I_{avg} \qquad (19)$$

For 100 per cent modulation, I_1 clearly equals I_{avg}, and m = 1. When I_1 is zero, m = 0, and the wave is unmodulated; thus m varies between zero and unity.
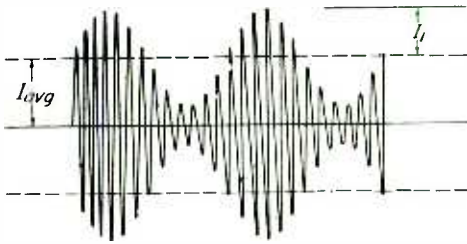

Fig. 26.—Factors determining the modulation fraction m.

The antenna ammeter reading is proportional to

$$\sqrt{1 + \frac{m^2}{2}}$$

or

$$I_m = \left(\sqrt{1 + \frac{m^2}{2}} \right) I_c \quad (20)$$

Thus, if the unmodulated carrier current reading is 20 amperes, and the percentage modulation is 60%, so that m = 0.6, the ammeter reading during modulation will be

$$I_m = 20 \sqrt{1 + \frac{.6^2}{2}} = 20 \sqrt{1.18}$$

$$= (20)(1.087) = 21.74 \text{ amperes}$$

The percentage increase is given by

$$100 \left[\sqrt{1 + \frac{m^2}{2}} - 1 \right]$$

and for m = 0.6, is 100(1.087 - 1) = 8.7 per cent.

To summarize, note that the

instantaneous peak current will be 1.6 × 20 = 32 amperes; the minimum current will be (1 - .6)20 = 8 amperes; and the average current will remain 20 amperes, but the *r.m.s. current* I_m corresponding to *average power* during modulation will be 21.74 amperes, and it is this latter value that the a-c ammeter will read, since it is of the r.m.s. type, such as a thermogalvanometer movement.

The increase in power during modulation will be proportional to the square of the ammeter reading, or for m = 0.6, it will be in the ratio of

$$(21.74)^2 / (20)^2$$

$$= (1.087)^2 = 1.18 \text{ to one}$$

Alternatively, this ratio is very simply given as

$$\left(1 + \frac{m^2}{2} \right) \text{ or} \left(1 + \frac{.6^2}{2} \right) = 1.18$$

As a guide, the following chart and graph (See Fig. 27) gives the relation between the percentage modulation and the percentage increase in antenna current.

These figures are for sinusoidal modulation. In actual practice the carrier is modulated by whatever speech or music is impressed upon the microphone; the modulation is neither sinusoidal in shape nor or constant percentage. As a consequence, the antenna meter will vary constantly from its minimum unmodulated carrier value to higher values, which should not exceed approximately 122.5 per cent of the minimum value.

Overmodulation (in excess of 100 per cent), particularly on inward

modulation, will produce a distorted envelope of the carrier wave, and will be heard as distortion in the receiver. On the other hand, it is desirable to modulate the carrier to as nearly 100 per cent as possible in order to deliver a strong useful signal, but the modulation peaks should not exceed 100 per cent.

PERCENTAGE OF MODULATION	PERCENTAGE INCREASE IN ANTENNA CURRENT
0	0
5	0.063
10	0.25
15	0.56
20	1.00
25	1.55
30	2.23
35	3.02
40	3.92
45	4.94
50	6.01
55	7.30
60	8.63
65	10.05
70	11.57
75	13.19
80	14.89
85	16.67
90	18.53
95	20.47
100	22.47

Limiter amplifiers employing supercontrol tubes or the equivalent are now generally used at a point in the system preceding the final modulator stage. The supercontrol tubes are so biased (somewhat like delayed a.v.c.), that the gain is constant so long as the audio input level does not exceed a certain value. On loud passages of music or speech, however, the bias is exceeded, and the transconductance of the super-

control tube is thereupon reduced. This produces a corresponding reduction in the gain of the limiter amplifier, so that the *output* for this increased input signal is very little greater than that for a threshhold input signal. Thus, overloading of the power amplifier stage is prevented.
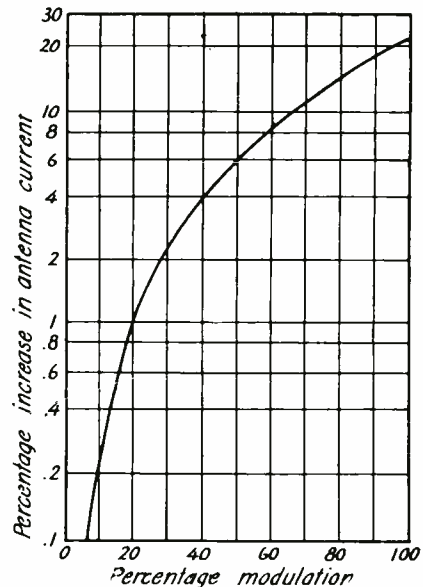

Fig. 27.—Relation between precentage modulation and percentage increase in antenna current.

The practical advantage of such a limiter amplifier is that the control room engineer can raise the average level of the program so as to obtain more nearly 100 per cent modulation on passages of average loudness, without incurring the danger of overmodulation on sudden loud passages.

For the actual measurement of the percentage modulation, a modulation meter is not entirely satis-

factory because its operation in general depends upon the heating effect of an element, so that it is relatively sluggish and unable to respond to mementary peaks of speech and music. It may be observed that even if a more rapid principle of operation were employed, the actual mechanical inertia of the moving element would prevent the meter from responding with sufficient rapidity. Hence the meter is best suited for following broad changes in volume level, and for making routine measurements of modulation in transmitter adjustment and checking.

For continuous indication of the depth of modulation during program operation, a more rapid and definite indication is required. Many modern broadcast transmitters employ, as a part of the permanent installation, a cathode ray oscillograph. This instrument allows the engineer continuously to observe the modulation peaks and so to adjust the audio input level that the peaks approach 100 per cent modulation

but do not over-modulate.

Another very convenient device is a neon tube which flashes when the modulation peaks reach the overload point. Thus when frequent flashes on a program indicate that it is not properly monitored, the audio, level is reduced to below the overload point. The neon tube is connected to the output of a simple vacuum tube voltmeter circuit.

FREQUENCY MODULATION

The fundamental theory and characteristics of frequency modulation, as well as some typical transmitter and receiver circuits, have been discussed in a previous assignment. This assignment will go into somewhat more detail concerning the modulation methods.

THE ARMSTRONG SYSTEM. —In Fig. 28 is shown a block diagram of the Armstrong system. The audio input is modified by the correction amplifier so that:
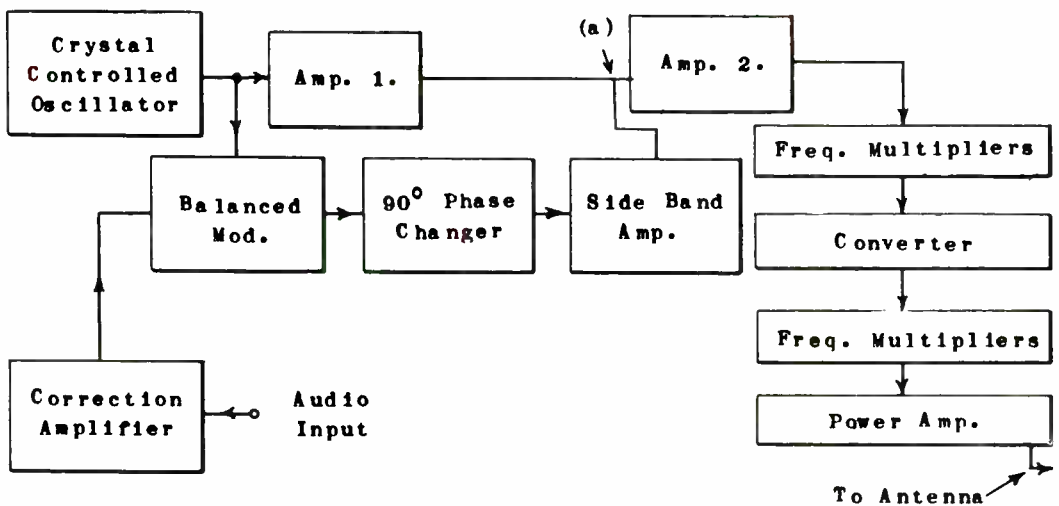

Fig. 28. —Block diagram of the Armstrong system of frequency modulation.

1. The amplitude varies inversely with the frequency in order to obtain frequency rather than phase modulation.

2. The higher frequencies are accentuated in accord with the preemphasis desired.

This, it will be recalled from a previous assignment, is a characteristic of the same shape as the impedance of an R-L circuit whose time constant L/R is 75 μsec.

The output of the correction amplifier is applied in push-pull to the screen grids of the balanced modulator (shown in Fig. 29). The crystal-controlled master oscillator feeds both control grids of the balanced modulator in parallel. A usual crystal frequency is 200 kc.

Note that the plates of the balanced modulator are connected in push-pull. In the absence of modulation, the screen grids are at a common d-c potential E_{sc}, and the control grids, being driven in phase, cause both plate currents to vary in phase. Since the plate currents flow in opposite directions through L_1 and L_2, and since these are equally coupled to L their magnetic effects cancel in L, and hence no voltage is induced in it; i.e., there is no carrier output.

When modulation is impressed on the screen grids, one grid goes positive when the other goes negative, so that the plate currents are unbalanced, and differentially effect L. A voltage is induced in L
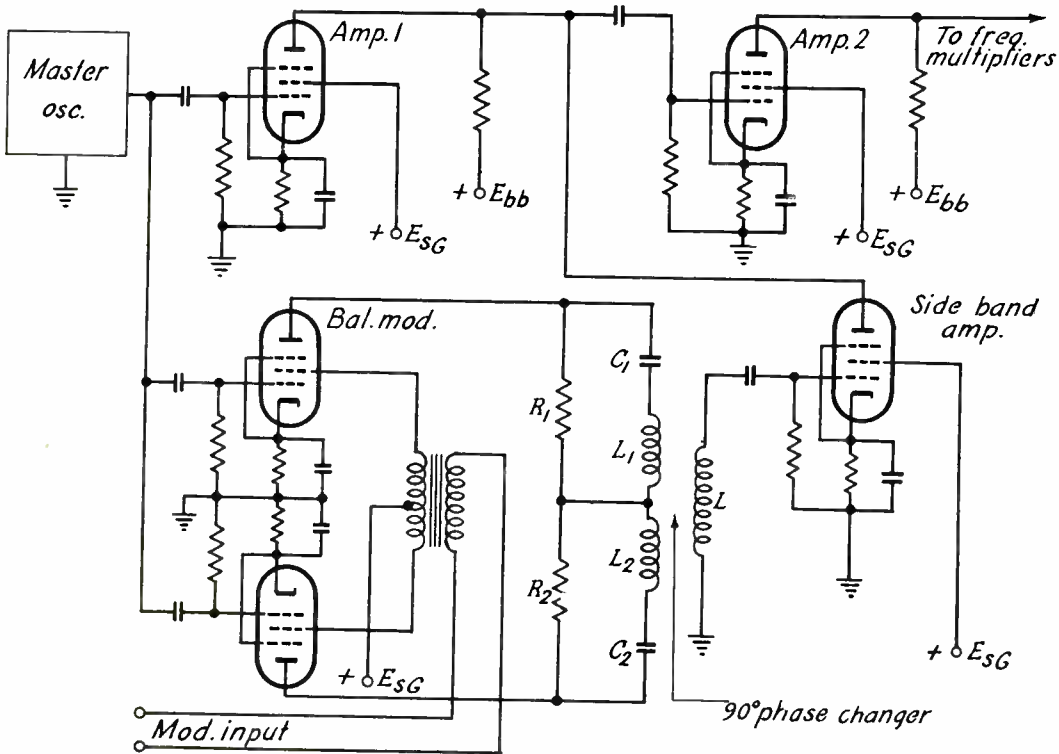

Fig. 29.—Schematic circuit diagram of Armstrong's balanced modulator.

as a result. When the modulation reverses polarity on the next half cycle, the voltage in L reverses, but otherwise has the same shape as before, so that the envelope of the wave repeats its shape of the previous half cycle.

Its appearance is therefore as in Fig. 30. Note that the envelope (shown in dotted lines) has the shape of a full-wave rectified sine wave.
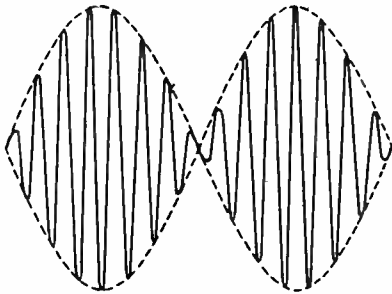

Fig. 30. —Appearance of output voltage of balanced modulator.

This effect can be obtained by taking a 100 per cent sinusoidally modulated wave, as shown in Fig. 31(A), and
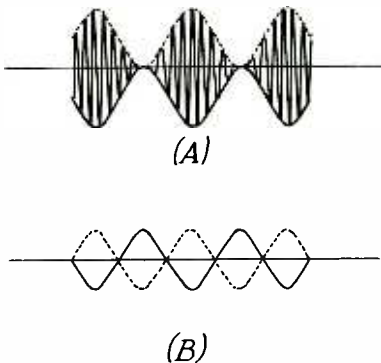
moving the two envelopes together until they cross, as in Fig. 31(B).

But this process is nothing more nor less than that of subtracting out the carrier, since it is the latter that prevents the envelopes from crossing one another, and at most, at 100 per cent modulation, permits them to touch one another. In short, the balanced modulator furnishes only the side bands and deletes the carrier.

The two circuits $L_1 C_1 R_1$ and $L_2 C_2 R_2$ are non-reactive at the oscillator frequency. This means that the current in $L_1 C_1$ and in $L_2 G_2$ is in phase with the plate current of each tube, and hence with the incoming control-grid driving voltages. Each of these currents, however, induce a voltage in L which is 90° lagging the current, as has been explained in previous assignments.

Hence the resultant of the two side band voltages in L is 90° lagging the currents in L_1 and L_2, and hence 90° lagging the r-f drive voltage on the grids. This will be made clearer by reference to Fig. 32.

(A)

(B)

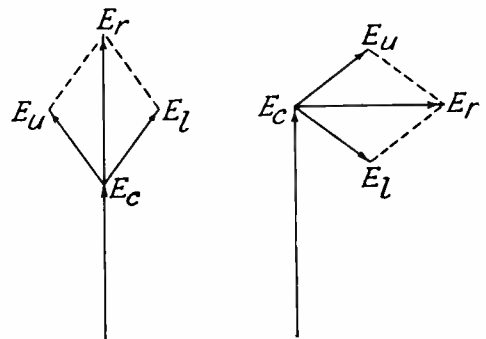Fig. 31. —Removal of the carrier frequency moves the envelopes together.

Fig. 32. —Shifting in phase by 90° of the side bands.

Here E_r is the resultant of the upper
side-band voltage E_u and the lower
side-band voltage E_l. As was ex-
plained in a previous assignment, E_r
is ordinarily collinear with the
carrier voltage E_c.

When E_r is rotated in phase
from its former position by 90°,
both E_u and E_r are similarly rotated,
as indicated in the right-hand fig-
ure. Hence, when the output of the
side-band amplifier is now combined
with the carrier output of Amplifier
No. 1(refer once more to Fig. 28),
the overall resultant of E_r and E_c
(the output of Amplifier No. 2) swings
back and forth through a range $\pm\theta$,
as indicated in Fig. 33, thereby
producing both phase and frequency
modulation, as well as some ampli-
tude modulation. The latter is elim-
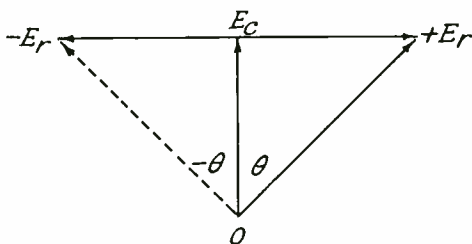inated by limiting in Amplifier No. 2

Fig. 33.—Production of phase modu-
lation by recombination of carrier
and 90° lagging side bands.

There is one factor in this
process that must be taken into
account. The voltage induced in L
is greater, the greater the fre-
quency. This means that E_u is some-
what longer than E_l, and this throws
E_r out of line with E_c in the left-
hand diagram of Fig. 32, and also
slightly less than 90° lagging with

respect to E_c in the right-hand dia-
gram.

To compensate for this, a
series R-C circuit is connected to
the output of Amplifier No. 2, with
the voltage across the capacitor fed
to the frequency multipliers. This
voltage is approximately in propor-
tion to the reactance of the capaci-
tor, and therefore decreases with
frequency. Therefore, by a proper
choice of R and C the upper side
band E_u can be reduced to equality
of amplitude with the lower side
band E_l.

The frequency multipliers raise
the frequency to some higher value,
and also raise the frequency devia-
tion in the same ratio. Since the
modulating process described above
produces a certain amount of phase
shift, and since the accompanying
frequency deviation is proportional
both to the phase shift and the
modulation frequency, the least fre-
quency deviation (for a fixed phase
shift angle) will be produced by the
lowest modulating frequency, and
hence the maximum multiplication
will be required in this case. The
required multiplication may, how-
ever, raise the carrier frequency
above the required value.

For example, suppose the lowest
modulating frequency is 30 c. p. s.
For low distortion, it has been
shown in a previous assignment that
the phase shift angle should not
exceed 25.5°. For convenience,
assume the maximum shift is 28.7°,
which is one-half radian.

The relation between phase
shift angle in *radians*, θ, the modu-
lating frequency f_m, and the fre-
quency deviation f_D is given by

$$\theta = f_D / f_m \qquad (21)$$

Thus, if $f_m = 30$ and $f_D = 75,000$ c.p.s.,

$$\theta = 75,000/30 = 2,500 \text{ radians}$$

Since the initial phase shift angle is 1/2 radian, the frequency multiplication will have to be $2,500/(1/2) = 5,000$ times. If the initial carrier is 200 kc = 0.2 mc, the final carrier frequency will be $5,000 \times 0.2 = 1,000$ mc, which is more than 10 times the desired value, for example, 88.1 mc.

Furthermore, an overall multiplication of 5,000 requires not only an unduly large number of multiplier stages, but the inherent noise produced in these stages, (which produces in turn a certain amount of noise frequency modulation), will become a serious problem where so many multiplier stages are involved. Hence, in the practical system, no more than about 3,000 times is attempted in the frequency-multiplying process; this corresponds to a final deviation of

$$\frac{3000}{5000} \times 75 = 45 \text{ kc}$$

instead of 75 kc at 30 c.p.s. The deviation is then allowed to rise linearly until at 50 c.p.s. a deviation of 75 kc is attained.

However, a direct multiplication of 3,000 times will produce a final carrier frequency of $3,000 \times 0.2 = 600$ mc, which is still too high. Hence, a heterodyne process is employed which lowers the carrier frequency but leaves the frequency deviation unchanged.

As a specific example, suppose the 200 kc wave is first multiplied 128 times to 25,600 kc. A further multiplication of $3,000/128 = 23.5$ or about 24 times is required.

Suppose first a local oscillator (crystal-controlled) of a frequency 21,929 kc is heterodyned with the above 25,600 kc output. The result is a difference beat frequency of

$$25,600 - 21,929 = 3,671 \text{ kc}$$

When this is then multiplied 24 times, the carrier frequency is raised to

$$3,671 \times 24 = 88.1 \text{ mc}$$

and the deviation to slightly over 45 kc (owing to the use of the whole number 24 instead of the fraction 23.5 for the second multiplication factor).

It is important to notice one point in this process. It is claimed that this system is direct crystal-controlled. Actually, however, it is *two-crystal*-controlled, since the heterodyne oscillator is involved as well as the primary crystal-controlled oscillator. The present F.C.C. regulations call for a maximum permissible frequency drift of $\pm 2,000$ c.p.s. On the basis of a carrier frequency of 88.1 mc, this corresponds to 22.7 parts in a million, or 0.00227 per cent. This same percentage holds for the heterodyne beat frequency, since it and its drift are both multiplied by the same factor, namely, 24 times.

However, the heterodyne oscillator must be accurate to a much greater degree, since every cycle per sec. drift in it is a cycle/sec. drift in the beat frequency. Therefore, it must be accurate to

$$\frac{3671}{21929} \times 22.7 = 3.8 \text{ parts in a million}$$

to cause no more than 22.7 parts drift in a million at 88.1 mc. This

does not take into account the drift in the primary oscillator. When multiplied 128 times, its frequency is 25,600 kc, or 3,671 kc above the heterodyne oscillator and therefore relatively close to it.

Hence, if as a first approximation, both are assumed of the same frequency, then their combined drifts must not exceed 3.8 parts in a million, or each must be good to approximately 1.9 parts in a million, if equal constancy is assumed for the two.

At the standard broadcast frequency a drift of ±20 c.p.s. is permitted by the F.C.C. This corresponds to about ±12.5 parts in a million at the high end of the band. It can be seen that even at the high end of the band the requirements are not nearly as stringent as for an f-m transmitter. Several solutions of this problem have been made, of which two will be described here. (Other methods, including an improved version of the Armstrong system, are given in the specialized broadcast section.)

THE PHASITRON TUBE. — If the initial phase shift can be increased without producing distortion, then less frequency multiplication will be required, and hence less constancy of frequency required of the initial oscillator, and perhaps the need for a heterodyne oscillator dispensed with.

One system is that developed by General Electric, based on a method proposed by Robert Adler of the Zenith Radio Corporation. The heart of the system is a special electron device called the phasitron tube. This tube has a cathode that projects a radial sheet of electrons, rotating at radio frequency, upon a perforated cylindrical target surrounding the axial cathode.
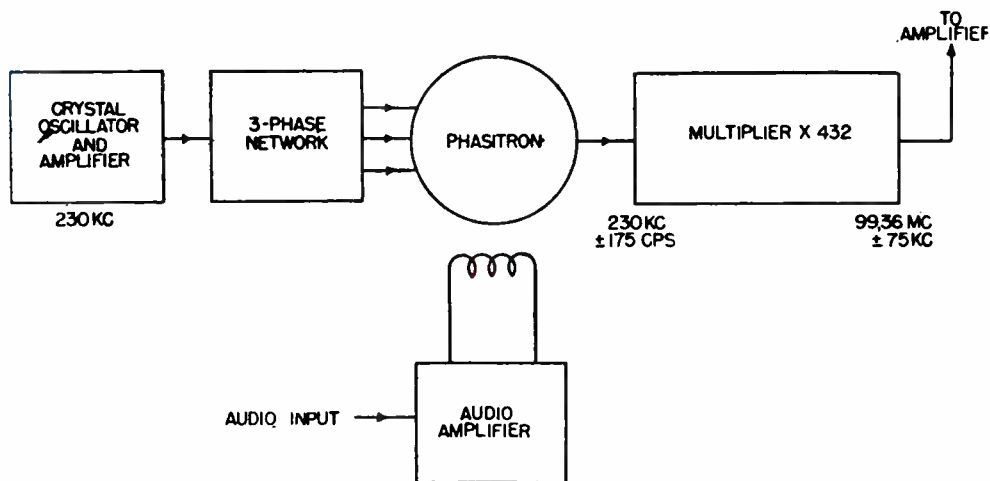
A series of grids warps the radial sheet of electrons into a pleated sheet, which more or less penetrates the holes in the perforated target and deposits its electrons upon the outer surrounding electrode. An axial magnetic field set up by the modulator causes this pleated sheet to rotate through an additional angle depending upon the strength of the modulation current, and thus permits penetration of the perforated target earlier or later in the r-f cycle. This makes the maximum current collected by the outer electrode occur earlier or later in the r-f cycle, and thus constitutes a phase shift in the r-f current wave. The phase shift can be several cycles instead of one-half radian or less, and therefore requires less subsequent frequency multiplication.

Before describing the phasitron tube in greater detail, a block diagram of the system, as illustrated in Fig. 34, will be discussed first. The r-f output of a crystal oscillator is converted into a 3-phase supply of the same frequency, and applied to three sets of grids in the phasitron tube to producing a rotating field. This produces the rotating pleated effect in the radial sheet of electrons.

The audio modulation input is fed to a magnetic coil, which produces the phase shift in the output. A phase shift of 3.5 radians with but 1.5 per cent distortion is possible as compared to 0.5 radian in the Armstrong system. As a result, frequency multiplication of the order of 432 times produces both a desired carrier frequency of 99.36 mc and a frequency deviation of ±75 kc. Note that no heterodyning

is required, so that the initial oscillator need have a stability of 2,000/99,360,000 = 20.1 parts in a million, the same as the final carrier. This compares favorably with the requirements in the standard broadcast band.

neutral plane. Directly below this sheet of electrons are 36 separate grid wires, in 3 sets of twelve each, as shown (ABC, ABC, etc.). All the A grid wires are connected together, and similarly for the B and C sets of grid wires.
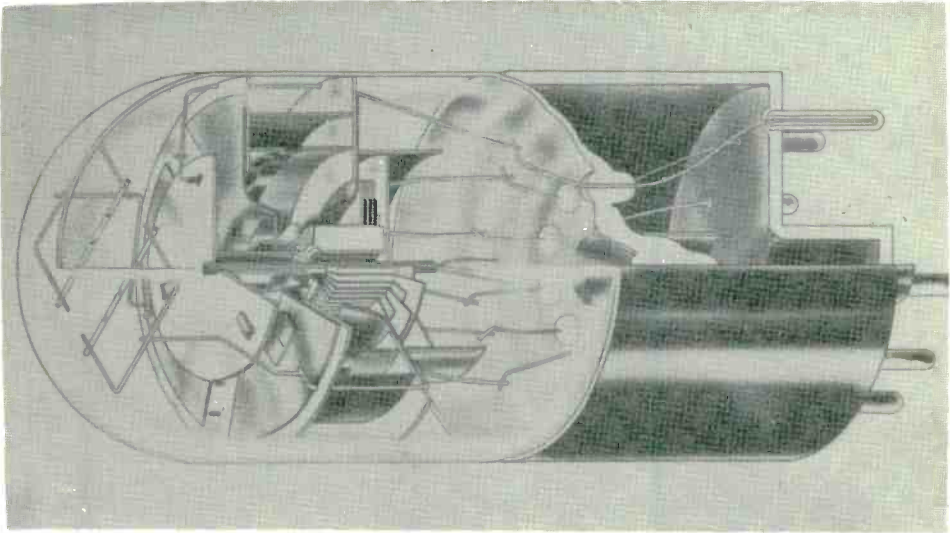

(Courtesy Electronics)

Fig. 34.—Block diagram showing the use of the phasitron tube in an f-m transmitter.

The phasitron tube itself is shown in Fig. 35. It is a maximum of $4^9/16$ in. high by $1^5/8$ in. in diameter; in short, a tube slightly larger than the ordinary radio receiving tube. A simplified sketch of the internal structure is shown in Fig. 36. The axial cathode is surrounded by a cylindrical focus electrode #1, a neutral plane electrode, and a second focus electrode #2.

The two anodes (#1 and #2) are at a positive potential and draw electrons from the cathode. The two focusing electrodes cause the electrons to flow radially in the form of a sheet directly below the

If a negative potential, for example, is applied between the cathode and the A set of grid wires, the radial sheet of electrons is repelled by this set, and is pleated or ruffled as shown in Fig. 37. If instead, the B set of grid wires is given a negative potential, the ruffles will move over a notch so that their maxima are directly over the B set, and similarly for the C set.

If now, a three-phase group of voltages is applied to the three sets of grid wires, a rotating field will be set up, which will cause the ruffles or pleats to rotate around the cathode as an axis, and at a
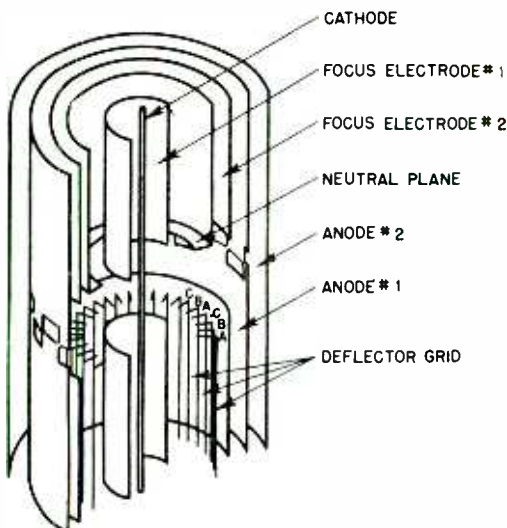
Fig. 35.—Phasitron tube 2H21.

speed such that they make one re-volution in one cycle of the three-phase supply divided by 12 (the number of sets of grid wires). Since the supply is obtained from

may be so positioned that they pass almost completely through the perfo-rations, whereupon anode #2 collects all the space current; or they may be so positioned that the two anodes

(Courtesy G.E. Co.)

Fig. 36. — Simplified internal view of the GL-2H21 phasitron tube.

the 230-kc oscillator, there will be

$$\frac{230000}{12} = 19,167 \text{ revolutions per sec.}$$

Observe the two sets of twelve perforations each, in anode #1, Fig. 36. Normally the unruffled sheet of electrons penetrates both sets about equally; i.e., it falls on a circle coinciding with the bottom of the upper set of perfo-rations and the tops of the lower set. Hence, anodes #1 and #2 collect about equal amounts of space current.

If the sheet is ruffled, the pleats may be in such a position as to strike anode #1 where it is not perforated, whereupon it collects all the space current; or the pleats
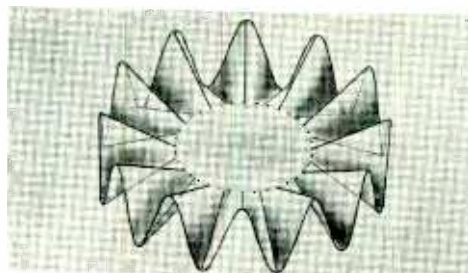
share the space current in any ratio whatsoever.

Fig. 37. — Ruffling of radial sheet of electrons by repelling action of one set of grid wires.

This is shown in Fig.38. When the ruffles are as indicated by the solid line, maximum current flows to is given an additional displacement in the direction of rotation of, say, 10° by some additional means. The
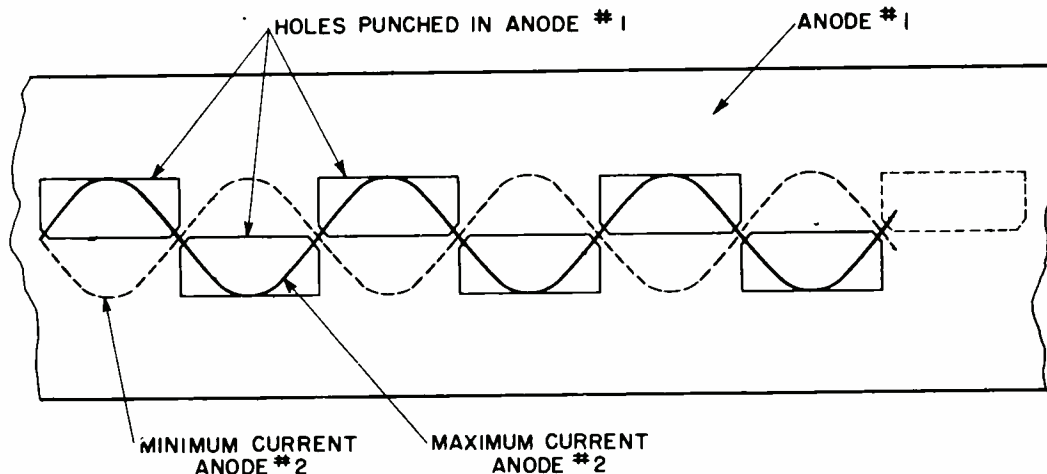

Fig. 38.—Alignment of ruffled disc of electrons with the perforations in anode #1.

anode #2 and minimum current to anode #1; when they are as indicated by the dotted line, minimum current flows to anode #2 and maximum to anode #1. When three-phase voltages are applied to the three sets of grids, the ruffles appear to rotate and take on the positions indicated, as well as intermediate positions, whereupon the current to each anode fluctuates between a maximum and minimum value, and therefore has an a-c component, and furthermore, the a-c component to one anode is 180° out of phase with that to the other. Since the ruffled disc makes 1/12 revolutions in one r-f cycle, and since there are 12 sets of perforated holes, so that one pair spans 1/12 of the circumference, the anode currents will fluctuate at exactly the frequency of the crystal oscillator.

Now suppose the ruffled sheet currents to the two anodes will now reach their maximum and minimum values 10° × 12 = 120° earlier in the r-f cycle; i.e., they will have received a 120° leading phase shift. (The factor 12 arises from the fact that 1/12 of a revolution or 30° of the circumference shifts the ruffles one set of perforations and hence one complete cycle of 360°, so that 30° of the circumference corresponds to 12 × 30 = 360° electrical.)

The method employed is that of magnetic deflection. The audio amplifier feeds current into a magnet coil surrounding the tube. The coil sets up an axial field that is perpendicular to the electron sheet, and proportional to the audio current. According to Fleming's rule, the electrons are deflected at right angles both to the radial flow and to the direction of the flux; i.e.,

along the tangent.

This is illustrated in Fig. 39. Since electron current flow is involved, rather than conventional current flow, the right-hand rather than the left-hand must be employed in applying the motor rule. For the directions of flux and electron flow shown, the deflection is clockwise along the tangent perpendicular to the particular radius.
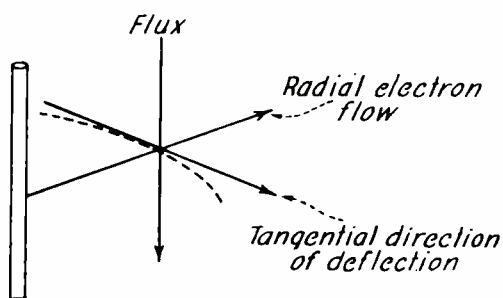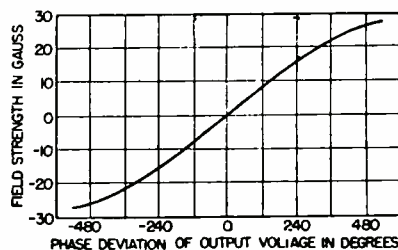

Fig. 39. —Relation between directions of electron flow, flux, and deflection.

The tangential motion, reacting with the axial magnetic field, in turn produces a force on the electrons. This force, if the motor rule is applied, acts *radially inward* on the electrons, thus reducing their radially outward velocity. This, combined with the tangential deflection gives rise to a circular motion around the cathode, which is a true rotation, and hence corresponds to an angular deviation or phase shift.

In Fig. 40 is shown the relation between magnetic field strength in guass and phase deviation of the output voltage in degrees. It will be observed that approximately one radian phase shift is obtained per

2.7 gauss change in magnetic field, and that the curve is essentially linear over a range from -360° to +360°. As stated previously, normally a 3.5 radian phase shift is the maximum employed, and the distortion is only at most 1.5 per cent.

Fig. 40. —Relation between phase deviation of output voltage and magnetic field strength.

The magnetic flux is concentrated in the region of the second focusing electrodes. The latter are made of magnetic material—siliconnicaloi, of low hysteresis loss— are conical shaped, and placed one above and one below the electron sheet so as to concentrate the flux through the sheet in the air gap between them. The anodes are made of nonmagnetic material to prevent their acting as a short-circuiting shunt for the flux.

The conversion of the crystal-oscillator output into three phase is accomplished in a relatively simple manner. As indicated in Fig. 41, a series R and C connected across a course of voltage E_a, draws a current i that leads E_a by an angle θ depending upon the relative values of R and C. The voltage drop across R is iR, and is in phase with i; the drop across C is $iX_c = E_o$,
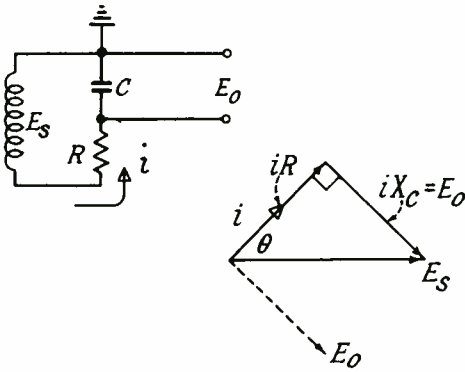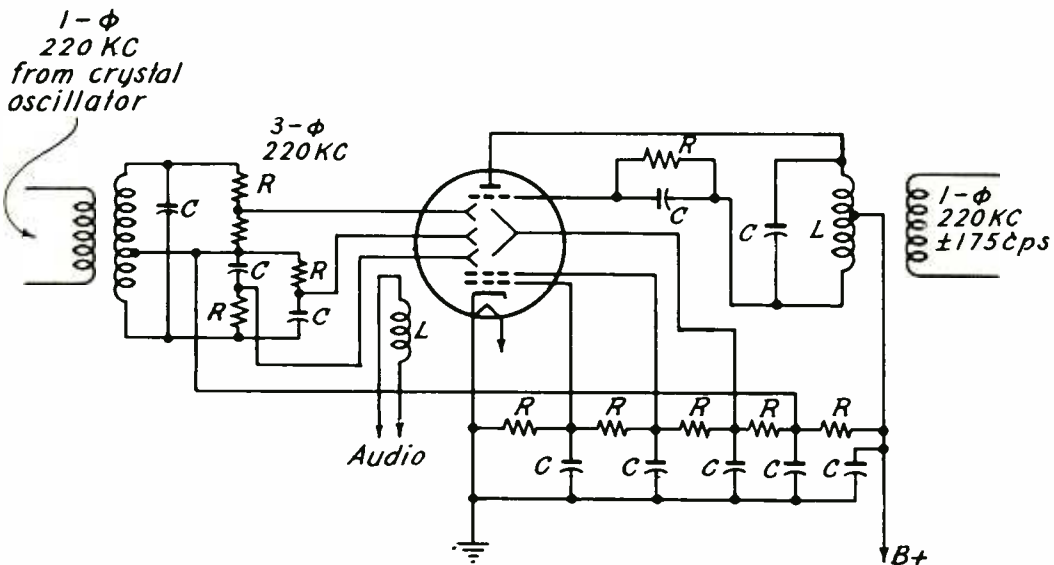
and is 90° lagging iR.

Fig. 41. —Simple phase shift network.

Hence, by choosing the circuit elements properly, three voltages of equal magnitudes, but 120° displaced in phase can be obtained. They are shown at the left in Fig. 42, as

applied to the phasitron tube's grids. Note the two R-C combinations connected across the lower half of the secondary of the transformer fed by the crystal oscillator. All three voltages are measured with respect to the center tap. Thus, the voltage across the resistor is a voltage that leads the lower secondary voltage by 60°; the voltage across the capacitor lags the secondary voltage by 60° (by suitably choosing each R and C); and the voltage across the top half of the secondary, and the tap on the resistance voltage divider, are 180° out of phase with respect to the lower secondary voltage, *when measured with respect to the center tap.*

The result is the three phase voltages of Fig. 43. Each is 120° out of phase with either of the others. When applied to the three deflector grids, they produce the ruffles in the electron sheet that

Fig. 42. —Basic GL-2H21 phasitron modulator circuit,

rotate at synchronous speed.

The phase shift that results when current flows through the magnet coil produces phase modulation. In order to produce frequency modulation, the magnet field strength, hence current producing it, must vary inversely as the audio frequency.
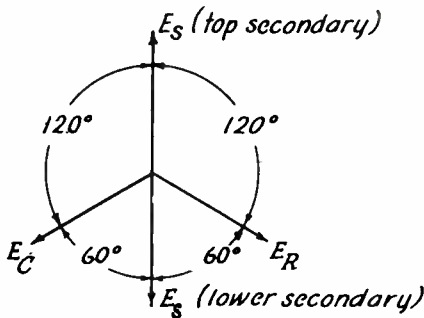

Fig. 43.--Vector relations of the three phase voltages.

This is atuomatically accomplished if the electrical impedance of the magnetic coil and associated circuit is essentially inductive in nature. Thus, suppose that the impedance of the audio output transformer (including the R_p of the output tube or tubes as reflected through the transformer), the resistance of the magnet coil, and its inductance, L, are each of such magnitude that the inductance of the magnet coil predominates. In this case the impedance is essentially an inductive reactance $2\pi fL$, where f is the audio frequency.

If the audio amplifier is flat, then the voltage developed in the magnet coil circuit will be of constant amplitude e_a over the audio range. The current flowing in the circuit will therefore be

$$i_a = e_a / 2\pi fL \qquad (22)$$

Since all factors in the right-hand side of Eq. (22) are constant except f, it is clear that i_a will vary inversely with f.

The magnetic flux ϕ_a set up by the coil is in direct proportion to i_a, hence ϕ_a will vary inversely with f. As a consequence, the phase shift will vary inversely with f, and thus give rise to true frequency modulation, in which the frequency shift or deviation, rather than the phase shift, is in proportion to the magnitude of e_a. In short, so long as the inductive reactance of the magnet coil is the predominating impedance in the circuit, the phase shift will inherently vary inversely as the audio frequency, and thus give rise to f.m.

To take care of the small resistive components, the frequency characteristic of the audio output voltage can be suitably modified. This is readily accomplished by altering the frequency characteristic of the inverse feedback employed in the amplifier so as to feed back less degenerative voltage at the lower frequencies.

This permits the gain to rise slightly at the lower end of the spectrum, where the inductive reactance of the coil is inherently lower and the resistive component therefore relatively more noticeable. A parallel R-C combination placed in series in the feedback path, and having a time constant R-C equal to the time constant L/R of the coil, produces the desired compensation.

It is to be noted from Fig. 42 that the output tank circuit is connected between the perforated anode

and the target surrounding it. It will be recalled that electrons that miss the perforated anode by passing through its holes, strike instead the outer target. The total space current remains constant, hence if more current flows to the target, less flows to the perforated anode, and vice versa.

This means that a current rise or positive half cycle for one electrode is accompanied by a current drop or negative half cycle for the other electrode. The two electrodes therefore behave like a push—pull circuit, and an inspection of Fig.42 will show that the tank coil is fed the plate-supply voltage through its center tap, just as in any other push-pull arrangement.

Preemphasis (of the highs) is an effect super imposed on the inverse frequency relationship, and is readily obtained by the use of a network in an earlier stage of the audio amplifier. One important precaution in this system is to exclude stray magnetic fields from the phasitron tube, particularly those from 60-cycle transformers. This is accomplished by surrounding the tube and coil in a high-permeability shield, and in addition, feeding the heater of the tube with d.c. obtained by rectifying 60-cycle a.c. by means of a selenium rectifier.

The output of the smallest transmitter unit is 250 watts, furnished by two Eimac 4-250-A tetrodes operated in push-pull. Higher outputs of 1 kw, 3 kw, 10 kw, and even 50 kw are readily obtained by using the basic 250-watt unit to drive the larger power amplifiers. The latter are Class C units of high efficiency.

This points out an advantage of the f-m transmitter over the a-m transmitter that was mentioned

previously; namely, since the amplitude of the f-m wave is at all times constant, the amplifier stage can be operated Class C at maximum output and efficiency, just like a telegraph stage in the keyed-down position. Thus, efficiencies in excess of 70 per cent are readily obtainable; this is to be contrasted with an efficiency of 50 to 60 per cent at full output for a linear amplifier.

THE W. E. TRANSMITTER. — Another solution to the problem of obtaining adequate frequency stability in an f-m transmitter is that employed by the Western Electric Company. This transmitter uses the reactance tube type of direct frequency modulation, with stabilization of the carrier frequency by comparison with a crystal oscillator.

A typical circuit is shown in Fig. 44. Tubes V3 and V4 constitute a push-pull oscillator circuit, whose tank circuit is composed of inductance L1 and capacitors C1 and C2 (each in two sections, as shown). The oscillator grids are cross-connected across L1 so as to obtain regenerative feedback and oscillation. A balanced or push-pull modulator stage is employed, and consists of reactance tubes V1 and V2, whose plates are connected across the tank coil L1. The grids of these two tubes are fed oscillator voltage essentially between the common cathodes and the center tap of the secondary of the modulation transformer T1 (actually the tap between the two capacitors connected across the secondary). Thus, the two grids are fed in parallel from the oscillator.

The exact path involves secondary coil L2 coupled to the tank coil L1, and pi network N1. The latter
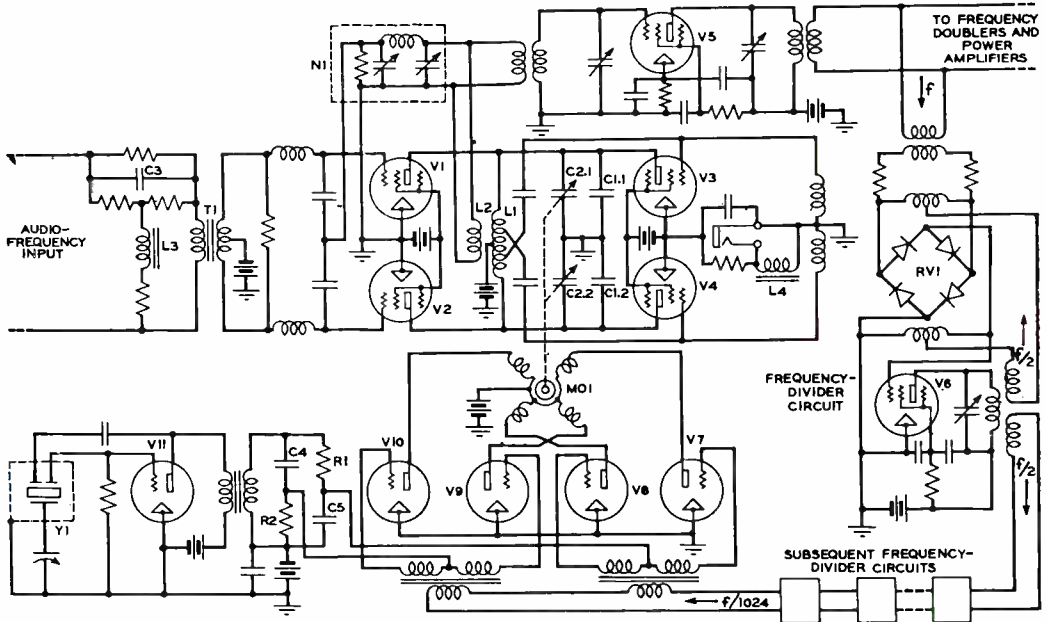
can be adjusted so that the phase of the voltage at the left-hand terminals can be shifted with respect to that at the right-hand terminals by the desired amount, this being such that the resulting r-f components of the plate currents are exactly 90° out of phase with the oscillator currents in L1.

The r-f components of the plate currents of the two reactance tubes are in phase, because the grid excitations via L2 and N1 are in phase. The two currents flow, however, from opposite ends of L1 to its center tap, therefore they cancel out. However, the grids of V1 and V2 are also connected in push-pull to the audio source via modulation transformer T1.

When modulation is applied, during one half cycle the grid of V1 is driven less negative, and the grid of V2 is driven more negative than their initial common value of bias. As a result, plate current from V1 exceeds that from V2; the two no longer cancel their effects in L1, and their difference represents a reactive current flowing in L1. Assume this net current is leading; if so, it cancels some of the lagging current drawn by L1, thus making it appear to have a higher reactance since it draws less net current. This is equivalent to increasing the inductance of the tank coil, and the oscillating frequency is therefore lowered.

On the other half cycle the reactive current flow is reversed, hence it appears as if the tank coil inductance were lower, and the frequency is therefore increased. The

Fig. 44. —Modulation and carrier-frequency stabilizing circuit of a Western Electric f.m. transmitter.

frequency variation is in proportion to the amplitude of the modulating input voltage, and thus frequency modulation of the oscillator is obtained, in which the frequency deviation is proportional to the strength of the modulating voltage.

It is important that the reactance tube plate currents (a-c components) be exactly 90° out of phase with the tank voltage, or otherwise the reactance tubes will appear as a combination of resistance and reactance to the oscillator. The significance of this is that during alternate half cycles the resistance and reactance change sign; the inductance reactance jX will become a capacitive reactance -jX, and the positive (damping) resistance R will become a negative (regenerative) resistance -R.

The change in sign in R will produce amplitude modulation in the oscillator output; this in turn means that the tank current will vary in amplitude at the modulation frequency, and in conjunction with the modulator current, will produce unequal positive and negative frequency deviations; i.e., asymmetrical and therefore distorted frequency modulation. This is obviated by the phase-shift network N1, which can be adjusted to make the reactance tubes appear as a pure reactance.

Since subsequent limiting will not prevent the asymmetrical frequency modulation already incurred in the oscillator stage, it is important to suppress it at that point. Any residual amplitude modulation is therefore reduced to a minimum by degenerative feedback. This is accomplished by inserting L4 in the common or mid-branch portion of the grid circuit. The cross-connected grid taps on L1 produce the regener-

ative push-pull feedback necessary for oscillation in the center-tapped grid coil. Amplitude variations, *which occur in phase* on both oscillator grids, cause a common current to flow through L4; the resulting r-f voltage drop is of the proper phase to counteract the amplitude modulation producing it.

At the a-f input, there will be noted a bridged-tee network consisting of C3, L3, and associated resistors. Since L3 is the shunt arm, it shunts the audio waves less and less as their frequency is increased; at the same time the reactance of C3 decreases and permits their transmission more readily to the primary of T1. Thus, the audio frequency response has a rising characteristic; the unit is the pre-emphasis network in the system. A proper choice of L3, C3, and the resistors makes the transmission characteristic sijilar to that of the impedance of a series L-R circuit whose time constant is 75 μsec.

The salient feature of the method of carrier frequency stabilization is that by demultiplying the f-m wave to a point where the deviation is less than the lowest audio frequency, all side bands except the first order are eliminated, and the carrier is so emphasized, that it can easily be compared with a crystal oscillator in order to detect and correct any drift in the original carrier. This will be explained in greater detail very shortly.

Another important characteristic is that the correction for drift is not applied as a bias change to the grid of the reactance tube, which would thus use up some of the range for modulation, but instead is applied directly to the tank circuit

capacitor. As a result, the reactance tube is free to perform the sole task of producing frequency modulation, and can be adjusted for optimum performance.

It will be recalled from a previous assignment that there are two quantities involved in f.m.: the modulation constant k_f, which is the ratio of the frequency deviation to the carrier frequency; and the modulation factor m_f, which is the ratio of the frequency deviation to the modulation frequency. In symbols.

$$k_f = f_D/f_c \qquad (23)$$

and

$$m_f = f_D/f_m \qquad (24)$$

where f_D is the frequency deviation, f_c is the carrier frequency, and f_m is the modulation frequency.

It will further be recalled that when a carrier is frequency modulated by even a single sinusoidal modulating frequency, an infinite number of orders of side bands are produced. For example, if a 100-mc carrier is modulated by a 1,000-cycle note, a first-order pair of side bands is produced 1,000 cycles on either side of the carrier frequency; a second-order pair of side bands is produced 2,000 cycles on either side of the carrier frequency; a third-order pair is produced 3,000 cycles on either side of the carrier frequency, and so on. In actual practice only a limited number of these orders is important from an amplitude and hence energy viewpoint, so that but a finite band width is actually necessary to transmit this frequency-modulated wave.

The amplitudes of the various components of the f-m wave are given by certain infinite series known as Bessel functions of the first kind, of the quantity m_f. Thus the amplitude of the carrier is given by $J_o(m_f)$, where $J_o(m_f)$ is a Bessel function of the first kind and zero order of the argument m_f. The amplitudes of the first order side bands are given by $J_1(m_f)$, where $J_1(m_f)$ is a Bessel function of the first kind and first order of the argument m_f, and so on.

Each Bessel function varies with m_f somewhat like a damped sinusoidal wave. This means that as m_f changes, the various J's vary from positive to negative values through zero, and have progressively smaller peak amplitudes as m_f increases.

But m_f is the ratio of f_D/f_m and is independent of the value of carrier frequency. When $f_D = 2.4 f_m$, or $m_f = 2.4$, $J_o(2.4) = 0$, or the amplitude of the carrier just equals zero. For other values of m_f, the amplitudes of one or other orders of side bands may become zero; this depends solely upon the frequency f_m and loudness of the modulating wave (which determines f_D).

When m_f is very small (less than unity, which means f_D is less than f_m), $J_o(m_f)$ is nearly unity, and the higher order J's are very small and nearly equal to zero. This means that the carrier amplitude is a maximum, and the side band amplitudes are very small. On the other hand, if m_f is very large (such as due to a loud low-frequency audio tone) J_o is smaller (and may be zero for some particular value of m_f), and the side-band amplitudes are fairly large.

The significance of this is that the final frequency-modulated wave normally has most of the energy
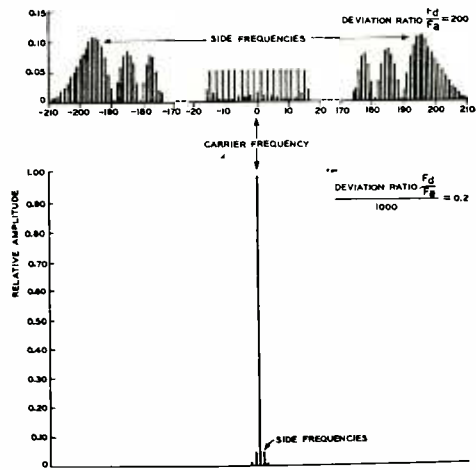
in the side bands, and very little
in the carrier, since the monitor
man in the studio usually strives to
obtain a maximum modulation of the
wave. It is therefore difficult to
select such a weak carrier amplitude
for comparison with a crystal-
controlled oscillator to correct for
drift.

Suppose, however, that the f-m
wave is demultiplied in frequency by
suitable circuits. If f_c is halved,
f_D is also halved, *but f_m remains
unchanged.* This means that m_f is
halved. Sufficient demultiplication,
therefore, can reduce m_f to a value
less than unity.

Under these conditions, the
carrier amplitude is high, and the
side-band amplitudes are low, as has
just been pointed out. Indeed, the
low-frequency carrier merely oscil-
lates in phase by a few degrees, and
is therefore accompanied by very
weak side bands. Such a wave is
easily compared with a crystal-
oscillator for drift correction.
This is illustrated in Fig.45, which
shows the relative amplitudes of the
side bands and carrier for the
initial high value of m_f (at the
top), and the relative amplitudes
after frequency demultiplication,
when m_f is reduced to less than
unity (bottom figure). The relative
increase in the carrier amplitude is
strikingly great.

The process of comparison is as
follows: The demultiplied carrier
and the crystal oscillator output
are fed into a mixer tube whose out-
put is at the difference beat fre-
quency. By also feeding a 90° dis-
placed wave from the oscillator two
output voltages are obtained, 90°
out of phase, and hence constituting
a two-phase supply. This is ampli-
fied, and connected to a synchronous

motor. The latter then turns the
plates of an air-type capacitor
connected across the tank circuit.

Fig.45.—Spectrum of a frequency-
modulated wave before and after the
wave is passed through a system of
frequency dividers.

If the demultiplied carrier is
lower in frequency than that of the
crystal oscillator by 100 c.p.s.,
for example, the beat frequency will
be 100 c.p.s. As a two-phase supply
it will actuate the motor, causing
it to revolve the capacitor plates
in the proper direction to raise the
carrier frequency, thus bringing its
demultiplied value up to that of the
crystal oscillator, whereupon the
beat frequency drops to zero, and
the motor stops revolving.

If the carrier frequency drifts
100 c.p.s. above the oscillator, the
beat frequency is again 100 c.p.s.,
but one voltage *reverses* in phase.
The motor now revolves in the oppo-
site direction, and lowers the
carrier frequency until it equals

that of the oscillator. In this way, any continuous small drift of the carrier from the oscillator frequency produces a rotating field that actuates the motor and brings the carrier back to the oscillator in frequency.

At the extreme right of Fig. 44 will be found a frequency-divider circuit. Its action is explained in the appendix. This divides the frequency in the ratio of 2 to 1. Ten such stages produce a total frequency division of $f/1,024$. The resulting low-frequency accentuated carrier is fed to the primaries of two transformers in series.

At the same time, the output of a crystal-controlled oscillator tube V11 is fed through a transformer and 2-phase network (similar to the 3-phase network employed in conjunction with the phasitron tube) to the center taps of the secondaries of these same two transformers.

As a result, the grids of V10 and V9 receive the demultiplied carrier frequency in push-pull arrangement, whereas they receive the oscillator voltage in phase via the common center tap. The two frequencies are mixed in this stage to produce, among other frequencies, the difference beat between the two. An inspection of the circuit shows it to be essentially a balanced modulator similar to that employed in the Armstrong system; as a result, only the difference beat frequency appears in the output.

The same considerations hold for the other transformer and tubes V8 and V7. Here, however, the oscillator frequency is injected 90° out of phase with that for V10 and V9, whereas the demultiplied carrier is applied in phase. As a result, the difference beat output from the

plates of V8 and V7 is 90° out of phase with that from V10 and V9.

The two coils of the motor therefore receive two-phase excitation, and the rotating field causes the rotor to revolve (except, of course, if the beat frequency is zero). The motor is mechanically connected to tank capacitor C2, and can therefore alter the oscillator frequency.

As an illustrative example, suppose the oscillator frequency is 4.9 mc, in order that when it is multiplied 18 times, it produces a final carrier of 88.1 mc. Assume ±75 kc maximum deviation for the latter. Then the oscillator frequency must deviate $\pm 75,000/18 = \pm 4,170$ c.p.s. about the 4.9 mc value.

The oscillator frequency is demultiplied by 1,024 to a value $4,900,000/1,024 = 4,780$ c.p.s. This, then, is also the frequency of the crystal-controlled oscillator. The reason for such a low final comparison value of frequency is that the difference beat will be very low in frequency, so that the motor will be able to respond to it without having to race first in one direction, and then the other, etc.

The deviation is demultiplied to $\pm 4,170/1,024 = \pm 4.07$ c.p.s. At 30 c.p.s. audio frequency, this gives rise to a phase variation in the 4,780-cycle low-frequency carrier of $\pm 4.07/30 = \pm 0.136$ radians or ±7.8°. The inertia of the motor will prevent its rotor and associated load from oscillating through this angle 30 times a second. On the other hand, a continuous drift in frequency in one direction (rather than a back-and-forth deviation) will produce an ever increasing phase angle or difference beat frequency which the motor can follow,

thereby correcting the drift.

Note that the above rotation is determined by the magnitude of the *beat frequency*, and *not* by the amplitude of the *beat voltage*. Thus variations in circuit gain do not affect the frequency correcting device except, to some extent, its sensitivity. A further point is that any variation in the frequency-determining elements of the modulated oscillator circuit are mechanically corrected by the motor and associated capacitor, even if the variation tends to produce a frequency drift of several megacycles in the final output carrier, and this correction does not disturb the adjustment of bias voltages, etc., of the tubes, so that they operate on the optimum portion of their characteristics. In addition, even if the motor control circuit should fail, the program continuity will not be interrupted, since the modulator is separate from the control circuit.

Western Electric f-m transmitters are available in 250 watts, 1 kw, 5 kw, and 50 kw. Between the frequency multipliers and the power amplifier stage is an isolating amplifier; the smallest power amplifier stage is 250 watts. This, in turn, is capable of driving the 1-kw and 5-kw power amplifiers, if more output is desired. Where 50 kw is required, the 5-kw stage is used in turn to drive it. It is to be noted that f.m. lends itself particularly well to ultra-high-frequency operation in that the power output is constant at all times, whereas in a.m. the peak power, corresponding to 100 per cent modulation, is four times the carrier power.

As a result, in a.m. much larger tubes are required than in f.m., and larger tubes are not only more expensive, but difficult to design at the higher frequencies. The lower peak rating required by f.m. makes the design of u-h-f tubes a much less difficult problem.

CONCLUSION

This concludes the assignment on modulation. The first section dealt with amplitude modulation, and discussed the various methods of modulating the carrier. These are:

1. Plate modulation.
2. Control grid modulation.
3. Screen grid modulation.
4. Suppressor grid modulation.
5. Cathode modulation.

In the case of plate modulation, the audio power required is high, but the load resistance that the r-f stage presents to the modulator stage is practically constant, and therefore tends to minimize distortion.

In the case of grid modulation, the audio power requirements are far less, but on the other hand the load impedance presented by the grid of the power tube or tubes is variable over the audio cycle, and tends to flatten the positive audio peaks. Design methods are given to minimize this form of distortion, which is principally second-harmonic distortion for a sinusoidal modulating wave.

Screen-grid modulation is used mainly in conjunction with plate modulation in the case of a tetrode and pentode tube. Modulating voltage for the screen grid can be obtained from a tap on the plate-modulation transformer, or by means of an audio impedance interposed in

the screen circuit (in the case of some tetrode tubes), in which case the screen modulation voltage is developed by means of an audio voltage drop in the audio impedance owing to the audio screen current flowing through it.

Suppressor-grid and cathode modulation are used occasionally, but have no particular advantage over control-grid and plate modulation. Indeed, the latter is the preferred method, especially in high-level modulation systems. An alternative is the Doherty high-efficiency linear amplifier used by Western Electric, and similar systems proposed by others.

The second section dealt with frequency modulation. The theory has been presented in a previous assignment; in the present assignment three systems are discussed. The first is the original Armstrong system. The main disadvantage of this system is the requirement of a heterodyne as well as an initial crystal-controlled oscillator; the stringent frequency-stability requirements impose rather difficult restrictions on the stability of each oscillator. A later improved method due to Armstrong is described in the Specialized Broadcast Section.

A second system is that developed by G. E. and employing a special phasitron tube. This tube affords a larger initial phase shift, and therefore requires less frequency multiplication. As a result, no heterodyne crystal-controlled oscillator is required, and constancy of frequency is more readily obtained.

A third system is that developed by the Bell Laboratories and used in the Western Electric f-m transmitter. In this case the initial oscillator is modulated by a reactance tube stage. Its output is then multiplied to the desired carrier frequency and with the required frequency deviation.

For frequency control, the frequency of the initial oscillator is demultiplied down to a value where essentially the deviation is negligibly small, and only the subcarrier remains. This is then compared with a crystal-controlled osciloator; any drift causes a difference beat frequency between the two, and this beat frequency, consisting of two voltages 90° out of phase, is applied to a synchronous motor, which thereupon revolves an auxiliary tuning capacitor in the initial oscillator's tank circuit, bringing the drift back to zero. Frequency control is therefore mechanical in nature and is independent of the frequency-modulating reactance-tube stage. As a result, the latter can be adjusted for optimum bias with respect to maximum linear frequency deviation range, without requiring that any part of this range be consumed in correcting for frequency drift.

APPENDIX I

FREQUENCY-DIVIDING CIRCUITS

The frequency-dividing circuit has some very interesting features. In order to understand its action, it is necessary to review the phenomenon of feedback from a broader viewpoint. In Fig. 1 is
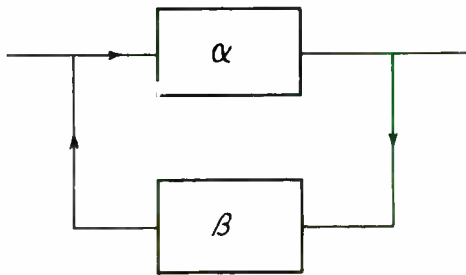

Fig. 1.—Generalized block diagram of a feedback circuit.

shown a generalized block diagram of a feedback circuit. The unit marked α is some kind of an amplifier having an amplification or gain of α times from the left-hand to the right-hand terminals.

The unit marked β feeds a fraction β of the output back to the input terminals. Both α and β may be complex quantities; i.e., they may introduce a phase shift as well as a gain or loss between their input and output terminals.

If the fraction of the output that is fed back is equal to the input necessary to produce that output, and is, moreover. *in phase with the input voltage,* then there is no need to supply the input voltage from an external source; the circuit itself is capable of supplying its own input, and therefore can maintain its operation by it-

self. In short, it can oscillate.

From a most general viewpoint, either α or β may be an amplifier or "losser" circuit; so long as the previous condition holds, the circuit will oscillate. In particular, if α is a losser circuit, and β is an amplifier circuit, the level of the signal at the right-hand terminals will be lower than that at the left-hand terminals, but the loss in α will be compensated for by the gain in β in this unconventional case, and the circuit as a whole will oscillate.

From this the following generalization is apparent:

If the gain around the feedback loop is unity, and the phase shift is zero, or 360°, or an integer multiple of 360°, the circuit will oscillate at a stable and constant amplitude.

If the gain is greater than unity, the amplitude of the oscillations will continue to increase; in a practical equipment, a point is reached in which the equipment breaks down, or simply overloads, in which case the gain drops to unity, and the circuit continues to oscillate from then on at this amplitude.

Consider now a more unorthodox feedback system in which the unit α is a modulator, and the unit β is a frequency multiplier or harmonic generator. The modulator α has the property of producing summation and difference beat frequencies when two voltages are applied to it.

Suppose, for example, that a voltage of a frequency of 1,000 cycles, and also one of 800 cycles, are applied to the modulator. There will appear in its output (in addition to the 1,000- and 800-cycle voltages) a difference beat of

1,000 - 800 = 200 c.p.s., and a summation beat of 1,000 + 800 = 1,800 c.p.s.

Now suppose that the output is fed into a resonant circuit tuned to 200 c.p.s. Then only the difference beat frequency voltage of 200 c.p.s. will appear across the output of this tuned circuit. If this 200-cycle voltage is then impressed upon the frequency multiplier, harmonics or multiples of 200 cycles will appear across its output. By connecting a suitably tuned circuit across the frequency multiplier, any desired harmonic can be selected.

Suppose the fourth harmonic is chosen. Its frequency is 4 x 200 = 800 c.p.s.; this is the resonant frequency of the tuned circuit connected to the output terminals of the frequency multiplier. If this is impressed upon the modulator, in conjunction with a 1,000-cycle voltage from an external source, then a difference beat frequency of 200 c.p.s. will be produced; this upon passing through the frequency multiplier, will develop an 800-cycle voltage; the latter will react in the modulator with the externally applied 1,000-cycle voltage to produce the 200-cycle beat output; and the action will continue indefinitely so *long as the 1,000-cycle voltage is applied.*

This is shown in Fig. 2. The incoming frequency at the left is 1,000 c.p.s. If, at the start, an 800-cycle voltage is also obtained by some means and impressed on the input, a difference beat frequency of 200 c.p.s. is obtained at the right-hand side of the modulator and 200-cycle tuned circuit.

A portion of the 200-cycle output can then be fed into the frequency multiplier, and the fourth harmonic selected by the 800-cycle tuned circuit and fed into the modulator. The operation then continues as long as the 1,000 c.p.s.
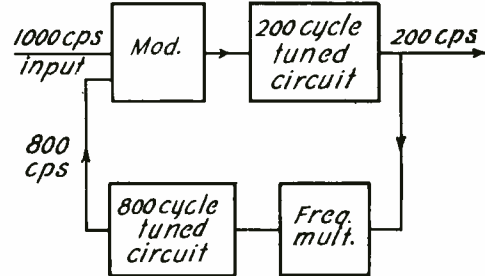

Fig. 2.—A combination modulator frequency-multiplier feedback arrangement provides a submultiple of the frequency input.

voltage is impressed at the left-hand terminals, and its fifth submultiple, or 200 c.p.s. can be obtained at the right-hand terminals.

The action, when the circuit is first put into operation, by which an 800-cycle voltage is obtained is essentially the same as that which causes an oscillator to start oscillating when the switch is first closed. Any thermal noise or other disturbance that has enough energy in its 800-cycle component will initiate the action; thereafter it is self-sustaining.

In the above example numerical values were used, such as an input of 1,000 c.p.s., a desired submultiple of 1/5 to give a 200 c.p.s. output, and a frequency multiplication of 4 in the frequency multiplier to give an 800-cycle voltage. In the general case, let f be the frequency of the input voltage, and n be the desired submultiple. The output frequency is then simply f/n. In the example f = 1,000 c.p.s., n = 5, and the output frequency is 1,000/5 = 200 c.p.s.

The multiplying action of the frequency multiplier is then $(n - 1)$ times, so that its output frequency is $f(n - 1)/n$. In the example, this has the value of $1,000(5 - 1)/5 = 800$ c.p.s. The modulator produces a difference beat frequency equal to

$$f - f'\left(\frac{n - 1}{n}\right) = \frac{nf - nf + f}{n} = \frac{f}{n}$$

which is obtained at the output. The above general relationships are seen to be consistent with a continuous self-sustaining operation.

This type of circuit has certain advantages over the ordinary multivibrator. In the first place, it has no free-running frequency like the multivibrator, and when the input frequency f fails, it ceases to operate, thus serving notice of failure of operation.

In the second place, it appears less sensitive to adjustment, and less apt to skip to some other submultiple, particularly if the submultiple is low, although the multivibrator is fairly stable under these conditions.

A third advantage is that for a given tube, it is capable of operating at a higher frequency than the multivibrator, because it generates essentially a sine wave, and thus does not require a broad band in its amplifier stages, as does the multivibrator, which generates essentially a square wave requiring the amplification of a broad band of harmonics.

Where only a 2 : 1 frequency division is desired, the circuit becomes much simpler. In this case $n = 2$, so that the harmonic generator β has to produce a frequency

$$\frac{n - 1}{n} f = \frac{2 - 1}{2} f = f/2$$

or the same as the output of the a stage. This means that no harmonic generator stage β is required. Indeed, instead of an a or modulator vacuum tube stage, a dry-disc rectifier can be used. However, since the difference beat output is low, and insufficient to sustain itself, additional gain is required in the form of a vacuum tube amplifier stage following the dry-disc modulator. The circuit then has the appearance of that shown in the right-hand side of Fig. 42.

The action of a bridge-circuit dry-disc modulator is best understood by reference to Fig. 3 and
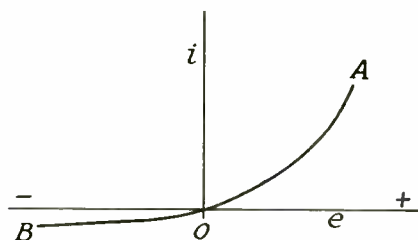

Fig. 3.—Typical current-voltage characteristic of a dry-disc rectifier.

succeeding figures. In Fig. 3 is shown the relation between current and voltage in a typical dry-disc rectifier. For positive values of voltage e, branch OA is involved, and indicates that a large amount of current flows; i.e., the forward resistance is low. For negative values of e, branch OB is involved; in this direction very little current flows, and the backward resistance is high. For an ideal rectifier the current would be zero, and the resistance infinite.

An important characteristic of the other branch OA is its upward curvature. This indicates that the

current increases more than in proportion to the voltage, so that if two voltages are simultaneously impressed upon the rectifier in the forward direction, each affects the current flow of the other by changing the portion of the branch OA that is traversed. This factor produces modulation in the bridge-circuit arrangement of four rectifiers, as in Fig. 42.

alternation, current (dotted line) flows through CD and DA in series; AB and BC are inoperative. Note that R_1 and R_2 are directly across the generator, and hence have no effect on the rectifier currents.

If it be assumed that all rectifier elements have the same current-voltage characteristic, then for the first alternation, since the current is the same in AB as in BC,
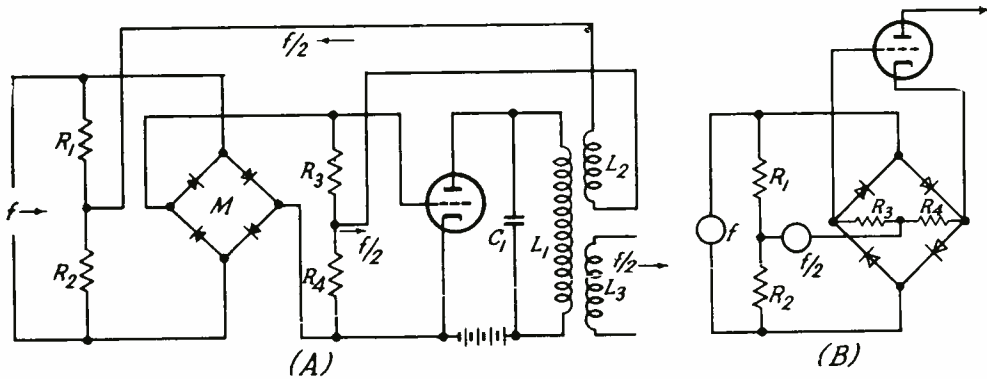

Fig. 4.—Frequency divider and simplified circuit.

In Fig. 4(A) is shown the frequency divider of Fig. 42, and in Fig. 4(B) is shown a simplified diagram of the modulator portion. The source of the incoming signal is denoted by f; the source of the half-frequency signal, actually the secondary coil L_2, is denoted by the generator f/2. The circuit L_1 C_1 is tuned to f/2, and thus the plate circuit of the tube selects and accentuates this frequency from all the other frequencies, such as f, 2f, f + f/2, that are applied to the grid by the modulator M.

In analyzing the action of the modulator, consider first input generator f. As shown in Fig. 5, during one half cycle or alternation, the current (solid line) flows through AB and BC in series; CD and DA are inoperative. During the next
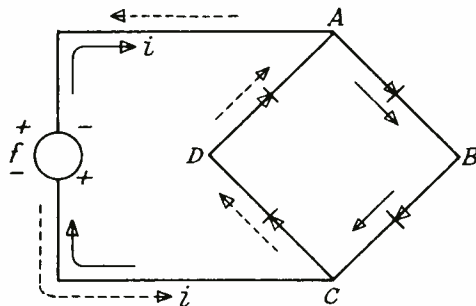
the resistance of each is the same, and point B is halfway between A and C in potential. Similarly, D is halfway between A and C in potential, since the reverse resistances of AD and DC are also equal.

During the next alternation, AD and AB, and DC and CB interchange

Fig. 5.—Currents through rectifier elements from generator f, alone.

their roles, but points D and B are still halfway in potential between A and C, and hence have no potential difference between them. Since the grid of the amplifier tube is connected to D, and the cathode to B, no signal appears between the two, and hence there is no output from the plate circuit.
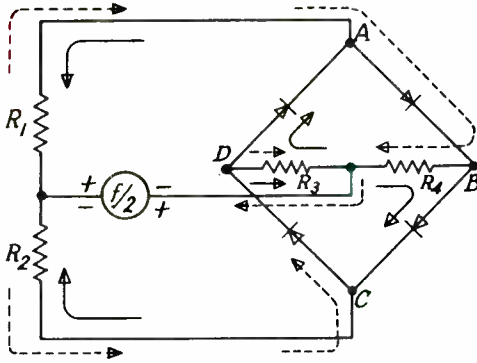

Fig. 6—Currents through rectifier elements from generator f/2, alone.

Next, consider the generator f/2 acting by itself. As shown in Fig. 6, during one alternation, current (solid line) flows through rectifier elements BC and DA; during the next alternation, currents (dotted line) flow through CD and AB. The currents are determined principally by the resistors R_1, R_2, R_3, and R_4, since these are very high compared to the forward resistances of the rectifer elements.

However, in the absence of generator f, the currents from generator f/2 encounter equal resistances in the two shunt paths for either alternation, hence the two currents are equal in R_3 and R_4, produce equal voltage drops, and therefore no voltage of frequency f/2 between D and B. Hence once again, so far as generator

f/2 is concerned, no voltage appears between the grid and cathode of the amplifier tube and there is no output.

When both generators are applied simultaneously, conditions are changed. Each generator, in forcing current through the rectifier elements, changes their forward resistance and thus tends to change the magnitude of the current flow of the other generator. However, in the case of generator f/2, any change in the forward resistance of the rectifier elements owing to current from generator f, does not materially change the magnitude of the currents of generator f/2, because, as mentioned previously, these current magnitudes are determined principally by linear resistors R_1, R_2, R_3, and R_4. Hence no appreciable voltage of frequency f/2 appears across the terminals D and B owing to generator f/2 even when generator f is also connected; this means that at no time is there any appreciable signal of this frequency applied to the amplifier tube *directly from generator f/2.*

The effect of generator f/2 on the current magnitudes of generator f, however, is very appreciable, because the forward resistance of the rectifier elements is the major resistance in generator f's circuit. To study the effects, combine Figs. 5 and 6, as has been done in Fig. 7(A) and (B). In Fig. 7(A) are shown conditions for one alternation of f; in (B), for the next alternation of f. For convenience, assume that f/2 changes so slowly that its current flow remains substantially unchanged during one cycle of f.

Note in Fig. 7(A) that rectifier element AB has both currents

flowing in it, whereas rectifier element BC has only current f flowing through it. The flow of current f/2 through DC has no effect on current f, since DC is an open circuit to the current direction of f.

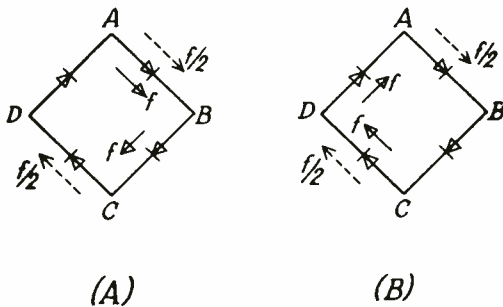Hence, D remains halfway in potential between A and C so far as current f is concerned. But

(A)   *(B)*

Fig. 7.—Combination of currents from generators f and f/2 in the rectifier elements.

rectifier element AB has less resistance than BC because more current is flowing through it; this makes B move closer to A, which is negative to C. Hence, B, to which the cathode of the amplifier

tube is connected, goes negative with respect to D, to which the grid is connected.

As a result, the plate current increases during this alternation of f. On the next alternation of f, the circuit is as in Fig. 7(B); now D moves closer to C in potential, and since C is now negative to A, D becomes more negative. On the other hand, B now remains halfway in potential between C and A. The result is that the grid of the amplifier tube goes negative with respect to the cathode, and as a result the plate current decreases.

This means that both alternations of f now appear in the output of the amplifier tube, whereas in the absence of current f/2, no such output appears. This is shown in Fig. 8(A), where the plate current of the amplifier tube is plotted against time. If current f/2 is zero, then the plate current contains no component f, and is pure d.c. For small values of f/2, the output contains a small amplitude of f in addition to the initial d.c.; for large values of f/2, the output contains a large amplitude of f in addition to the initial d.c.

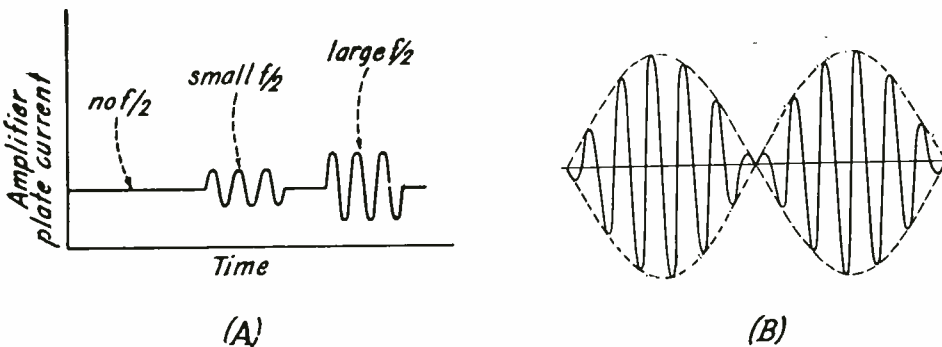If current f/2 varies in a

(A)   *(B)*

Fig. 8.—Plot of current f vs. time, for various values of current f/2.

sinusoidal manner, then the ampli-
tude or envelope of current f varies
as shown in Fig. 8(B). Note that
both half cycles produce exactly the
same shape envelope, but that the
phase of f abruptly reverses at the
end of each half cycle of f/2. The
appearance of the wave is exactly
the same as that shown in Fig. 28
for the Armstrong balanced modula-
tor, and indicates that the dry–disc
rectifier modulator is also of the
balanced type.

Hence, the output of the ampli-
fier tube does not contain the
carrier frequency f, but its two
side bands produced by interaction
with modulator current f/2. In
Fig. 8 the modulator current f/2
was assumed, for clarity in il-
lustrating, to be of much lower
frequency than f, so that many
cycles of the latter appear during
one cycle of the modulator current.
If the actual frequency of the
latter be taken into account, then
only two cycles of f will appear
during one cycle of f/2, but ex-
actly the same conclusion can be
drawn.

This is that only the two side
bands of f appear, of frequency
f + f/2 = 3f/2, and f - f/2 = f/2.

The tuned circuit $L_1 C_2$ of Fig. 4(A)
responds to the latter frequency
only, and passes it on to the two
secondaries L_2 and L_3. Note for
this particular pair of values of
frequency (f and f/2), that al-
though generator f/2 does not
directly produce an output voltage
from the modulator, it nevertheless
does so indirectly by cross–modu-
lating with f to produce a differ-
ence beat frequency f/2 of exactly
the same frequency as itself. For
higher ratios of demultiplication
the modulating frequency is

$$\frac{n-1}{n} f$$

and different from the difference
beat frequency f/n in the output.

The action can now be sum-
marized briefly as follows: The
input frequency f cross–modulates
with the output frequency f/2 to
produce two frequencies, 3f/2 and
f/2, which are applied to the grid
of the amplifier tube. The latter
amplifies f/2 to a magnitude suf-
ficient to furnish output, and al-
so to feed back sufficient energy
to cross–modulate with f and main-
tain the action, *so long as f is*
continually applied to the input
terminals.

MODULATION

EXAMINATION

1. (A) Which method of amplitude modulation requires the most modulating power?

   (B) Which method operates the radio-frequency stage at the highest efficiency?

2. (A) Explain the difference between low-level and high-level modulation.

   (B) In low-level modulation, how is large modulated r-f power output obtained?

   (C) The antenna meter reads 10 amperes during zero-modulation periods. If the reading during modulation rises to 12 amperes, what is the percentage modulation (assumed sinusoidal)?

3. An r-f Class C stage using a Type 805 tube is plate modulated. Under carrier conditions, the plate-supply voltage is 1,250 volts, and the d-c plate current is 160 ma.

   (A) What is the value of the resistance that this tube presents to the modulator stage?

   (B) What is the required modulator power?

4. Assume a pair of 809 tubes are employed in Class B push-pull to modulate the above 805 tube. The Tube Manual gives the following data for such a stage:

   E_{bb} = 750 volts   Peak a-f signal = 145 volts (grid-to-grid)
   E_c = -4.5 volts I_b = 40 ma at no signal (2 tubes)
                          I_b = 200 ma at full signal (2 tubes)

   Plate-to-plate load resistance = 8,400 ohms
   Grid driving power = 2.4 watts   Output power = 105 watts

   (A) Is the power output sufficient for 100 per cent modulation of the 805 tube?

   (B) What turns ratio must the modulation transformer have for correct impedance match?

MODULATION

5. In the problem in the text on grid modulation of an 803 tube, assume that 2 per cent distortion at 100 per cent modulation is permissible. Suppose a pair of 6F6 tubes are employed in Class AB$_2$ (grid current flows during part of cycle). The operating characteristics are as follows: E_{bb} = 350 volts; E_c = -38 volts; Positive grid swing = 20 volts for each tube; μ = 6.8; R_p = 2600 ohms per tube.

   (A) Calculate the turns ratio required in the modulation transformer.

   (B) Check to see if the modulator stage can deliver the required 50 volts for modulating the grid of the 803 tube.

6. Discuss the method of feeding modulation voltage to the screen of a tetrode and of a pentode tube that is being plate-modulated.

7. In the Armstrong system, the initial crystal-controlled oscillator has a frequency of 210 kc. This frequency is multiplied 128 times, and then heterodyned with a 22,830-kc crystal-controlled oscillator. The resulting beat frequency is then multiplied 22 times.

   (A) What is the final carrier frequency?

   (B) If the initial phase shift angle is .445 radians at a modulating frequency of 50 c.p.s., what is the corresponding frequency deviation of the final carrier?

8. Assume that the phasitron tube can undergo a maximum phase shift of 3.5 radians. The initial oscillator frequency is 206.25 kc, and the final desired carrier frequency is 89.1 mc.

   (A) What frequency multiplication is necessary?

   (B) What is the relusting frequency deviation if the modulating frequency is 50 c.p.s.?

9. (A) In the phase-shift systems of frequency modulation,

9. What means must be employed to obtain frequency rather than phase modulation?

    (B) How is this brought about in the G. E. phasitron tube system?

10. (A) In the Western Electric system, why is the frequency of the frequency-modulated initial oscillator demultiplied before it is compared with the crystal-controlled oscillator?

    (B) What further advantage as regards control of the synchronous motor is obtained by demultiplying the frequency?

    (C) What is the advantage of having the motor mechanically control the turning capacitor instead of the bias of the reactance tubes?