TRANSMITTER ADJUSTMENTS

By J. G. Sperling

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FCC Rules, Regulations, and Standards of Good Engineering Practice have required for many years that the audio frequency and radio frequency characteristics of AM broadcast transmitters shall be in accordance with certain prescribed limits. It may ordinarily be assumed that all stations operate within these limits. However, the operator of a broadcast station is quite often faced with the problem of diagnosing transmitter faults or failures which prevent operation in accordance with the standards set by FCC. The possible faults and failures, if listed, would fill many pages. Those which will be discussed here are: (1) Limited Audio Frequency Response; (2) High audio harmonic content; (3) Excessive second and/or third RF harmonic content; (4) Improper neutralization of the RF circuits; (5) Excessive hum on the carrier.

These difficulties or faults may be easily remedied as follows:

1. The use of high fidelity pre-, program, line, and modulator amplifiers will settle the question of audio fidelity in the a-f end. It should be noticed, however, that the amount of inductance of the modulation choke or other device connecting the modulator to the modulated amplifier materially determines the amount of low audio frequency response. For the response at 30 cycles to be equal to that at 1,000 cycles, it is necessary that the value of the above inductance be equal to the reflected load resistance of the class C modulated amplifier divided by at least 200, or $L = E_p/(I_p \times 200)$ where L is in henries, E_p is the plate voltage of the modulated amplifier and I_p is the plate current of the modulated amplifier.

The capacity of the plate by-pass condenser in the modulated amplifier and the following class B stages, if any, determines the amount of high a-f response. If the response at 10,000 cycles is to be the same as that at 1,000 cycles it is necessary that the capacity, in micro-microfarads, be equal to the load resistance divided by 100, or $C = E_p/I_p \times 100$.

2. The audio harmonic content for a properly adjusted class A audio amplifier is negligible. The only precaution to watch is to see that the amplifier is not overloaded and that all the vacuum tubes are in good operating condition.

If class B audio amplifiers are used, the use of proper tubes in a correctly adjusted circuit will usually result in a third harmonic content of not over 2 per cent.

The use of a suitable KVA. to KW. ratio in the r-f amplifiers and proper adjustment of all circuits will result in audio distortion of not over 1 per cent. This is discussed more thoroughly further on.

3. The r-f harmonics present can be greatly attenuated, usually to a value around .05 per cent of the fundamental power, by the use of a suitable KVA/KW ratio and harmonic suppressing networks.

4. In spite of all that has been written regarding proper neutralization procedure, many transmitters can be found that are improperly neutralized. The easiest check for improper neutralization is as follows: Disconnect the plate voltage on the stage to be checked and all succeeding r-f stages. Note the d-c meter in the grid circuit of the stage following the stage to be checked. If there is any grid current flowing, merely adjust the neutralizing control of the preceding stage until the current drops to zero.

5. Excessive hum on carrier may be reduced by increasing the amount of filter used in all the power rectifiers, and by using properly matched r-f and rectifier tubes.

HOW THE R-F AMPLIFIER WORKS

One of the greatest problems that one encounters in transmitter adjustment is the question of how much L and C to use in the tuned circuits, and the design and adjustment of a proper coupling system for the transmission line. Before entering a discussion of the amount of L and C to use, it is necessary to see how a r-f amplifier tube works.

In Fig. 1 is seen a plot of the various voltage and current relations during an electrical cycle. A sinewave voltage, e, from the oscillator or r-f amplifier stage, is impressed on the grid of the tube along with the d-c grid bias. This bias in the case of a class C amplifier is about two times cut-off bias. For a class B stage it is just cut-off. The a-c voltage on the plate, e_p, is superimposed upon the d-c plate voltage, E_p. This e_p is 180 deg. out of phase with the voltage e_g. Grid current i_g, is drawn when the grid voltage e_g is positive. The a-c plate current i_p starts to flow when the grid voltage e_g is positive and above the theoretical cut-off bias line. This a-c plate current is not sinusoidal, as is the grid voltage, but unsymmetrical due to being operated on or past the bend of the characteristic curve.

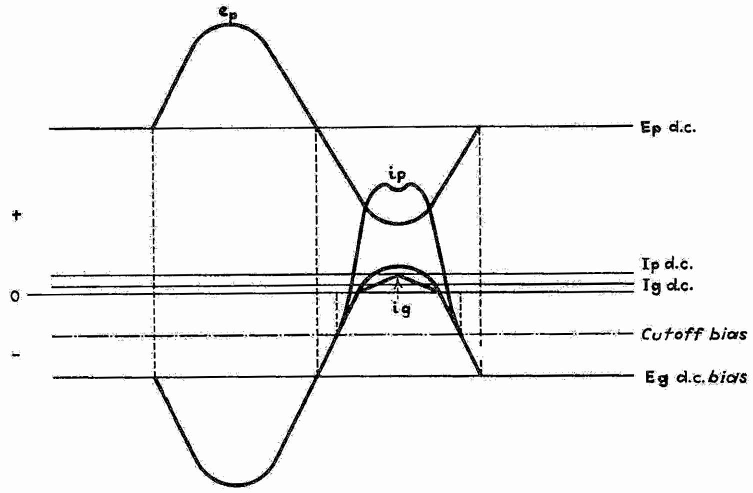


Fig. 1. Instantaneous relations between plate and grid voltages and currents in r-f amplifier

This simply means that the wave-form is replete with harmonics of the fundamental frequency. To reduce this harmonic content to a minimum, it is necessary to have a large circulating current present in the tank circuit to smooth out the wave form and transform it into something resembling a sinewave. This effect is termed the "flywheel effect." The greater the capacitance C in the tuned tank circuit, the greater the circulating current.

The fly-wheel effect operates as follows: When the a-c grid voltage e_g goes positive, the a-c plate current pulse i_p flows through the tank circuit, comprised of L and C in parallel. It produces an r-f voltage across it, charging the tank condenser C. At the moment e_g starts to go negative the condenser C discharges to L via the plate end of the tank circuit, and charges the other set of plates of C, which is the end connected to the plate supply. When the grid voltage e_g is negative no i_p flows, but the condenser C discharges in the opposite direction to which it did at first because the other set of plates of C has been charged. This completes the cycle of output r-f voltage, and explains why only one tube is necessary in a class B or C r-f amplifier for correct operation.

The ratio of KVA/KW, the ratio of volt-amperes in the tank-circuit to the d-c plate input power, $E_p \times I_p$, for maximum reduction of harmonics should be at least 12.6. It is customary to use a value between 15-25. In the preceding stages, those preceding the modulated amplifier, it is not necessary to use such a large ratio because the tank circuit is usually shielded. A value of 5 or thereabouts will do in these stages. The values of inductance and capacitance in the tuned plate circuit can be computed.

If the stage is single ended, not push-pull, the rms value of the peak plate voltage is:

$$e_p(r.m.s.) = \frac{2 \times E_p \times Eff}{3.14 \times .707} = 0.9\, E_p \times Eff$$

As the efficiency of a class C stage is usually taken as 72 per cent,

$$e_p\ (r.m.s.) = .65\ E_p.$$

If push-pull is used,

$$e_p\ (r.m.s.) = 1.3\ E_p.$$

For a class B single stage at 63 per cent efficiency.

$$e_p\ (r.m.s.) = \frac{2 \times E_p \times .63}{3.14 \times .707} = 0.567\, E_p$$

For a push-pull stage

$$e_p(r.m.s.) = 1.13\ E_p.$$

At resonance $X_C = X_L$ and therefore the amount of circulating current through these two branches is equal.

$$I_L = I_C = \frac{\frac{KVA}{KW} \times E_p \times I_p}{e_p\ (r.m.s.)}$$

Therefore:

$$X_c = \frac{e_p\ (r.m.s.)}{I_L}\ ; C = \frac{1}{6.28 \times F \times X_c}$$

$$\text{and } L = \frac{X_L}{6.28 \times F.}$$

By the use of Fig. 2, all computations necessary for the derivation of C and L, other than those on the chart, are eliminated. This chart is not absolutely accurate but close enough for all adjustment purposes.

In the modulated amplifier and the Class B stages, if any, it is customary to use a KVA/KW ratio of 15-25.

TRANSMISSION LINE TROUBLES

Much trouble is encountered in the design and correct adjustment of a transmission line coupling unit. Dietsch has shown the proper method of designing such coupling devices.[1] If we wish to terminate the transmission line into a tank circuit, it is necessary to provide such values of C and L so as to provide a suitable KVA/KW ratio and at the same time offer a unity power factor or resistance load to the transmission line so there will be no reflection losses. Before proceeding with the design of the coupling units it is necessary that the characteristic impedance of the transmission line and the antenna resistance be accurately known. Any suitable measuring device may be employed for these determinations.[2]

The value of tank capacity may be derived from the following formula:

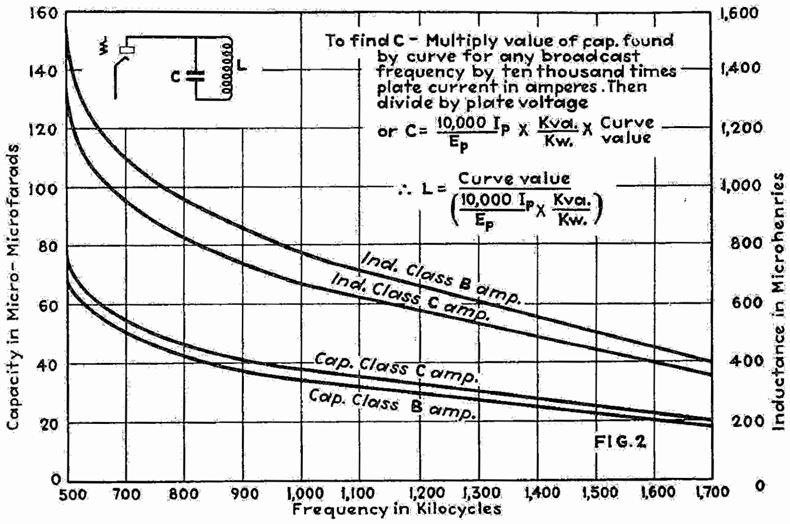
$$C_{\mu\mu f} = \frac{KVA}{KW} \cdot \frac{10^{12}}{6.28 \times F \times Z_o}$$

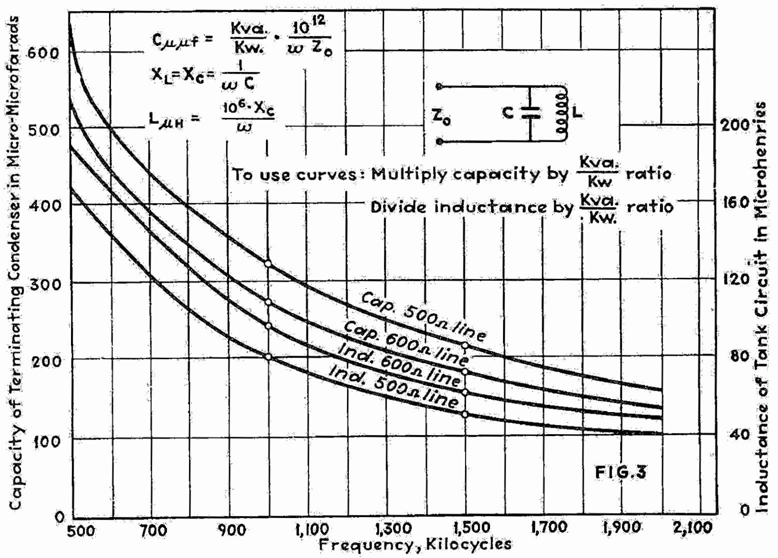
Where F = frequency in cycles and
Z_o = characteristic impedance of transmission line

Since $X_L = X_C$ at resonance, $L = X_c/6.28\ F$.

Calculations have been made for the 500 and 600 ohm. transmission lines only, because of their widespread use as compared to the low impedance concentric transmission line.

The values of C and L versus frequency for the 500 and 600 ohm. transmission lines are shown on Fig. 3.





Figs. 2 and 3. Charts for calculating capacity and inductance of radio-frequency amplifier tank circuits

The value of mutual inductance, M, between the tank coil and the antenna coupling coil can be derived from the following formula:

$$M = \frac{10^6}{6.28} \cdot \frac{X_C}{2\pi F}\sqrt{\frac{R_a}{Z_o}}$$

Where:

M = Mutual Inductance in μ

X_c = Reactance of tank capacity

R_a = Antenna resistance

Z_o = Transmission line impedance

F = Frequency in cycles

When it is merely desired to use a terminating capacity at the antenna end of the transmission line and not a tank circuit, the value of capacity necessary to provide a unity power factor load is equal to:

$$C\,(\mu_f) = \frac{10^6}{6.28 \times F \times Z_o}\sqrt{\frac{Z_o - R_a}{R_a}}$$

Where F = frequency in cycles

Z_o = transmission line impedance

R_a = antenna resistance

The antenna will be properly coupled to the transmission line, when using a tank circuit at the antenna end of the transmission line, when the antenna current, and the current in the capacitive and inductive branches of the tank circuit are the pre-calculated value.

The current in the antenna can be easily calculated.

$$P = I^2_{ant} \times R_{ant} \text{ or } I_a = \sqrt{P/R_a}$$

Where P = input to transmitter.

The current in the capacitive branch of the tank circuit may be determined as follows:

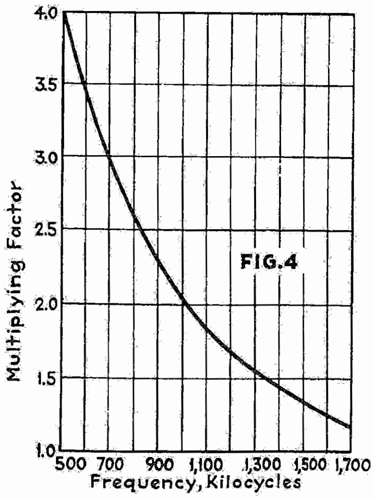


Fig. 4. Correction curve for applying Figure 5

$$E_{(rms)} = \sqrt{Z_0 \times P}$$

$$\text{and } I_{cap} = \frac{KVA/KW \times P}{E_{rms}}$$

It is usually customary to use a KVA/KW ratio slightly higher than that in the final amplifier tank circuit if the antenna impedance is relatively high. This impedance can be determined from any good r-f measuring set.[2] When the antenna impedance is low, a lower KVA/KW ratio may be used.

The value of current in the inductive branch of the terminating tank circuit may be derived as follows:

$$X_L = R \times \frac{Z_o}{X_c};\ Z_o = R + \left(\frac{Z_o}{X_c}\right)^2 R;$$

$$\therefore I_L = \frac{E_{rms}}{X_L}$$

Where X_c = reactance of tank cond.

X_L = reactance of tank ind.

R = tank inductance resistance plus reflected resistance of antenna coil.

The antenna should be resonated at the transmitter frequency by means of a driver oscillator coupled to the antenna coupling coil, with the tank circuit disconnected from the transmission line. Then the tank circuit should be connected across the transmission line and the antenna coil opened. At this point tune the tank circuit to resonance. The amount of inductance in the antenna may be calculated from any of the many formulas in Bulletin 74 of the Bureau of Standards. Now reconnect the antenna and ground or counterpoise to the antenna coupling coil and put the transmitter into operation. The current readings in the antenna, and the capacitive branches of the tank circuit should indicate the values already calculated. When the antenna current and the capacitive current are both low, there is insufficient excitation from the transmitter. Increase the r-f drive. If the capacitive current is high and the antenna current is low, there is insufficient inductance in the antenna coupling coil. With the above readings reversed, the opposite is true. Whenever the antenna coil is touched the antenna must be again tuned to resonance.

It will be noticed that the current in the inductive branch is not its pre-calculated value when the other two readings are correct. It is merely necessary to vary the inductance of the tank circuit until the inductive current is correct. At this point the tank circuit offers a unity power factor or resistance load to the transmission line.

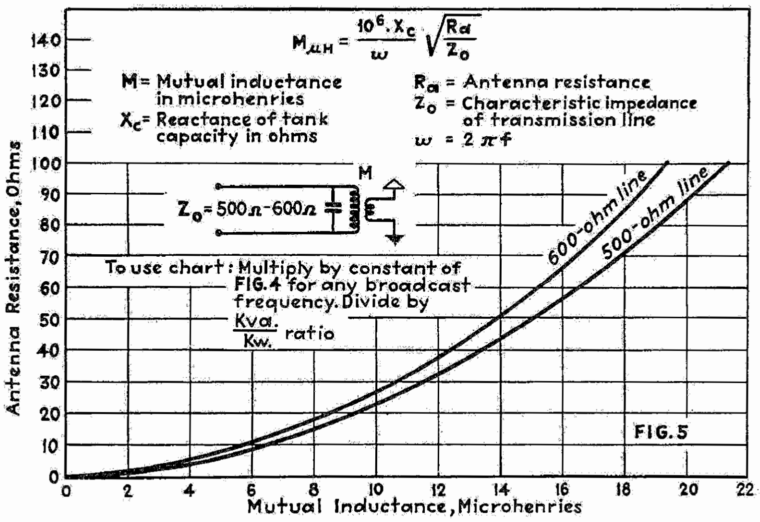
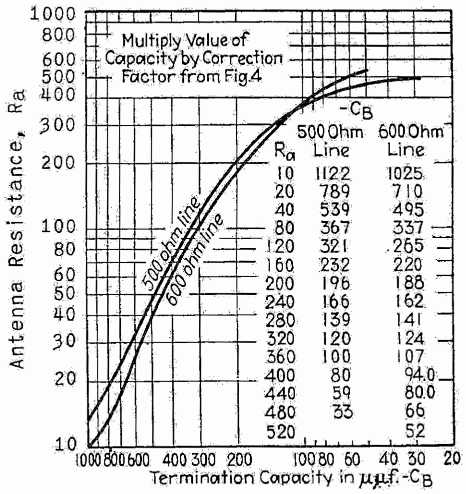


Fig. 5. Chart for computing proper mutual inductance between the tank and the antenna for two impedance values of transmission line



R_a	500 Ohm Line	600 Ohm Line
10	1122	1025
20	789	710
40	539	495
80	367	337
120	321	265
160	232	220
200	196	188
240	166	162
280	139	141
320	120	124
360	100	107
400	80	94.0
440	59	80.0
480	33	66
520		52

Fig. 6. Terminating capacity vs. antenna resistance for 500-600 ohm lines

It is recommended that the tank circuit be used to terminate the transmission line and not merely the capacitor as in Fig. 6, for increased harmonic reduction.

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1-Antenna Termination, Carl G. Dietsch. *Electronics*, September, 1935.

2-Antenna Measuring Set, W. B. Lodge, *Radio Engineering*, April, 1934.

ELIMINATING SPURIOUS RADIATIONS FROM BC TRANSMITTERS

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PRACTICAL METHODS FOR DETERMINING EXISTENCE AND EXTENT OF HARMONIC GENERATION AND OF DESIGNING AND APPLYING CORRECTIVE FILTERS

Even though commercial design of radio transmitters is such that most spurious signals are eliminated or greatly reduced, there are many cases where such radiation does not appear until the transmitter is in operation and must be corrected by the station engineer. Among the common cases of such are combination frequencies resulting from two different transmitters, and cases where a harmonic is particularly objectionable because it falls on the frequency of another service nearby.

The first requisite is an accurate diagnosis of their origin. Failure to find a successful cure has usually resulted from a hasty conclusion as to the nature of the trouble, and consequent improper corrective measures. One of the essential instruments needed is a simple resonant wavemeter used as a wave trap, with a wide, carefully calibrated frequency range. This instrument does not contain harmonics or respond to combination frequencies, as does a transmitter, receiver, heterodyne frequency meter, or any other instrument containing a vacuum tube. Any conceivably spurious response from this wavemeter is likely to be several octaves removed, and not in exact harmonic relationship with the primary response frequency.

A spurious response in a receiver from a spurious signal from the transmitter can be distinguished quickly by inserting the resonance wavemeter as a wave trap in the antenna lead of the receiver. The signal will drop when the wavemeter is tuned to the frequency at which the signal enters the receiver. If the fault lies in the receiver, this drop will be the correct frequency of the transmitter. If the fault lies in the transmitter, or at least somewhere external to the receiver, the wavemeter will indicate a spurious frequency at whatever point the receiver is tuned.

A direction finder is another desirable instrument for locating the source of an undesired signal. Cases frequently occur where a spurious frequency carries the program of one radio transmitter, but is actually generated in another transmitter, or in something completely foreign to any transmitter. When a direction finder is not available, useful results may be obtained with a portable radio receiver which has a signal strength meter.

Location Radiating Source

Such a direction finder or a signal strength indicator leads to the antenna radiating the signal. The signal itself is generated in some non-linear (or rectifying) device, and may be carried some distance by wires which serve as a transmission line, before it reaches the point of radiation. Before correction is made, it is ordinarily necessary to locate the non-linear device. The correction usually consists of adding or modifying networks adjacent to the non-linear device in such a manner that the spurious signal cannot reach the radiator, or else in such a manner that a contributing frequency necessary to produce the spurious radiation is prevented from reaching the non-linear device.

With the recent rapid increase in the number of broadcast stations, a new spurious frequency is sometimes formed by a combination of the frequencies of two (or more) stations. To produce the new frequency, energy from both of the base frequencies must flow through one non-linear device. Common devices for production of such interference are:

1—The final amplifier tube of one of the stations.
2—Any vacuum tube in an indicating instrument connected to the final amplifier, such as an inverse feedback rectifier, audio monitor rectifier, remote ammeter rectifier, or phase monitor.
3—The crystal oscillator tube or other low level rf or af tube in the transmitter or amplifier preceding the transmitter.
4—Rusty iron or corroded copper in the antenna system.
5—Man-made structures which are not part of a transmitter, possibly at some distance (a mile or more) from the transmitter. Poor electrical connections, old plumbing, and corroded metal roofs fall in this category.
6—The ionosphere (assuming that the debated "Luxemberg Effect" is accepted as a reality).
7—A vacuum tube in the radio receiver in which the interference is observed.

Combination frequencies usually are frequencies which are defined by $f = mf_a + nf_b$, where f is the spurious frequency, and f_a and f_b are the correct operating frequencies of two radio transmitters; m and n are any integers, positive, negative, or zero. The entire range of spurious frequencies expressed by all values of m and n are ordinarily produced when two frequencies are combined and passed through a non-linear device. However, the only one which ordinarily concerns us is the one which meets three conditions:

It is initially strong, the associated networks, transmission line, and radiator convert a substantial part of it into radiation, and finally it falls on a frequency which interferes with some other service.

Harmonic Suppression

Good engineering requires suppression of all spurious radiations. This is not always attained in the design, construction, and operation of transmitters. Strong radiations of the kinds most frequently occurring (such as harmonics) are eliminated fairly well in the manufacture of the transmitters. Other spurious radiation is usually eliminated only if it is observed either by a government monitoring station, or by someone who suffers interference from it. Even after an interfering radiation is observed, it is not always possible to identify the station responsible for it and to persuade the owners or engineers to eliminate it. A few specific cases might be cited.

The FCC ordered a broadcast station to eliminate third harmonic interference at a communications receiving station about two miles distant. Field intensity measurements indicated about 10 microvolts per meter at a mile, but there was such a large fluctuation in the field that no useful intensity measurements could be made. The loop antenna of the field intensity meter gave a bearing toward the broadcast station.

It was assumed that the harmonic (which is always generated in the final amplifier of the transmitter) was being radiated and a diagnosis was first made to determine whether the radiation was from (1) the antenna, (2) power lines, audio input, or other wires leaving the transmitter, (3) direct radiation from the final tank circuit (particularly likely because of poor shielding). When the transmitter was operated into a dummy antenna, the interference disappeared, establishing the fact that the only radiation was along the transmission line to the antenna. To our surprise the normal cure (a trap circuit resonant

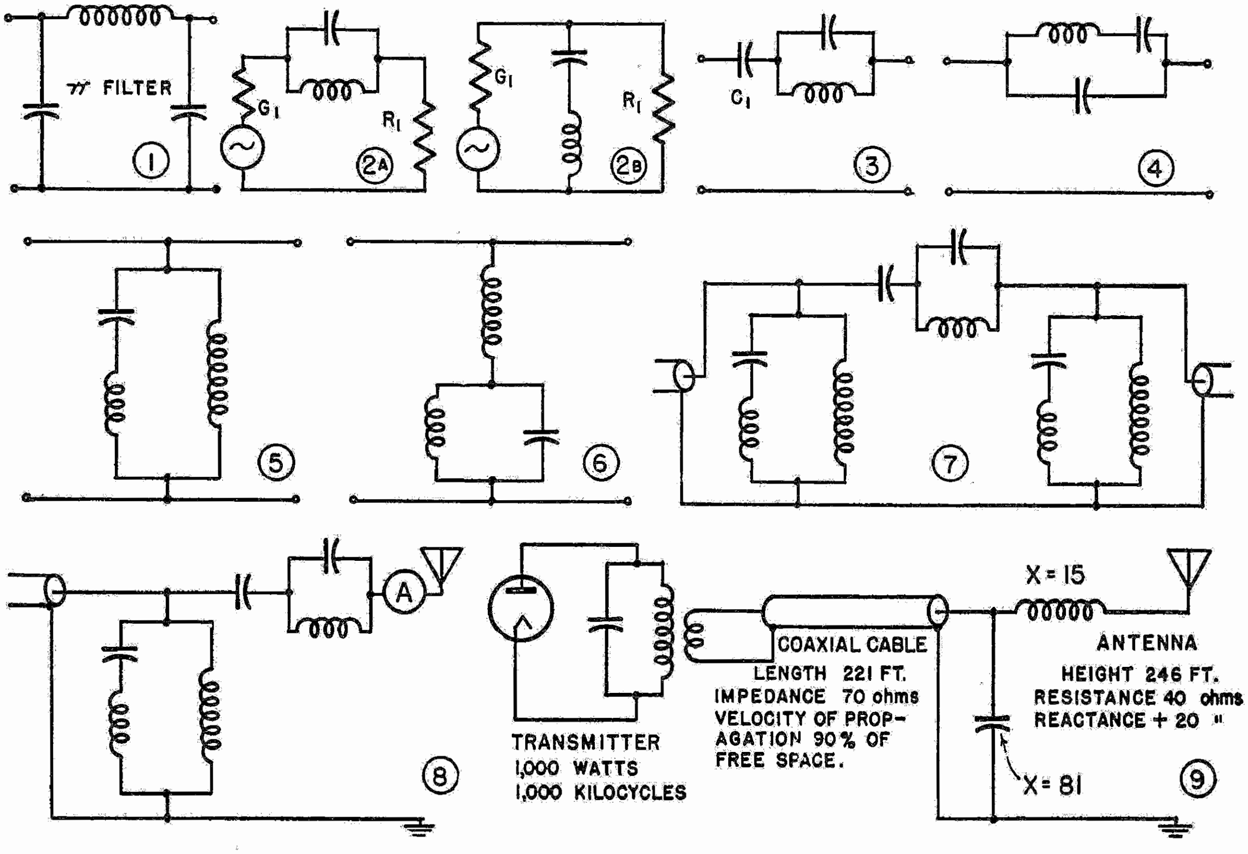

Fig. 1--Simple pi filter. 2A--Simple wave-trap filters. 3--Type A filter with reactance a pass frequency tuned out by C_1. 4--Filter with Fig. 3 characteristics. 5--Type B filter with reactance at pass frequency tuned out by an added inductance. 6--Filter with Fig. 5 characteristics. 7--A pi filter of resonant circuits. 8--L network prevents induced antenna current from nearby station from registering on antenna meter. 9--Typical filter for rejection of 1100 kc from transmission line input.

at the harmonic frequency, connected between the transmitter output and the transmission line input) was ineffective.

Trap Characteristics

The performance of the trap was proven to be good. Supplying a signal at the harmonic frequency from a signal generator established that a signal at harmonic frequency was adequately suppressed in the trap. But when the main transmitter was fed into the line, the harmonic was unaffected by tuning the trap near the harmonic frequency, and was reduced only when the trap was tuned to the fundamental frequency of the transmitter. This paradox seemed unbelievable for a time. The implication, however, was that the signal passed from the transmitter to the antenna at fundamental frequency, and was converted to third harmonic in the antenna.

The remote meter rectifier tube in the antenna turning unit was removed, but the harmonic remained. Nothing else in the antenna seemed to be capable of producing a harmonic. The field intensity meter was set up about two hundred feet from the antenna, with a long cord to the headphones, and various parts of the antenna system were disturbed while listening for any change in the harmonic.

Finally it was found that the harmonic fluttered violently when the guy wires of the antenna tower were jarred. The guys were rusty iron, apparently with poor contacts at splices or other attachment points that became generators of the harmonic. The obvious correction was replacement of the guy wires. Such problems from a little rust seem far-fetched. However, it is a matter of public record in this case that this harmonic interference on a communication frequency caused the death of an aviator.

Unusual Source

In another similar case a harmonic which did not respond to the usual wave traps was finally found to be generated in the vacuum tube rectifier at the antenna which was used for a remote antenna ammeter. This was a rectifer which produced several milliamperes dc.

There have been numerous cases recently of spurious radiations in the broadcast band at a frequency of $f = 2f_a - f_b$. For instance, where there are stations in the same city on 1000 kc and on 1100 kc, a signal is heard on 900 kc. It carries modulation of both transmitters, but much stronger modulation of the 1100 kc transmitter. On investigation it is found to be radiated from the 1000 kc station. It is produced in the final amplifier tube of the 1000 kc station, when the 1100 kc signal is received in this tube from the other station.

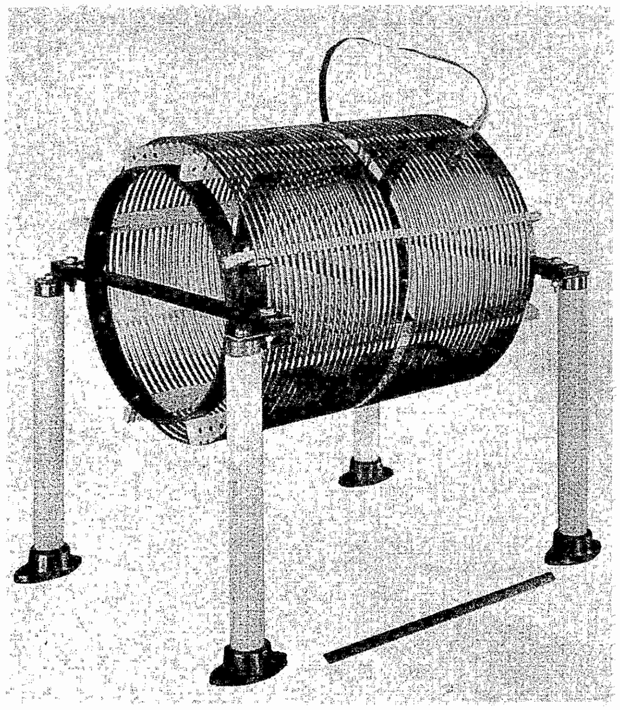

Fig. 10--Transmitting inductance well designed to give high Q factor.

Correction consists of inserting a filter somewhere between the final amplifier tube and the antenna of the 1000 kc station, and tuning the filter to reject 1100 kc. When properly tuned, the signal on 900 kc disappears. Final tuning is most easily accomplished with both transmitters in operation, and a receiver tuned to 900 kc to indicate filter adjustment. In a similar manner, the other station produces a spurious radiation on 1200 kc, and must have comparable corrective measures.

Particularly troublesome cases occur where the frequency separation is small and the distance between antennas is small. In numerous cities both 1450 and 1490 kc are assigned. The frequency separation of 2.7% is very difficult to separate with resonant circuits. In one city, stations on these frequencies have antennas only 500 ft. apart. Here an added difficulty appears: each antenna picks up so much current from the other station that the antenna current indication is inaccurate by as much as 20% when the other transmitter is operating. This trouble also has been corrected by proper filter design, but it is a severe complication to the design.

Spurious Modulation

It was found that a 250-watt broadcast station (tested without modulation) actually had a substantial modulation from the program of a higher power station several miles away, and widely different in frequency. Preliminary check showed that this signal was not coming in on the audio lines to the new station. If this was a case of picking up the spurious signal in the antenna and rectifying it in the final amplifier, a trap in the antenna lead to reject the carrier frequency of the other station would be needed. However before building the trap, the inverse feedback rectifier tube was pulled out of its socket. The spurious modulation disappeared completely! This rectifier takes its input from a small sampling line loosely coupled to the final tank circuit of the transmitter. All that was necessary was a small resonant trap in this input circuit to the rectifier.

The trap was built out of miniature receiver components, and worked perfectly! A tank built in the main transmission line from the transmitter to the antenna would have done the job too, but it would have required large and expensive transmitter components, its adjustment would have changed the loading on the transmitter, and it would have caused some loss of transmitter power.

In a similar case spurious radiation appeared as a hash 30 kc on each side of the carrier frequency. When the transmitter was not modulated, no spurious radiation was found.

Other Interference

Here again it was found that by removing the inverse feedback rectifier tube, the spurious signal disappeared. A squeal at 30,000 kc was apparently set up in the inverse rectifier in some manner, probably by feedback completely around the circuit of this rectifier, and the transmitter audio and radio circuits. A slight reduction in the amount of feedback used was all that was necessary to eliminate the trouble.

In a communications station, where the same frequency was used for transmission and reception, when receiving, the plate voltage was turned off the transmitter, but the filaments were left on. Interference was found on the operating frequency, heavily modulated at 60 cycles. The tubes were obviously generating oscillation due to plate and grid being grounded to the center of the filament. The cure: leaving the grid bias on the final amplifier when the plate was turned off.

Police were using a frequency on the third harmonic of a broadcast station. They complained of hearing the broadcast station in their cars when near the transmitter. While examining the harmonic, we used a field intensity meter about a thousand feet from the transmitter.

It was found that the field intensity meter did not have sufficient selectivity in the input circuits, and therefore the fundamental frequency reached a tube in the receiver with sufficient amplitude to produce the harmonic in the receiver. A communications receiver with an additional wave trap at the fundamental frequency in its antenna lead disclosed the true third harmonic of the transmitter. This permitted correct tuning of the third harmonic trap in the transmission line input to reduce the interference.

The police car receivers apparently were subject to the same defect, insufficient selectivity in the input circuits, which lead to actual production of the harmonic in receiver. Correction of such difficulty cannot be accomplished at the transmitter, except by such drastic change as moving to a distant location or changing operating frequency.

We have mentioned earlier the necessity for filters where two stations have antennas near each other. In reality, where antennas must be near, the engineering problems are often simpler if only one antenna is used. Both transmitters are connected to the same antenna, with filters at the points of connection which pass the proper frequency and reject the other. There have been several installations of this type in the past, even including one where both stations used directional antennas. One tower was common to both directional systems. Other towers were not.

Monkey Chatter

The modulation sidebands of a transmitter often cause interference with reception on adjacent channels. There are two common causes: the audio frequency input to the transmitter may contain sufficient power at frequencies above 5000 cycles to cause noticeable interference on a station on the next channel. This can easily be eliminated by insertion of a 7500 cycle cutoff filter in the audio input to the transmitter.

The other cause of monkey chatter (overmodulation of the transmitter) is cured by keeping modulation down. Limiter amplifiers now in widespread use in broadcast stations have contributed greatly to correcting this difficulty.

The first filter in a transmitter output is the tank circuit. It is tuned to pass one frequency with small loss, and to offer high attenuation to other frequencies. By decreasing the ratio of inductance to capacity in the tank circuit, the attenuation of other frequencies can be improved, but at the same time the loss of power in the fundamental frequency increases.

Pi Net Suppressors

The addition of a pi or tee network in the output of a transmitter is often used for harmonic suppression. Such a network substantially attenuates all frequencies from the second harmonic upward, but is of little service where the separation between the pass frequency and the rejection frequency is small (under 30%).

Fig. 1 and Table I show design data for a pi filter suitable for harmonic suppression, for specific frequencies and load impedances. Satisfactory design for other frequencies and load impedances may be obtained by interpolation. The number of turns of inductance applies only to an inductance 6 in. in diameter, with 4 turns per inch.

Capacitances may be varied as much as 25% in order to use commercially available fixed capacitors. The currents shown for the capacitor are actual unmodulated amperes for a 250 watt transmitter. For other powers, the current varies as the square root of the transmitter power. For proper factor of safety, the capacitors should have commercial rating of at least twice the indicated values.

Filters of high selectivity such as are often required in elimination of combination frequency radiation are best constructed with one or more resonant circuits at the rejection frequency. This type of filter finds its best applications where the frequency difference between the pass and the rejection frequency is in the range from 2% to 50%.

When the frequency separation is greater than 50%, a pi filter is adequate, unless an unusually high degree of suppression is required. Where the frequency separation is less than 2%, separation usually is beyond the ability of resonant circuits. Where the frequency separation is less than 5%, it is ordinarily necessary to use a filter containing more than one resonant section.

Resonant Filters

There are two ways of connecting the resonant circuits in this kind of filter. In Type A (Fig. 2) the resonant circuit is parallel resonant, and is connected in series with the main circuit. This wave trap approximates an open circuit for the rejection frequency. The other form (Fig. 2B), a series resonant circuit connected across the main circuit, approximates a short circuit across the line at the rejection frequency.

The following formulas are useful: (1) to determine whether to use Type A or Type B filter, to evaluate its components and to compute the results which can be obtained. They contain numerous assumptions and approximations, but have sufficient accuracy for practical use in filter design.

Here the term rejection ratio refers to the ratio of the voltage passed at the rejection frequency (f_2) after insertion of the filter to voltage passed without the filter. The loss ratio is the power dissipated in the filter at the pass frequency (f_1) compared to the power delivered from the filter to the load.

D = $(f_1-f_2)/f_1$ the frequency separation.

X_2 = the reactance of the inductance (or capacitor) in Type A filter at f_2.

Y_2 = the reactance of the inductance (or capacitor) in Type B filter at f_2.

Q = the figure of merit of the inductance used in the filter. For typical transmitting inductances, Q may be assumed to be 300, if the Q of the capacitor is essentially high. In the diagrams the generator is assumed at the left, and the load at the right.

TABLE I

Operating kc	Line or antenna impedance, ohms	L=µh	Inductance turns (6" dia.)	Capacitance, µf	Capacitor Current, Amps.
600	20	5.3	4	.013	3.5
600	70	19	10	.0038	1.9
600	300	80	31	.00088	0.91
1000	20	3.2	3	.0080	3.5
1000	70	11	7	.0023	1.9
1000	300	48	20	.00053	0.91
1500	20	2.1	2	.0053	3.5
1500	70	7.4	5	.0015	1.9
1500	300	32	15	.00035	0.91

Values of components required for a pi filter for harmonic suppression for a 250 watt broadcast station.

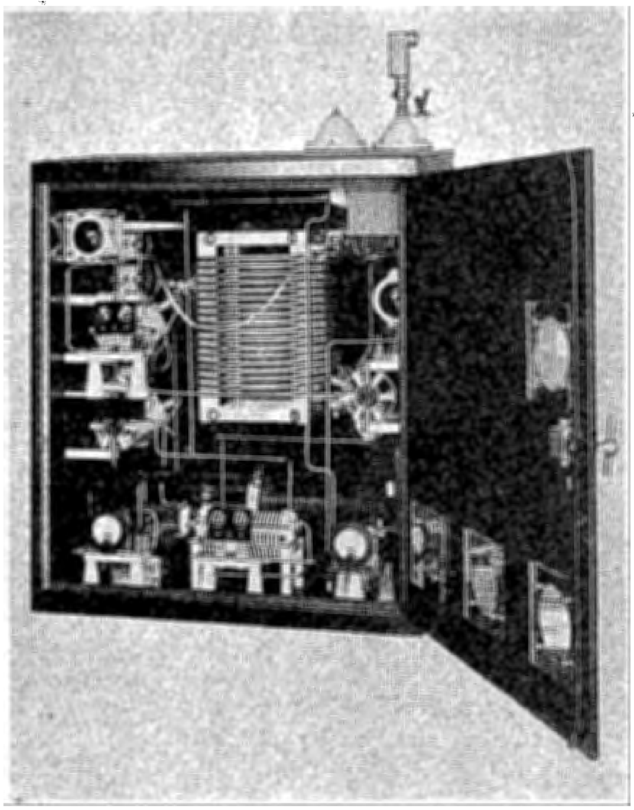

Fig. 11--Typical network incorporated in an antenna tuning unit

However, f_2 may be impressed on either left or right side of the filter, in various applications, and the filter is intended to prevent it from reaching the other side. The designation of one side as "generator" and the other side as "load" is therefore not always clear.

R_1 = the load impedance at f_1.
R_2 = the load impedance at f_2.
G_1 = the generator impedance at f_1.
(G_1 is assumed to be small compared to R_1).
G_2 = the generator impedance at f_2.

In choosing between Types A and B, the former is more effective when the impedance of the circuit is lower at f_2 than at f_1, and vice-versa. Definite formulas for selection of types are:

When R_2G_2 is less than $R_1{}^2$, use type A.
When R_2G_2 is greater than $R_1{}^2$, use type B.

Any of the following methods may be used to determine which case prevails.

(1) If all components in the circuit are known, it is possible to make calculations of the impedance at the filter point, for both f_1 and f_2. Usually the impedance at f_2 is found to be so greatly different from that at f_1, the choice is obvious.

(2) The impedances may be measured with an rf bridge.

(3) Types A and B may each be tried out, and the preferable one determined from the comparative performance of the two. If this method is followed, it is best to design the two for equal loss at f_1 (according to the formulas given hereafter), and then observe the difference in performance at f_2.

Methods (1) and (2) are applicable to determination of load impedance, and from this impedance alone, the advantage of one type of filter may be so great, or the needed performance so easily obtained, that there is no need for further analysis. These methods, however, do not readily give the generator impedance. Method (3) for this reason is sometimes the only satisfactory method.

To design a type A filter, a value of X_2 must be selected which will give satisfactory performance. Then Rejection Ratio = $(R_2+G_2)/QX_2$ = R_2/QX_2 (when $G_2 \ll R_2$).

Loss Ratio = $X_2/4D^2QR_1$.

Losses Must Be Lowered

It is seen that a larger X_2 makes the rejection ratio lower (an improvement), but also makes the loss ratio higher (a disadvantage). Since performance in both respects is improved by increasing Q, use the lowest loss inductance available, a good capacitor, make good connections between them, and see that the housing and other objects around them do not greatly reduce the Q. Also since the loss varies as D^2, performance deteriorates rapidly as the frequency separation is decreased.

In practice, it is rarely necessary to obtain either of these performance factors lower than 0.001. Where frequency separation is small, it is sometimes necessary to allow each factor to be as high as 0.1. If either factor is higher than 0.1, the filter usually is considered useless, and the project abandoned, or a more complex filter used. If we set 0.1 as the maximum limit of each factor, and assume that Q = 300 and $R_1 = R_2$, we find that this type of filter is useful only for frequency separations of 1.7% or greater.

In resonant filters, high values of current and voltage appear in both the inductance and the capacitor. Components of suitable size must be selected to prevent failure from overheating or voltage breakdown.

In Type A filters the current is ½D times the current in the load. For example, if the frequency separation is 3% the load current is 5 amperes, the current circulating in the filter inductance and capacitor is 83 amperes. This calls for rugged components and makes it clear why severe losses may occur. The voltage across these components is the current times the reactance, and is likely also to be so high that precautions must be taken to prevent corona.

Component Selection

In selecting components, the current rating of fixed capacitors should be 1.22 times that shown to allow for rise in RMS current with 100% modulation. The voltage rating of inductances and air condensers should be 2.30 times that shown in order to allow for maximum RMS voltage with 130% positive peaks of modulation. To find peak voltage, this RMS value must be multiplied again by 1.41. Furthermore, for adequate factor of safety in commercial equipment, it is well to have actual current and voltage ratings about double again the maximum values just computed.

These current and voltage calculations are for the pass frequency (f_1). In most applications, the current and voltage at the rejection frequency (f_2) is so much smaller that it may be neglected. This is so however in some cases, such as operation of two transmitters into one antenna.

For the design of Type B resonant filters the corresponding formulas are:

Rejection Ratio = $(R_2+G_2)Y_2/R_2G_2Q$
Loss Ratio = $R_1/4D^2QY_2$

Here the voltage across each component in the filter is ½D times the voltage across the load and the current is found by dividing the voltage by the reactance.

The insertion of a Type A network in series with a circuit introduces a new reactance in the circuit at f_1. This may be tuned out by an equal and opposite reactance, as shown in Fig. 3. If $f_2 > f_1$ the added reactance must be capacitive. If $f_1 > f_2$ it must be inductive.

In a different arrangement of the three components accomplishing the same end (Fig. 4), the two upper components are series resonant at the pass frequency, and therefore form a zero impedance. They have an inductive reactance at the higher rejection frequency, and a parallel capacitance of the same value of reactance forms a parallel resonant circuit, which approximates an infinite impedance at the rejection frequency.

The small difference in performance of these two circuits permits choice from ease of adjustment, and commercially available components. In the form shown in Fig. 3, the parallel resonant circuit must be tuned to (f_2) first. Then the series capacity C_1 may be tuned for (f_1) without disturbing the (f_2) adjustment. In Fig. 4, the circuit must be tuned first at (f_1), and then the (f_2) tuning does not affect the (f_1) adjustments.

Since tuning adjustment must be made to closer than 1%, construction must be used which will be stable against mechanical shocks and temperature changes. In each three-element group, at least two elements (including the third element) must be continuously variable. It is ordinarily necessary to have these variables so constructed that they can be adjusted with full power on the transmitter, and with all doors closed and other electrostatic shields in place. In any ordinary cabinet construction, opening a door will detune the inductances several percent.

The Type B filter has similar possibilities for adding a third component to neutralize the effect of the filter at (f_1). Figs. 5 and 6 correspond in principle to Figs. 3 and 4.

Where a simple resonant filter does not offer good enough performance, resonant groups may be used to replace single elements in a pi or tee network, as shown in Fig. 7. This network permits complete tuning at (f_2) first, and then tuning at (f_1) without disturbing the (f_2) adjustments. To attain this independence of adjustment, one of the things necessary is to see that stray inductive and capacitive coupling between components is kept small.

This nine-element filter is in use where stations on 1450 and on 1490 kilocycles are located within a quarter mile of each other.

In designing these filters consisting of two or more Type A and Type B groups connected in L, pi, or tee arrangement, the loss in each group must be kept low enough so the sum of these losses will not be objectionable. The rejection ratio in each group may be kept quite low, but the total rejection increases rapidly as the number of groups is increased, and so is as high as desired.

Where two antennas are so close together that the antenna ammeter of one shows an objectionable amount of current picked up from the other station, an L or tee arrangement of these groups must be used so that a Type A group is nearest the antenna and is designed to have a high enough impedance at (f_2) to reduce the antenna current to a harmless value.

For a numerical example of network design assume the previous example of a station operating on 1000 kc and another nearby on 1100 kc. A spurious signal is heard on 900 kc with the program of the 1100 kc station dominant. However, a direction finder, field strength indicator, or other means indicates that the spurious signal is being radiated from the 1000 kc station. The inverse feedback, audio monitor, remote ammeter, and any other similar rectifiers are disconnected without improving the situation. The presumption therefore is that the final amplifier of the 1000 kc station is the tube producing the spurious frequency. The correction consists of inserting a filter between this tube and the antenna which will reject the 1100 kc signal.

The constants of the station at 1000 kc are all known (Fig. 9). The antenna impedance is known to be 40+j20 at 1000 kc, and 50+j50 at 1100 kc. The L network is correct for matching the antenna impedance to the 70-ohm line at 1000 kc. By recomputing the performance of the antenna tuning network at 1100 kc, it is found that the load at the end of the transmission line is 107-j58 at 1100 kc. Then by computing the transformer effect through the transmission line, it is found that the load at the input to the transmission line (where the filter is to be inserted) is 35+j19, or an impedance of 40 ohms.

On the generator side we need not make an exact calculation, but we know that the small pickup coil offers low impedance, and is little affected by the tank circuit at this frequency which is not the resonant frequency of the tank circuit. Therefore both R_2 and G_2 are smaller than R_1, so we may be sure that we have the condition R_2G_2 less than R_1^2, and the Type A filter will be most effective.

The formulas for performance, after substituting $D = 0.1$, $Q = 300$, $R_1 = 70$, and $R_2 = 40$ are: Rejection ratio $= 0.13/X_2$; Loss Ratio $= 0.0012X_2$.

We select a value of $X_2 = 10$. Then the rejection ratio is 0.013, or 1.3% of the objectionable frequency is passed by the filter. The loss factor is 0.012, or a power loss of 12 watts in this 1000 watt station.

Calculating shows the current in the inductance and capacitor is 5 times the current in the load, or 25 amperes. The voltage across the inductance is 250 volts. Any transmitting components will handle this voltage, plus modulation peaks.

FACSIMILE

A Brief Theory of Present Day Facsimile Prepared by the NAB Department of Engineering with the Assistance of Mr. John V. L. Hogan, President of Faximile, Inc.

Electrical transmission of pictures was conceived more than a century ago by Alexander Bain, an English Physicist, who proposed a system so basically correct that the fundamentals of the present day system of Facsimile transmission show a remarkable likeness.

The work of many brilliant men has gone into the development of Facsimile as we know it today. It was first used only on wire line circuits and many methods of transmission were tried. Finally about the time radio broadcasting was born, pictures were transmitted by radio. In fact on July 6, 1924 the first picture, termed a photo-radiogram, was successfully relayed from New York to London and back to New York. Today the transmission of pictures is an accepted service, and one needs only to refer to his daily newspaper for this evidence.

The broadcaster's use of Facsimile was first authorized by the FCC in the late '20's utilizing frequencies in the range 1500 Kc to 2950 Kc. As a matter of interest two bands of frequencies were allocated to visual broadcasting which included still pictures and moving pictures (Television). In 1937 experimental use of facsimile was permitted on frequencies in the standard broadcast band from midnight to 6:00 a.m. Systems with varying standards were used for this service. In 1939 frequencies above 25 megacycles were authorized for experimental facsimile, and in 1940 the first facsimile rules were adopted by the Commission which provided for operation with FM broadcast stations.

On July 9, 1948 the FCC adopted its present Rules and Standards with respect to facsimile, permitting simplex or multiplex transmission on FM frequencies. (See Appendix I). A general and brief theory of the operation of one system operating in accordance with these rules is here outlined.

Facsimile Transmitting Equipment

The facsimile terminal equipment, for installation at the studios of an FM broadcasting station (or elsewhere, -- as for example in the editorial rooms of a newspaper) should comprise two complete transmitting scanners and their associated accessories including monitor recorders, scanner amplifiers and limiter line amplifiers. Two scanners are needed for the same reason that two turntables are required in a transcription studio. A single scanner necessarily causes some waste space (as well as delay) between successive pages, while the scanner is being reloaded and adjusted. The alternate use of two scanners permits a continuous, unbroken flow of facsimile copy.

A lamp, optical system, phototube and amplifier tube are mounted in the movable scanner head or carriage. The lamp illuminates a spot on the copy. The phototube receives reflected light from the illuminated spot through an optical system. The phototube current or output is controlled by the optical density of the desired portion of the original copy. For each revolution of the copy drum, the carriage advances along the drum, so that the copy is scanned in a spiral. The scanner produces electrical signals representative of the optical density of the copy scanned and feeds these signals to the scanner amplifier.

The scanner amplifier supplies power to the scanner, and uses the signals from the scanner to amplitude modulate a subcarrier which is developed in the scanner amplifier. The modulated subcarrier is amplified and fed into the limiter line amplifier which has two inputs. Signals applied to one of these inputs are amplified linearly. Signals applied to the other input are fed through an adjustable preset limiting stage whose output is either zero (or no signal), or equivalent to a full marking density. This limiting channel is useful for transmitting poorly prepared copy. The limiter line amplifier has a common output which operates with balanced circuits. This output of the limiter line amplifier may be fed directly into an FM transmitter for simplex facsimile broadcasting, or into the facsimile input channel of a transmitting multiplexer unit for multiplex operation.

Single scanner operation requires one of each of the following units: Scanner, Scanner Amplifier and Modulator, Limiter Line Amplifier, and a Pulse and Page Separation Signal Generator. (See Figure 1-A)

Dual scanner operation requires two each of the Scanner, Scanner Amplifier, and Limiter Line Amplifier units; one Pulse and Page Separation Signal Generator; plus control and output switching equipment. (See Figure 1-B)

The output of the facsimile input equipment may be fed directly into an FM transmitter, or relayed from the scanner location.

Synchronizing of broadcast facsimile transmitting and receiving equipment is accomplished by using identical or interconnected power sources. In areas where power line synchronization is not possible, automatic synchronizing amplifiers may be provided for receiving sets.

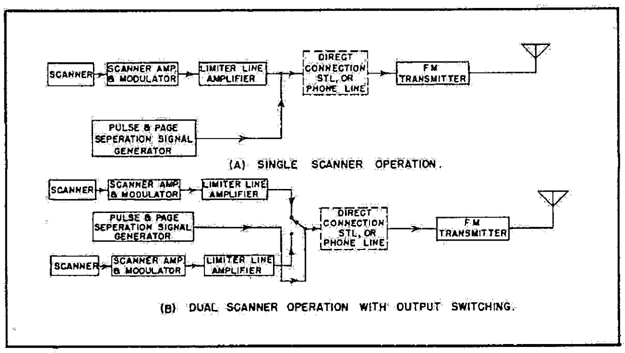

Figure 1--Input equipment used in Simplex Facsimile Broadcast Operation

The service area of an FM transmitter is affected in no way by the change from sound modulation to facsimile modulation. A good FM receiving antenna should be considered standard practice in most receiver installations.

A simplex facsimile broadcast operation can be advanced to a Multiplex FM sound and facsimile operation by feeding the two programs into a Multiplexer unit located at the FM transmitter. The mixed sound and facsimile output is then used to modulate the FM transmitter. (See Figure 2)

When planning a Multiplex aural and facsimile operation it should be noted that two program transmission lines are usually required to feed the two input channels of the Multiplexer unit. This may dictate the use of dual STL systems, dual telephone lines, or any acceptable combination. It is not necessary for the sound and facsimile programs to originate at the same place.

Various combinations of units may be used to form a mobile or portable facsimile input equipment. Basically, a Scanner, Scanner Amplifier, and Limiter Line Amplifier, are needed together with the Pulse and Page Separation Signal Generator.

Power source for portable or mobile operation must be synchronous or interconnected with that at the location of the broadcast station's fixed facsimile input equipment. It is suggested that, whenever possible, operations combining fixed, portable, and mobile facsimile input equipments be consistent in using the same phase of the normal three phase 115 volt 60 cycle AC power.

Facsimile Receiving Equipment

The signals needed to operate the facsimile recording equipment are taken directly from the output of the discriminator of an FM receiver. If the receiver is limiting excellent copy can be expected. (See Fig. 3 for simplex reception; Fig. 4 for multiplex.)

The discriminator output signal is applied to a recorder amplifier which amplifies the signal, rectifies it, and supplies current corresponding to the signal to mark the moist electrolytic recording paper in the recorder.

The recorder amplifier supplies all power to the facsimile recorder, including power to the synchronous recorder motor. A relay in the recorder amplifier interrupts the motor power when necessary to phase or "frame" the recorder.

The recorder has a humidor compartment which stores the moist electrolytic recording paper. Feed rollers advance the paper between the marking electrodes which consist of a printer blade and helix drum. After the paper is marked it passes over a heater which dries the paper and completes the marking process.

Marking the recording paper is nominally a constant current, low voltage process. In broadcast recording equipment 250 milliamperes of direct current are needed for full density marking. The voltage in recording may reach a maximum of about 50 volts depending on the amount of moisture in the paper. As the recording current passes through the paper from printer blade to helix, iron from the blade is deposited in the paper in proportion to the amount of current. The marking process gradually consumes the printer blade which is replaced along with the roll of recording paper.

Rotation speeds of recorder helix drum and scanner copy drum are 360 rpm. Paper feed in the recorder corresponds to scanner carriage advance rate of 3.43 lineal inches per minute. The recorder delivers approximately 28 square inches of copy in a minute. A fifteen minute broadcast program interval will result in four pages of recorded copy, 8.2 by 11.5 inches, in size, together with the required page separation signals.

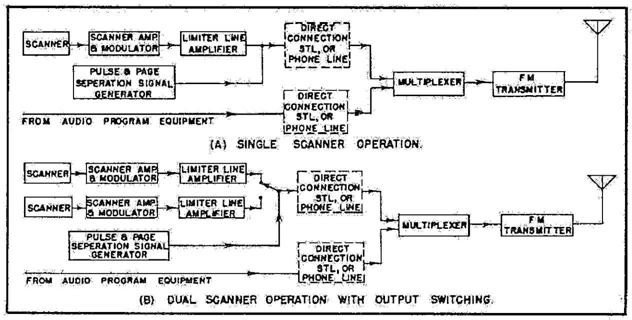

Figure 2--Input equipment used in Multiplex Facsimile Broadcast Operation

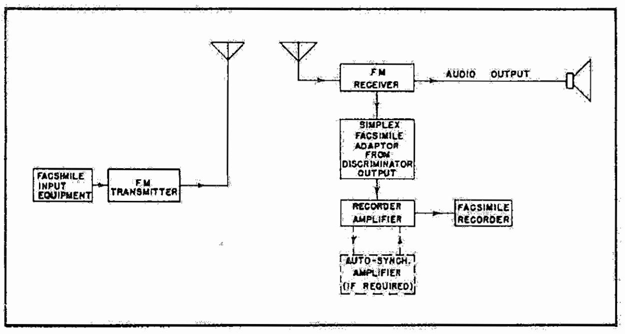

Figure 3--Equipment used in Simplex FM-Facsimile Broadcast Operation

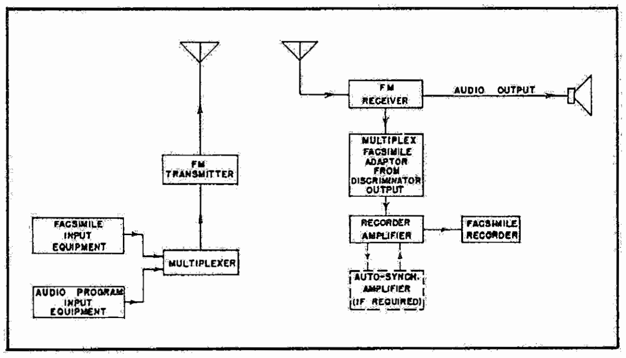

Figure 4--Equipment used in Multiplex FM-Facsimile Broadcast Operation

Discussion of FCC Standards

Standard (1) calls for straight line scanning from left to right and progressing from the top to the bottom of the page. This was selected, of course, so that the received copy can be comfortably read as it is being recorded.

Standard (2) states that the Index of Co-operation shall be 984. This is the number that defines the amount of intelligence that can be carried by a single facsimile page. It takes account of the lines per inch and the length of the scanning line. As usually applied, an Index of 984 results in a scanner whose effective line length, across the page, is 8.2 inches, whose line use ratio is 7/8 and whose definition is 105 lines per inch. The copy may be reproduced on recorders having the same dimensions or the recorders may deliver reduced or enlarged copy. The aspect ratio and the amount of intelligence will be the same in all cases, so long as the Index number remains approximately 984.

Standard (3) states the number of scanning lines per minute as 360. This standard is essential for synchronization, and determines the speed of transmission of copy having any particular definition or number of lines per inch. With the standard Index of 984 applied to an 8.2 inch scanning line, the standard drum speed of 360 r.p.m. results in a copy speed of 3-3/7 lineal inches, or 28.1 square inches, per minute. This produces four pages, each about a foot long, in each fifteen-minute period of broadcasting.

The remaining standards from (4) to (10) are principally important from the point of view of providing inherent interchangeability or compatibility between all transmitters and all receivers. They define the line use ratio as 7/8, describe the phasing and synchronization pulse which is to be transmitted during the 1/8 of the scanning cycle not used for copy, and prescribe a page separation interval, the use of sub-carrier amplitude modulation (and the laws which it is to follow) and finally specify sub-carrier noise level as at least 30 d. b. below maximum picture modulation. Noise levels may be higher in facsimile broadcasting than in sound broadcasting without causing comparable interference effects; with a 30 d. b. ratio, or even less, no signs of noise can be detected on facsimile copy.

APPENDIX I

FCC Rules and Regulations

S 3.266 Facsimile Broadcasting and Multiplex Transmission

(a) FM broadcast stations may transmit simplex facsimile in accordance with transmission standards set forth in the Standards of Good Engineering Practice Concerning FM Broadcast Stations during periods not devoted to FM aural broadcasting. However, such transmissions may not exceed one hour during the period between 7 A. M. and midnight (no limit for the hours between midnight and 7 A. M.) and may not be counted toward the minimum operation required by Section 3.261.

(b) FM broadcast stations may, upon securing authorization from the Commission, transmit multiplex facsimile and aural broadcast programs for a maximum of three hours between the hours of 7 A. M. and midnight (no limit for the hours between midnight and 7 A. M.) in accordance with transmission standards set forth in the Standards of Good Engineering Practice Concerning FM Broadcast Stations provided that the transmission of facsimile does not impair the quality of the aural program below 10,000 cycles per second, and that a filter or other additional equipment is not required for receivers not equipped to receive facsimile.

FCC Standards

Sections 1 and 8 of the Standards of Good Engineering Practice Concerning FM Broadcast Stations are amended by adding the following:

1 - DEFINITIONS

O. Index of Cooperation. The index of cooperation as applied to facsimile broadcasting is the product of the number of lines per inch, the available line length in inches, and the reciprocal of the line-use ratio. (e.g., 105 x 8.2 x 8/7 = 984)

P. Line-Use Ratio. The term "line-use ratio" as applied to facsimile broadcasting is the ratio of the available line to the total length of scanning line.

Q. Available Line. The term "available line" means the portion of the total length of scanning line that can be used specifically for picture signals.

R. Rectilinear Scanning. The term "rectilinear scanning" means the process of scanning an area in a predetermined sequence of narrow straight parallel strips.

S. Optical Density. The term "optical density" means the logarithm (to the base 10) of the ratio of incident to transmitted or reflected light.

8 - TRANSMITTERS AND ASSOCIATED EQUIPMENT

H. Facsimile-Engineering Standards.

The following standards apply to facsimile broadcasting under Section 3.266 of the Rules and Regulations.

1. Rectilinear scanning shall be employed, with scanning spot progressing from left to right and scanned lines progressing from top to bottom of subject copy.

2. The standard index of cooperation shall be 984.

3. The number of scanning lines per minute shall be 360.

4. The line-use ratio shall be 7/8, or 315° of the full scanning cycle.

5. The 1/8 cycle or 45° not included in the available scanning line shall be divided into 3 equal parts, the first 15° being used for transmission at approximately white level, the second 15° for transmission at approximately black level, and the third 15° for transmission at approximately white level.

6. An interval of not more than 12 seconds shall be available between two pages of subject copy, for the transmission of a page-separation signal and/or other services.

7. Amplitude modulation of subcarrier shall be used.

8. Subcarrier modulation shall normally vary approximately linearly with the optical density of the subject copy.

9. Negative modulation shall be used, i.e., maximum subcarrier amplitude and maximum radio frequency swing on black.

10. Subcarrier noise level shall be maintained at least 30 d.b. below maximum (black) picture modulation level, at the radio transmitter input.

THE ECONOMICAL OPERATION AND MAINTENANCE OF POWER TUBES*
NAB

OPERATION

Longer tube life means lower tube costs. Find a way to increase the useful life of tubes in your equipment and the unit cost per hour of operation drops accordingly.

Vacuum tube replacements constitute one of the three major cost items of operating a broadcast transmitting station. Because of this, and because of the importance of maintaining the best possible continuity of service, proper maintenance and operation of all tubes and in particular of the water-cooled and air-cooled power tubes used in the transmitter contribute to the efficiency of station operation. This is a subject deserving close attention on the part of the station engineer.

Many factors enter into the life of tubes. Of these the following are the most apparent:

1. Filament Voltage
2. Plate Voltage
3. Operating Temperature
4. Amount and nature of residual gas in tube
5. Number of times current is turned on and off
6. Fatigue of metal parts

Bright Tungsten Filament Tubes

The first mentioned factor, that of filament voltage, if carefully controlled will pay economic and operational dividends. As illustrated in Fig. 1, an extremely small change in filament voltage results in a considerable change in filament life. The possibility of increasing tube life by reducing filament voltage and consequently filament temperature is the result of the fact that bright-tungsten-filaments (water-cooled tubes) may be operated at complete saturation. In other words peak currents amounting in value to the total emission available may be drawn continuously without damage to the filaments. Obviously, Fig. 1 shows theoretical filament life based on normal evaporation of filaments and applies to bright tungsten filaments such as are generally used in water cooled tubes. While they may not hold for every installation the ratios or relationships may be considered an average for a large number of tubes.

Note that the increase in life obtainable is considerable even at slightly reduced filament voltages. For the same reason a correspondingly larger reduction in life results from even slightly increased filament voltages.

Bright tungsten filament tubes used both in amplifiers and rectifiers, should be operated at the lowest possible filament voltage consistent with satisfactory operation; i.e., power output, tolerable distortion, hum, and carrier shift. All broadcast transmitters should be operated in accordance with good engineering practice. However, it is not necessarily good engineering or good economy to adjust the transmitter in such a way as to reduce harmonic distortion from, let us say, 3% to 1% if in so doing it becomes necessary to operate the water-cooled tubes at a 5% higher filament voltage. The application of the cold eye of logic in arriving at a compromise in this regard will pay dividends and should not result in distortion of an objectionable nature. Class B modulator stages should be watched carefully in this connection.

Tubes are designed to provide a certain amount of emission at certain input voltages. Obviously, if the whole amount of emission designed into the tube is not required it becomes possible for the user to obtain more than the life expectancy provided for in the design of the filament.

For the same reason as shown in Fig. 1 an increase in tube life resulting from a decrease in filament voltage must be accompanied by a decrease in the available emission. The relationship of emission and theoretical filament life appears in the curves. It is important to note that reductions in filament voltages, therefore, are recommended only in conjunction with reliable distortion measurements because of the possible flattening of positive peaks.

Knowing the operating conditions of the tube, it is readily possible to estimate in advance the approximate amount of filament voltage reduction that can be made. The peak emission current requirements for the usual types of operation will run approximately as follows:

TYPE OF OPERATION	REQUIRED PEAK EMISSION
Class B Audio	3-1/2 times d-c plate current
Class B Radio Freq.	8 times d-c plate current
Class C Telephony	9 times d-c plate current

Curves of filament voltage versus emission are published for many tubes, and are usually

*Source Material: (1) Operation and Maintenance of Transmitter Power Tubes (from the engineering notes of the Columbia Broadcasting System), 2nd Edition of the NAB Engineering Handbook; (2) FTR Handbook of Tube Operations issued by Federal Telephone and Radio Corporation; (3) A Simple Test for Defects or Damage to Vacuum Tubes, 2nd Edition of NAB Engineering Handbook; (4) Engineering notices of the National Broadcasting Company.

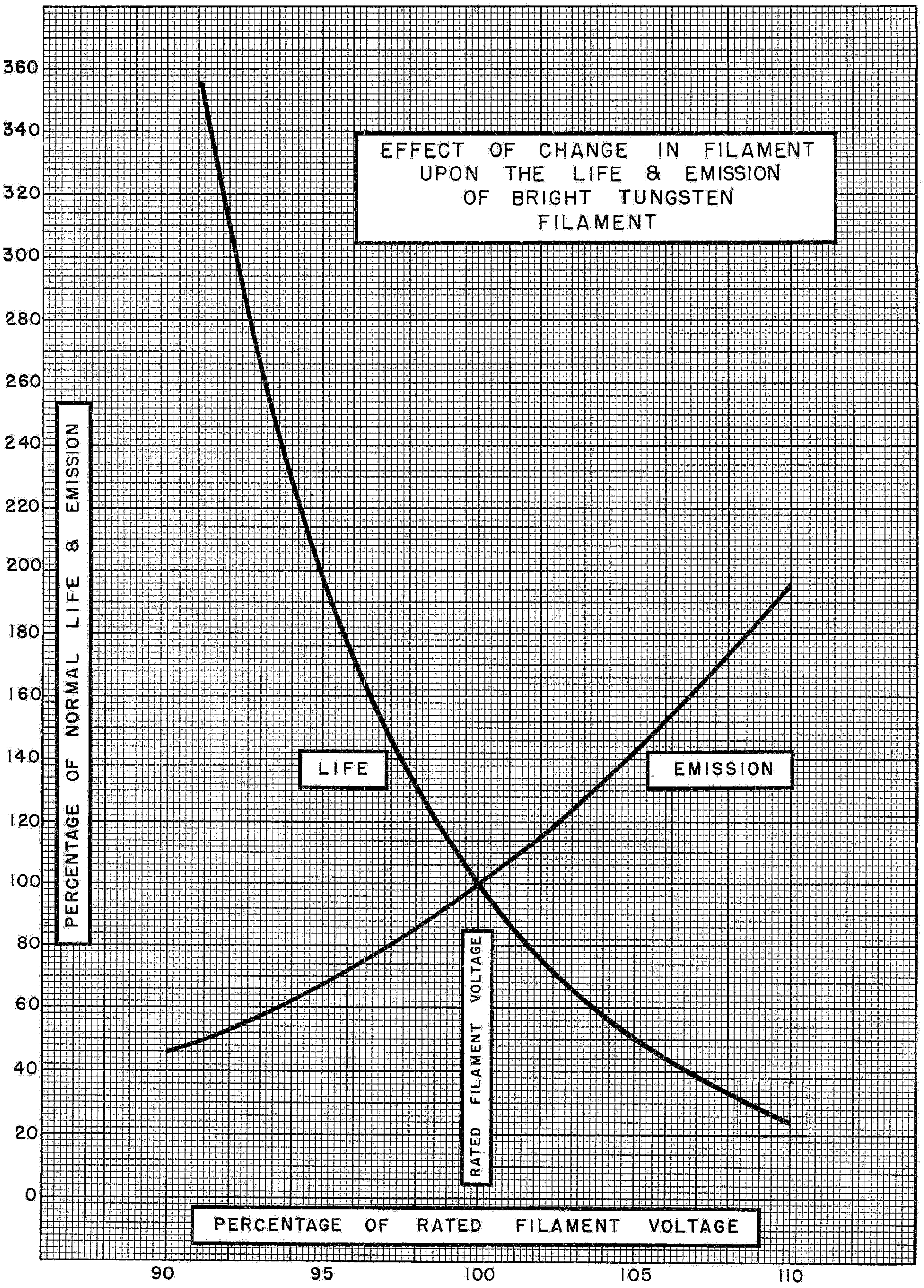

Figure 1 (Courtesy FTR)

available from the supplier. In the absence of these curves, an approximation of normal emission can be made by assuming 7.5 ma. emission per watt of rated filament power. Knowing the peak emission requirement and the total emission at rated filament voltage, it is possible to estimate from Fig. 1 a reduced voltage which will satisfy the peak requirements.

That filament voltage which fully satisfies the emission requirement can be applied with no sacrifice in transmitter performance. It may be advisable to determine if it is possible to reduce the filament voltage still further. Probably the best way to determine the proper voltage for a broadcast transmitter is to take a curve of distortion versus filament voltage at a high percentage of positive modulation. Limited emission of a class C tube causes a limiting and rounding off of the positive modulation peaks. Limited emission of a class B modulator tubes will cause a limiting and rounding off of both positive and negative peaks. If the filaments are heated by alternating current, some increase in hum level may result from a decrease in filament voltage. It is extremely important that filament volt meters be accurate and indicate the voltage at the tube and not at the source.

The following tabulation taken from Fig. 1 will indicate the relationship between filament voltage, total hours of useful life and unit cost per hour:

Filament Voltage		Total Hours of Useful Life		Unit Cost per Hour
90%		400%		25%
95%		194%		52%
100%	Normal	100%	Normal	100%
105%		50%		200%
110%		26%		415

Let us take a typical example. Suppose a tube has a rated filament voltage of 20 volts and theoretical average life expectancy of 3000 hours. Let us assume further that the tube costs the station $600.00. If the tube were operated with 20 volts on the filament the life expectancy would be 3000 hours and the cost of operation would be twenty cents per hour. Now let us assume that this same tube could be operated at 95% of the rated filament voltage. The theoretical average life expectancy would be increased to 5820 hours and the unit cost of operation would be reduced to about ten cents per hour.

Aside from the above discussion the following suggestions will also aid in prolonging the life of your tungsten-filament tubes:

1. Minimize anode dissipation by careful tuning of the transmitter
2. Be sure there is plenty of water flowing on the water-cooled anodes and plenty of air on air cooled anodes to prevent hot-spotting and gassing.
3. Keep plenty of air on the glass bulb--particularly on the seals where glass joins metal or leads go through--to reduce electrolysis and gas evolution from glass.
4. Switch leads every 500 hours, preferably once a week, when filaments operate on d.c.
5. During starting cycle be sure the instantaneous current does not exceed 150 percent of normal current.
6. Raise plate voltage in easy steps when starting.
7. Prevent damage caused by overloading the plate circuit. Use protective devices such as a fuse or relay.
8. Hard water (over 10 grains per gallon) should not be used for water cooling. Distilled water will reduce scale formation on anode.

Thoriated Tungsten And Oxide Coated Filaments

Thoriated tungsten and oxide-coated filaments used in intermediate sized air-cooled amplifier tubes and mercury vapor rectifier tubes <u>must</u> be operated within the rather critical filament voltage range specified by the manufacturer. Short life usually results and may be expected from the operation of such tubes much below or above (within approximately 5%) rated filament voltage.

It is very important that the filament voltmeters used be accurate and that the meters be connected in the circuit to indicate the voltage at the filament terminals--not, as is sometimes done, at a point beyond which there is an appreciable voltage drop in the filament supply bus. It is recommended that these instruments be calibrated periodically.

Thoriated tungsten filaments are operated at temperatures of such degree that evaporation is negligible. This means that the life of the tube is not controlled by the reduction of the tungsten wire and can not be extended by operation at reduced voltage as in the case of the water-cooled tubes.

In thoriated tungsten filaments the source of emission is a layer of thorium on the filament surface. During operation the thorium in this layer is constantly being removed by evaporation and bombardment and is constantly being replenished from within the wire. In order to maintain the balance between the loss and replacement of an active layer of thorium, therefore, an operation is required within a comparatively narrow range of temperature. This is evidenced by the fact that a reduction of only 1% in filament voltage may cause a loss of approximately 5% in emission.

Unlike bright-tungsten-filaments, dull emitter thoriated tungsten type filaments should never be operated at or near saturation. In other words, the peak currents drawn should not exceed more than one half of the maximum of which the filament is capable of emitting. These filaments are, therefore, designed to provide at least double the emission that would be needed in any normal class of operation.

Often thoriated filaments, due to momentary overloads or operation at improper voltages, provide low emission. A process generally known, but which is worth repeating, can often reactivate tubes subjected to the above strains. This process is merely that of flashing the filaments with 120% of normal filament voltage for about one minute, followed by ten minutes at normal voltage. No grid or plate voltage should be applied during these periods.

Concerning factors which enter into the life of tubes, number 5 is that number of times the current is turned on and off. It is always preferable to avoid intermittent operation of any filament. With tubes of 250 watt plate dissipation rating, or higher, it is suggested that filaments be turned off for standby periods of more than about two hours. For periods of less than two hours, the filaments may be reduced to 80% of the normal voltage. With tubes of less than 250 watt rating, the filament voltage should be removed for standby periods of more than 15 minutes.

Here are a few suggestions not discussed above which will help in extending the life of your thoriated tungsten filament tubes:

1. Don't overload the tubes. Use adequate protective devices such as fuse or relay. Heavy overloads are apt to evaporate the thorium surface from the filament, and permanently damage the tube.
2. Normal operating temperature for thoriated-tungsten-filament tubes is obtained by operating them at the rated filament voltage. Care should be taken to operate them at this voltage (except for standbys and when reactivating). Occasionally, under or over-loading will give longer life, but such operation should only be carried out after first consulting the tube manufacturer.
3. Increase the filament voltage progressively (only a small percentage at a time) when a tube no longer responds to reactivation. New filament transformers may be necessary for this operation.
4. Keep tubes well ventilated--with fans or blowers, if necessary.
5. Run at lowest possible anode current and voltage.
6. Minimize plate dissipation by careful tuning of the transmitter.

POWER TUBE MAINTENANCE

Inasmuch as broadcasting stations are required by the FCC to maintain a specified stock of spare tubes, some tubes may deteriorate on the shelf before being placed in service. This deterioration was at one time a serious problem to the users of large vacuum tubes. However, improvements in the methods of manufacturing tubes in the past few years have greatly improved this situation. For those users of large water-cooled tubes who may wish to either check new tubes upon receipt or who may wish to improve initial operation of tubes which have been kept on the shelf for extended periods the following checking on conditioning treatment is set forth for consideration.

Tests for Defects or Damage

Users of large vacuum tubes may save considerable time and money by utilizing a simple H. F. coil to rough-check all tubes before placing in service. The coil is a high-frequency 15 to 25 kilovolt source, operated from a 110 volt a.c. plug, and for this purpose may be small or of the type suggested below for conditioning. In use, the tip of the coil is placed against the external bulb surface of the tube far enough away from all external metal parts so that no arc can be struck externally. Any glow observed with the tube (glow may range from light pink to deep purple) indicates a poor vacuum and the tube should not be placed in service. If the tube is completely "down to air" (i.e. the internal and external atmospheres are alike), no color will be observed. However, to test this possibility, the coil tip should be placed against the bulb, completely clear of any external metal surfaces, but within an inch or less of some internal metal part. If the tube is nearly "down to air" an arc (usually a deep purple line) will appear, within the tube, between the metal part and the point of contact of the coil. A properly evacuated tube will show no glow or arc of any kind. (In rare cases when internal and external pressures are equal, no arc glow will be apparent.) Occasionally, however, the high frequency source will cause certain constituents of the glass or deposits thereon to fluoresce. This condition appears as a surface effect on the bulb and is readily distinguished from a glow within the tube.

This procedure, when used in conjunction with a continuity meter to check for open filaments or shorts between elements, provides a simple and adequate test to check shipment damage or manufacturing effects. Great care should, of course, always be exercised in handling vacuum tubes, since to damage one tube beyond any further use may negate operational savings built up over a period of months.

The above procedure to test the degree of evacuation does not apply to mercury vapor tubes since the results may be misleading with the presence of mercury in vapor form.

Conditioning Large Water-cooled Tubes

Before a tube that has stood for several months in storage is put into operation, it may be placed in an insulated mounting without the filament being lighted. The output from a high voltage, high impedance transformer, such as the Thordarson Type 2479[1], 1 kw, 110 - 25,000 volts, should be connected directly across from the anode to the filament and grid terminals tied together with the magnetic shunt adjusted for minimum voltage. The primary voltage should be applied until no arcing takes place over the transformer protective gap. The magnetic shunt is gradually pulled out until the maximum voltage is applied. This voltage should be left on for about ten minutes or longer if arcing takes place over the transformer protective gap.

If the tube envelope shows a hazy type of discharge on the first application of voltage, the tube may be gassy as a result of a small leak. The ten-minute treatment recommended above should clean up this haze condition unless the tube has developed a leak sufficiently large to cause trouble when put in service.

A second treatment consists of filament aging at the normal operating filament voltage without plate voltage and under normal water-cooling conditions, for several hours. This tends to clean up residual gases and cleans off both the filament and grid surfaces. It has been found that the above treatments are effective in cutting down initial flashing during operation, even though the tube is returned to the spare shelf for a week prior to being installed for regular service.

Tubes held as spares may be conditioned periodically to insure continuity of service when the tubes are inserted for use.

[1]This transformer is fairly large but may be used on the work bench. It is currently available from Thordarson on special order at approximately $165.00.

High vacuum rectifier tubes may be conditioned in the same manner as water-cooled tubes.

Mercury Vapor Rectifiers

Mercury vapor rectifiers should be preheated and tested under standard operating conditions as recommended by the manufacturer to determine that there is no vacuum leak and that all the mercury splashed on the tube elements during shipment and handling has been evaporated to the lower and cooler part of the envelope. After the initial acceptance test the tubes should always be handled and stored vertically to insure that excess mercury does not splash on the electrodes.

If a tube fails to rectify without arc backs after the recommended preheating time, another treatment has been found to be useful. This treatment consists of inserting approximately 1 megohm, such as 20 - 20 watt resistance units, connected in series with the anode of the tube which has shown arc-backs.

Using the maximum available voltage in the transmitter, the inserted resistance prevents excess currents from flowing through the tube which tends to arc back. The voltage should be left on for at least one hour, or until all visible mercury is removed from the anode. Room temperature air should be blown on the lower end of the envelope during this treatment. Then the tube may be operated under standard conditions without the resistance and put back into the spare tube cabinet for future use. Periodic tests during off hours will insure that spare tubes have not developed vacuum leaks.

Air Cooled Tubes

Air-cooled amplifier tubes should be placed in service as soon as received and then used alternately with other tubes of the same type. This will keep all the tubes "conditioned" and tend to decrease the possibility of a tube developing slow leaks. When placed in service the filaments should be operated at normal rated voltage for several hours in order to clean up residual gas and to activate fully the filament.

TELEVISION TRANSMITTER THEORY

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FCC REGULATIONS- Specifications of Performance:

a. Twelve channels are provided, from 54 to 216 mc.

b. The width of the standard television broadcast channel shall be 6 mc.

c. It shall be standard to locate the visual carrier 4.5 mc lower in frequency than the unmodulated aural carrier.

d. "Vestigial sideband" transmission shall be used with the idealized amplitude characteristic shown in Figure 5.1.

e. It shall be standard that a decrease in initial light intensity cause an increase in radiated power (negative transmission).

f. It shall be standard that the black level be represented by a definite carrier level, independent of the light and shade in the picture; that is, the "d-c component" must be transmitted.

g. It shall be standard to transmit the black level at 75% ±2.5% of the peak carrier amplitude.

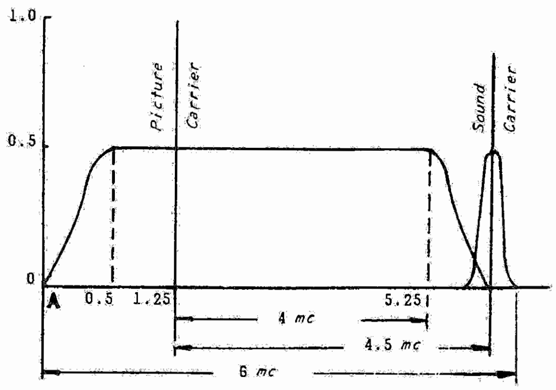
h. It shall be standard to rate the transmitter in terms of its peak power when transmitting the standard television signal. This represents peak of "sync" power. "Carrier power" has no significance when the d-c component is transmitted.

i. Amplitude modulation shall be standard in the visual transmitter, and the radio-frequency amplitude shall be reduced to 15%, or less, of the peak amplitude for maximum white.

FURTHER CRITERIA FOR TRANSMITTER PERFORMANCE have been specified by RMA:

a. The carrier frequency of the visual transmitter shall be held to within ±0.002%.

b. The peak-to-peak variation of transmitter output within one frame of the video signal due to all causes, including hum, noise, and low-frequency response, measured at both sync peak and pedestal level, shall not exceed 5% of the average overall sync peak



relative field strength not to exceed - 20 db at A

Figure 5-1 - Vestigial Sideband Amplitude Characteristic

signal amplitude. Thus all effects of hum and poor low-frequency response must be 25 db down.

c. The RMA recommended that, for normal transmission, the characteristic of transmitter output voltage versus brightness of subject be substantially logarithmic.

These regulations define, in general terms, the performance specifications of the visual transmitter. We see that we must operate in the medium- to semi-high-frequency part of the spectrum, radiating sidebands from dc out to 4.5 mc, using vestigial sideband transmission with amplitude modulation. The minimum depth of modulation must be 15 percent, and black in the picture must represent increased power. The carrier stability is defined, and the total of hum, noise, and low-frequency response on the carrier must be at least 26 db down. The specifications on amplitude linearity, as it concerns the transmitter only, are covered up by the logarithmic character of the camera output, and will be discussed later.

IDEALIZED TRANSMISSION CHARACTERISTIC

THE IDEALIZED TRANSMISSION CHARACTERISTIC of the FCC governs transmitter design in several respects. First of all, this vestigial sideband characteristic lops off part of the lower sideband, and ideally transmits out to 4.5 mc for the upper sideband and 1.25 mc for the lower. This is basically to save channel space, but it is also a receiver problem. In any vestigial or single-sideband transmission and reception system, we deliberately throw away an amount of information, quantitatively in direct relation to the percent of one sideband which is attenuated, since one sideband only yields 50 percent modulation. As long as one sideband is unaltered, the quality and resolution do not suffer. In the receiver, the i-f systems can be built with only half as much bandwidth, and the loss of signal is regained by adding a stage of video without any net increase in receiver cost.

BANDWIDTH

THE BANDWIDTH required to transmit a picture is basically a function of the number of picture elements which must be reproduced, and the speed at which an element must be scanned. In terms of picture constants, bandwidth is given by the relation

$$f_{max.} = \frac{w}{h} \frac{kfn^2}{2}$$

where w/h is the aspect ratio (4/3), k the utilization ratio (0.75), f the frame repetition rate, and n the total number of lines. The expression is for equal vertical and horizontal resolution. Thus for a 525-line, 30-frame picture, $f_{max.}$ equals 4.15 mc. For double-sideband transmission, a total bandwidth of 8.3 mc would be required.

There are several ways of thus squeezing a 525-line picture into a 6-mc channel. The most important are known as the RA, or Receiver Attenuation system, and tha TA, or Transmitter Attenuation system. The RA system is the one authorized by the FCC, and the frequency response through such a system is shown in the accompanying graphs, Figure 5-2. These curves are typical only of a high-level modulated-type of transmitter.

It is important to note the curtailment of high-frequency response in the receiver output, as compared to the response of the transmitter input. Studio equipment (and also the transmitter, to a lesser degree) is intentionally designed for greater response in the high end, to avoid any picture degradation, other than that actually imposed by the basic limitations of channel width. Since the receiver must adhere strictly to channel requirements, its reproduced picture will not contain as much detail as that actually transmitted. This factor should be taken into account when analyzing system performance.

The essence of proper vestigial sideband system operation is the fact, that, in the region of carrier out to 1.25 mc, the transmitter and receiver must have complementary phase and amplitude characteristics, as shown in (d) and (e), Figure 5-2. The transmitted signal has double amplitude response in the low-frequency end, which poses a monitoring problem, since a simple diode will not suffice.

Two other rather important matters arise in connection with sideband energy distribution. The first relates to the points designated as "a" in the FCC idealized characteristic. The regulations state that, at 1.25 mc below the visual carrier and at 4.5 mc above the carrier, the relative field strength shall not exceed 0.0005 to prevent adjacent channel interference. This value representa an attenuation of 60 db with respect to the 0.5-value of normal sideband amplitude. This appears rather severe, particularly at the point 1.25 mc below the carrier. However, Kell, Brown of RCA, and Keister and associates of GE, have made calculations and measurements which show that, for an average picture, the voltage components of sidebands in the vicinity of carrier ±1.25 mc are at least 40 db down. The transmitting equipment, then, must furnish the other 20 db. The RTPB has taken cognizance of this relationship, and has recommended to the FCC that equipment performance be judged on the basis of -20 db being satisfactory.

SIDEBAND ATTENUATION

SIDEBAND ATTENUATION may be obtained by two basic methods, which, for want of better terminology, may be referred to as the GE method and RCA method. The GE method uses low-level modulation followed by a multiplicity of broadband, linear r-f amplifiers, with each stage carefully tuned off-center so that the overall response gives the desired pass-band and attenuation. The RCA system uses high-level modulation, though not necessarily in the last stage, and then inserts a filter system between the PA stage and the antenna. In medium-power transmitters, *i.e.*, around 5-kw peak output, it is technically and economically feasible to build a video system with good frequency characteristic and a voltage output sufficient to modulate the grid of the final stage. If, however, we consider modulating the grid of a 40- or 50-kw output stage, we find that (1) the increased requirements of voltage necessitates water-cooled tubes; (2) water-cooled tubes with good modulator characteristics are not available; (3) overall cost is greater; and (4) we can get about the same power output by operating the 40-kw stage as a class B linear amplifier.

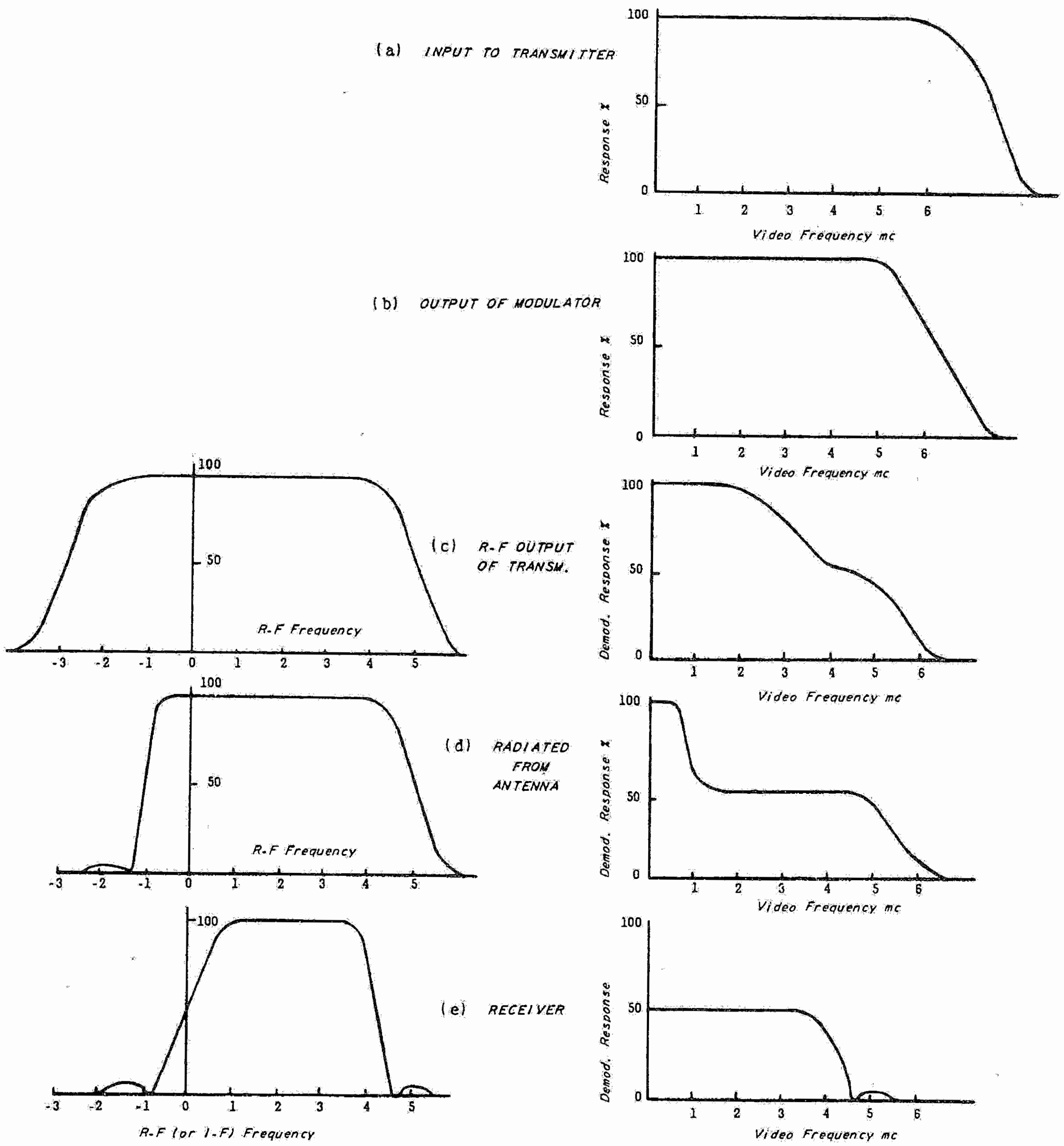


Figure 5-2 - Frequency Response Through a Television Transmitter System

The RCA system, as used in the TT-5A transmitter, modulates the final stage of a medium-power transmitter and the next-to-the-final stage of a high-power transmitter. In either case, there are tuned tank circuits between the points of modulation and the transmission line; and the response characteristic, which of course must be broadband, can be set with respect to the carrier frequency in such a way as to aid in the attenuation at the points "a". This not only provides a safety factor for the vestigial sideband filter, but, what is more pertinent, such adjustment allows the PA to work into a higher impedance, since the pass band is smaller, and hence to be capable of putting out more power.

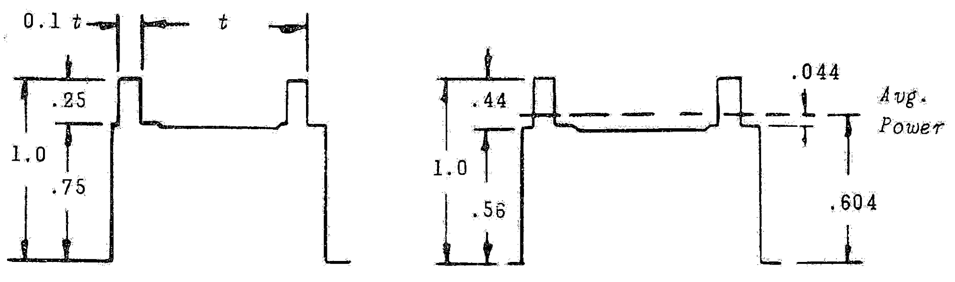
NEGATIVE TRANSMISSION

NEGATIVE TRANSMISSION means that the transmitter power decreases as the light intensity into the camera increases. In England and pre-war Europe, positive transmission was used, *i.e.*, increased power with increased light. The reasons for standardizing on negative transmission in this country are based on two facts: (1) Noise produces areas of lack of light, or black spots, which are considered to be less bothersome to the eye than bright spots, and (2) a much simpler receiver AVC system can be built if the blacks go in the increased-power direction. Since the blanking pedestal must represent black, the combined pedestal and sync pulse must modulate the transmitter in the "up" direction. Herein lies a disadvantage of negative transmission. To obtain increased output from a PA during the sync pulse interval, its grid must be driven into the positive or grid-current region, and the tendency is to saturate sync; or to reduce the amplitude of the sync pulses with respect to the rest of the signal. This, of course, does not happen with positive modulation; but a certain percentage of saturation of the whites is probably preferable to loss of sync voltage at the receiver. However, precautions can be taken to minimize sync saturation, and these will be discussed later.

There is another important advantage in using negative transmission, which ties in with the FCC specifications stating that the black level shall be transmitted at 75 percent of the peak carrier height. Since the blacks and the sync signals modulate in upward direction, the transmitter must hold peak power only for the duration of the sync pulse. Since the pulses are relatively short, about 6 microseconds for horizontal and 75 microseconds for vertical, they do not contribute appreciably to PA anode dissipation. The voltage and power relations are as shown in Figure 5-3.

The voltage curve shows the relationships of black level to sync tips as specified by the FCC, on the basis of sync tip height equal to unity. If we now re-plot this curve in terms of power, still using the sync tips as unity, it is seen that black level represents $(0.75)^2$, or 0.56, of peak power. Now, the sync pulses contribute additional power in proportion to their total duration and repetition rate. If a complete signal is analyzed, including both horizontal and vertical sync pulses, it is found that their total duration is equal to 0.1 of the repetitive time interval; therefore, the power they contribute is $0.1\ (1.00 - 0.56) = 0.044$, or a total of $0.56 + 0.044 = 0.604$, or 60% of peak of sync power. Anode dissipation is proportional to power output (assuming constant plate efficiency). Therefore, for a given tube complement with a given anode dissipation, the output can be increased 40%. If positive modulation were used, the output stage would have to be capable of handling continuous power equal to the peak power whenever a white picture came along.

In addition to the stipulation that the black level be 75% of the peak r-f amplitude, it is also specified that this black level represents a definite power level independent of changes in the picture itself. This means that the peaks of sync also represent a certain power level, since the ratio between black level and sync peaks is fixed. The picture transmitter could thus be rated (in power output) in terms



E_{rf} *vs. Time* P_{rf} *vs. Time*

Figure 5-3 - Voltage and Power Relations for Negative Transmission of either black level or sync peaks. The sync peak level was chosen because, at the time the old NTSC considered this matter, there was considerable discussion on what percentage of the signal should be devoted to synchronizing, and hence what percentage of the signal would be black level. So to avoid any possible confusion in the future, in case the percent of sync was changed, the peak of sync level was chosen.

To appreciate the significance of maintaining the sync peak at a certain fixed power or voltage level, it is necessary to look briefly at the d-c component of the signal; how it is obtained, and how it is handled. The discussion must start with the studio camera, and end with the receiver kinescope. It is in the output of the camera tube itself that the composition of the picture signal must be recognized and handled properly. In any picture scene which is to be transferred to electrical wave form, the signal can be said to contain two generalized components: One, the detailed information, which is the a-c component; and the other, the background, or d-c component. The background is simply the average illumination of the whole scene, taken over the interval of one complete picture frame. The image orthicon and the Farnsworth dissector tube are direct-coupled devices and automatically supply both components. The iconoscope, depending as it does, on a multiplicity of sensitized, capacity-coupled cells, puts out only the a-c component, referred to some arbitrary axis. The scene must therefore be viewed with an auxiliary light-sensitive device such as a photocell, whose output, suitably amplified, controls the bias of an appropriate tube in the unit in which the camera signal is combined with the sync signals (see Figure 5-4).

Thus, it is apparent that true d-c insertion is accomplished at the camera, and that fundamentally this consists of properly locating the camera a-c output voltage wave form with respect to the blanking pedestal or black level.

To preserve the background component, each amplifier between the camera and the kinescope could be d-c coupled. However, it is neither desirable nor necessary to go to these extremes. When a video signal passes through a capacity-coupled amplifier stage, the signal orients itself around a reference axis which is the average of the entire wave form. Thus, in the output of such a stage, black is no longer black in the sense that it represents a given voltage level, since that level will vary as the symmetry of the wave form varies. However, the relationship between the a-c wave form and the blanking pedestal has not been lost. In fact, the only thing that is lacking is the establishment of the black level to represent a given voltage or power output, as the case may be. Through relatively simple circuits, the black level may be re-established at any desired point in the television system. Not many years ago, before the value of establishing the black level in the transmitter was recognized, re-insertion was not applied until the signal reached the grid of the kinescope. At this point, a simple diode was connected across the grid resistor, and the desired reproduction was obtained.

LIMITING WAVE FORMS

Before pointing out the advantages of establishing the black level in the transmitter, let us look briefly at the *limiting wave forms* encountered in video signals, as illustrated in Figure 5-5.

The all-white and the all-black (with a white line) pictures represent the extreme cases, and typify the nature of the departure from symmetry which we nearly always encounter in video signals. The significance of this phenomenon

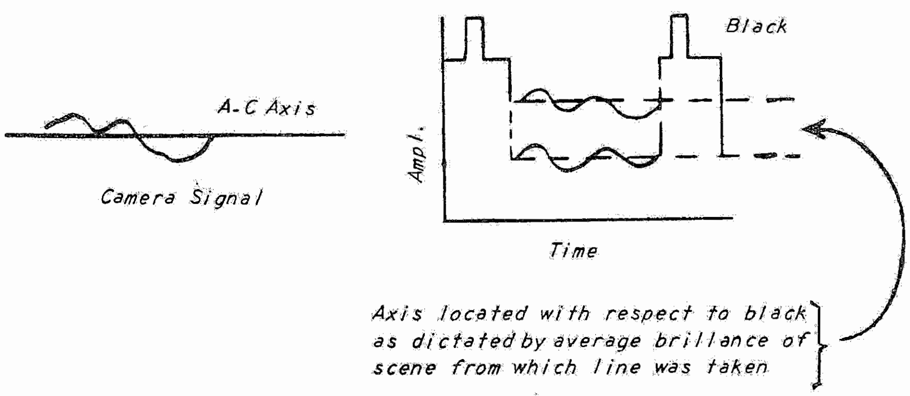


Figure 5-4 - Picture Signal Showing D-C Component

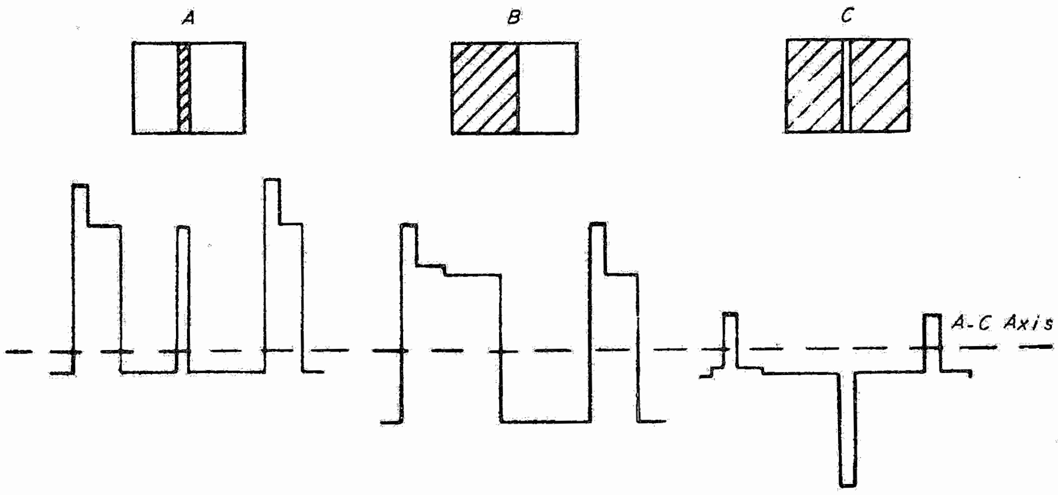


Figure 5-5 - Limiting Wave Forms in Video Signal

is seen when we think of these unsymmetrical wave forms in terms of voltage swings on the tube characteristic. The important fact, for transmitter design, is that a given tube will handle appreciably greater output with d-c re-insertion, although the actual voltage gain remains the same. Kell has shown, after analyzing the areas above and below the axis for the two extreme wave forms, that there is a ratio of 1.5 to 1 for the two cases.

D-C re-insertion allows 50 percent more intelligence to be transmitted in the sidebands, without increasing the carrier power, and hence without increasing the interference range of the transmitter. From the foregoing, it is seen that it is particularly advantageous to re-insert the d-c component in high-level stages, both video and r-f, where voltage-swing limitations are encountered. In present designs, this practice is followed, in that the d-c component is re-inserted in the grid of the last video or modulator stage, and preserved on the grid of the modulated stage by d-c coupling.

MODULATION

The last point in our standard specifications for television transmitter concerns modulation. Amplitude modulation for the video signal is specified and *grid modulation* appears to be the preferred method for the very-broad-band systems, and for systems requiring transmission of the d-c component. A similar analysis for narrow-band sound transmission will probably yield the opposite answer, and one of the deciding factors, in this comparison, would be distortion. High-output grid-modulated stages yield such high distortion as to be intolerable for high-fidelity transmission. In order to work on the truly linear portion of an r-f amplifier characteristic, it is necessary to stay almost entirely in the negative grid region, and thus a very considerable amount of power must be sacrificed. In television signal transmission, much greater distortion can be tolerated, and as a matter of fact the right kind of distortion is invited.

Referring to Figure 5-6, all camera tubes

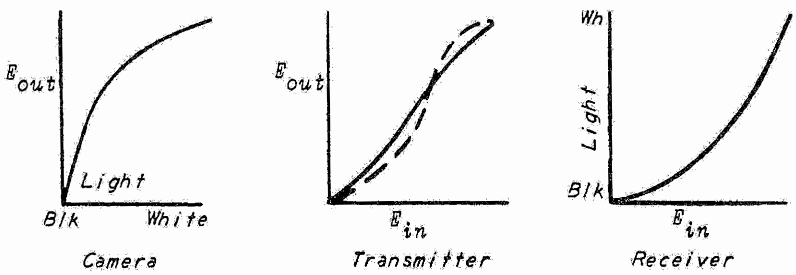


Figure 5-6 - Linearity Characteristics of TV System Elements have an amplitude-linearity characteristic which flattens off in the white region and is expanded in the black. Camera circuits could be designed to correct this, but it so happens that the receiver kinescope has an opposite, and nearly complementary, characteristic. Thus, at the receiver the response in the black region is contracted. Because, with negative transmission, noise, and interference bursts are in the black direction, this kinescope characteristic is very desirable, since it tends to saturate off noise peaks. It is very much akin to pre-emphasis of high's in a sound transmitter, so that the receiver can de-emphasize and attenuate noise. The RTPB recognized this feature, and recommended that the amplitude linearity of the transmitted television signal be "logarithmic", and the receiver "exponential". Furthermore, the transmitter and receiver characteristics can depart considerably from being complementary without affecting anything except the relative shades of black and white in the picture, the optimum value of which changes from one scene to another. It is apparent then, that the characteristic of the transmitter proper can be allowed to vary considerably from linearity, particularly in the white region, without appreciably changing the desired systems characteristic. It is important, however, to watch closely the linearity in the black region, since the sync signal must be held to +0, -2 percent of specified amplitude.

TERMINAL EQUIPMENT THEORY
THE TELEVISION SYSTEM

Reprinted from the RCA Manual for Television Technical Training Program through courtesy of the Radio Corporation of America

PART I

A REVIEW OF THE BASIC CONCEPTS OF THE SYSTEM AND THEIR EFFECT ON THE DESIGN OF STUDIO EQUIPMENT

INTRODUCTION

The problems which arise in the design of television equipment involve a branch of electronics which is strange to many technicians entering the field for the first time. This is especially true of those who have long experience in the field of sound broadcasting and reproduction, but whose education and experience antedate the post-war television boom as well as the war-time period, during which were developed so many of the techniques of electronic pulsing circuits. Basically the art involves the generation and reproduction of high-speed transient phenomena at both regular and irregular intervals and the consequent need for understanding circuits used for amplifying and transmitting wide bands of frequencies. Not only are the techniques new, but a whole new language has been developed to aid in their expression. It is hoped to provide here at least a brief glimpse of this new field and its language, thus helping the beginner to a firmer grasp of the tools he must now employ.

LIMITATIONS

No true appreciation of any system can be realized without some understanding of its basic limitations, and a discussion of the television system should therefore begin by reviewing these. The most serious limitation of a television system, as in the case of an aural system, is "noise." The same phenomena that cause hum, crackle, and hiss in the background of a sound broadcast, cause bar-like shadows, random blotches, and "snow storms" in the background of a television picture. The word, noise, has been carried over from aural terminology into television terminology with the same connotation; thus, any spurious elements in a television picture are generally called noise. In reading the following discussions, it will be helpful to remember that much of the reasoning behind the methods used in the television system is based on the need to minimize the effects of noise.

Spurious noise components in the signal arise from three general sources, (a) shot noise and thermal agitation in vacuum tubes and other circuit elements, (b) pickup from associated or remote electrical apparatus, and (c) microphonics. The best means for minimizing noise is to maintain a high signal-to-noise ratio in all parts of the system; but where this is impossible, special circuits are a distinct aid in extending the useful range of operation.

Noise limits, among other things, the ability of the system to resolve fine detail. However, a more direct limitation on the resolving power of the system is the frequency bandwidth available in the transmission system. This limitation has commercial aspects of more significance than the technical aspects because of the limited room available in the radio spectrum. As a result, the decisions of the Federal Communications Commission effectively determine the limits of resolution within the noise-free service area of any station. Long years of field testing have shown that a six-megacycle channel will provide adequate resolution and at the same time will yield a reasonable number of channels.

Other factors which limit overall performance are the fineness of scanning apertures*, the degree of accuracy with which tonal gradations

* *The use of the word aperture in television probably arose from the use of scanning disks where the light passed through small holes which traversed the projected area of the scene. Small holes traversing closely spaced lines in the area were capable of greater resolution than larger holes traversing more widely spaced lines. Though scanning disks are no longer used, the term aperture is still applied to the scanning device in a general sense. In electronic television, the diameter of the "aperture" is simply the diameter of the scanning beam of electrons in the plane of the scanned image. Similarly the term aperture correction is applied to means (usually the use of special circuits) for compensating the picture signal for loss of resolution caused by finite size of the beam and by non-uniform distribution of electrons over the cross-sectional area of the beam.*

are reproduced, and the brightness range of which the reproducing device is capable. However, if it can be assumed that the transmission system between the pickup and reproducing devices is reasonably linear, then the problems arising from these particular limitations are confined largely to the pickup and reproducing devices themselves, and do not affect system considerations to the same extent as limitations described in the preceding paragraphs, and as certain economic factors do.

Economic factors usually limit the degree to which technological development is used to improve the quality of performance. Methods may be known by which some of the physical limitations of the system can be overcome, but sometimes such methods are not used for a long time after their discovery because means for applying them economically are not developed simultaneously. In other words, their use increases the cost of equipment excessively. This is especially true in the case of receiving equipment which must be produced in large quantities at low unit cost.

Such methods often do find their way into transmitting equipment, where low unit cost is not so important and where quality of performance is paramount. Quality is stressed in transmitting equipment to provide reliability and to reduce the need for including in the receivers complicated and expensive corrective circuits. Examples are circuits for automatic correction of scanning linearity, and clamp circuits for accurate re-establishment of black level, or d-c restoration, as it is often called.

STANDARDS

During the decade preceding the entrance of the United States into World War II, Radio Corporation of America carried on an extensive program of research and development in television which has been largely responsible for the formulation of the standards governing our present black-and-white system. The earliest work on standards was done through the medium of the Radio Manufacturers' Association. Much more extensive work on standards was carried on later by the National Television System Committee and the Radio Technical Planning Board, the former body being set up to deal exclusively with television standardizing problems and to bring about agreement among the several interested groups on suitable standards for recommendation to the FCC. With the approach of commercial broadcast service, the FCC adopted the recommendations of these bodies as the basis for tentative standards of good operating practice. Activity of the RMA has continued on television and its recommendations have been extended to cover much of the detail of studio and transmitter operation, and of receiver design. While a considerable portion of this material still exists only in the form of recommendations to the FCC, it will undoubtedly constitute the major part of the final standards.

One of the most important standards recommended is the one which describes the waveshape of the picture signal. This standard is outlined in detail in a drawing which is reproduced in Figure 3-5. Reference will be made to this drawing from time to time in discussing the system, and an attempt will be made to clarify the reasoning involved in establishing many of the specifications included in it.

SCANNING SYSTEM

The standard system of scanning in television is one in which the scene or image is traversed by the aperture in lines which are essentially horizontal, from left to right, and progressively from top to bottom. The aim is to have the aperture move at constant velocity both horizontally and vertically during actual scanning periods because this type of motion is simple to duplicate in the reproducing aperture and because it provides a uniform light source in the reproducer. At the end of each line the aperture, or scanning beam, moves back to the start of the next line very rapidly. The time occupied by doing this is called the *fly-back* or *retrace* period. In a similar way, the beam moves from the bottom back to the top after the end of each picture scan. Motion during retrace periods need not be linear. The complete traversal of the scene is repeated at a rate high enough to avoid the sensation of flicker. This rate has been set at 60 times per second because most of the power systems in the United States are 60-cycle systems, and synchronism with the power system minimizes the effects of hum and simplifies the problem of synchronizing rotating machinery in the television studio (film projectors) with the scanning.

It has appeared rather recently that the choice of 60 cycles for the vertical scanning frequency was a fortunate one for another reason. The progress of the art has included means for obtaining brightness levels in the reproduced pictures which are appreciably greater than those used in motion picture theatres. It is well known that the threshold of flicker increases as the brightness increases. Thus, 48- or 50-cycle flicker would be noticeable to some observers at modern brightness levels in television receivers. Persistence of vision varies in different people, and those whose persistence characteristics are short are conscious of the 60-cycle flicker in the bright pictures on some present-day receivers. Therefore it appears that a still higher vertical frequency would be desirable if other factors would permit. Needless to say, the interline flicker, mentioned later in connection with interlacing, is also less objectionable with the higher scanning rate.

Another important factor affecting flicker is the persistence characteristic of the screen material in the receiver. This can be made long enough to overcome any appearance of flicker, even with scanning rates less than 50 cycles per second, but, if carried too far, such long persistence causes ghost-like trailing after moving objects in the scene. Judicious choice of screen persistence is a great aid in reducing flicker.

Obviously the scanning apertures in the pick-up and reproducing parts of the system must be in exact synchronism with each other at every instant. To accomplish this, synchronizing information is provided in the form of electrical pulses in the retrace intervals between successive lines and between successive pictures. The retrace intervals are useless in reproducing picture information, hence are kept as short as circuit considerations permit, but are useful places in which to insert the synchronizing pulses. These pulses are generated at the studio in the same equipment that controls the timing of the scanning of the pickup tube, and they become part of the complete composite signal which is radiated to the receiver. Thus scanning operations in both ends of the system are always in step with each other. Synchronizing is discussed in more detail in a later section.

The number of scanning lines is the principal factor determining the ability of the system to resolve fine detail in the vertical direction. The number of scanning lines is also related to the resolving power in the horizontal direction because it is desirable to have the same resolution in both directions. Thus, as the number of lines increases, the bandwidth of the system must also increase to accommodate the greater resolution required in the horizontal direction. The present system employs 525 lines, a number arrived at after thorough consideration of the related questions of channel width and resolution by the N.T.S.C. and the R.T.P.B.

INTERLACING

One of the most interesting features of the television scanning system is the interlacing of the scanning lines, a scheme which is used to conserve bandwidth without sacrificing freedom from flicker. The sensation of flicker in a television image is related to the frequency of the illumination of the entire scene. It has no relation to the number of scanning lines nor to the frequency of the lines themselves. Therefore a system which causes the entire area of the scene to be illuminated at a higher frequency, even though the same lines are not scanned during successive cycles of illumination, results in greater freedom from flicker. Interlacing does just this by scanning part of the lines, uniformly distributed over the entire picture area, during one vertical scan, and the

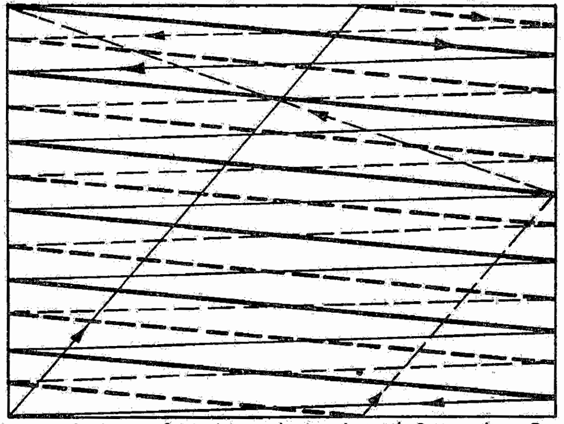


Figure 3-1 - Odd-line Interlaced Scanning System with 13 Lines. Consecutive Fields are Indicated by Solid and Dotted Lines, Respectively.

remaining part or parts during succeeding scans. Thus, without changing the velocity of the scanning beam in the horizontal direction, it is possible to obtain the effect of increased frequency of picture illumination.

In the standard two-to-one interlaced system, alternate lines are scanned consecutively from top to bottom, after which the remaining lines, that fall in between those included in the first operation, are likewise scanned consecutively from top to bottom. (Figures 3-1 and 3-2 il-

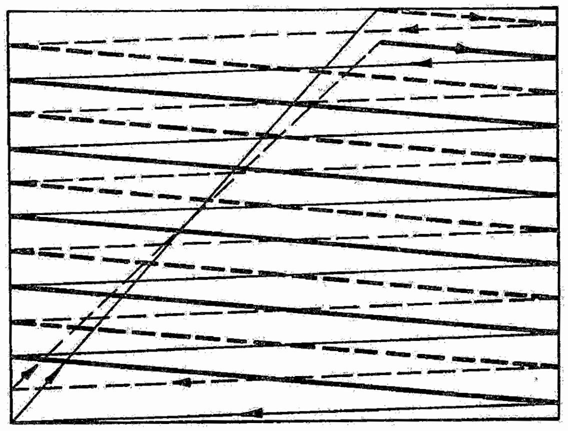


Figure 3-2 - Even line Interlaced Scanning System with 12 lines.

lustrate this principle). In the 525-line system, each of these groups, called a *field*, consists of 262-1/2 lines. Two consecutive fields constitute a *frame*, or complete picture, of 525 lines. Each field is completed in 1/60 of a second and each pair of fields or frame, in 1/30 of a second. The effect on the observer's eye, from the standpoint of flicker, is that of repetition of screen illumination every 1/60 of a second, yet the complete picture is spread out over 1/30 of a second.

The important result of interlacing is a reduction in the bandwidth of the frequencies generated in the picture signal, for a given value of limiting resolution, as compared to the bandwidth produced in a system using sequential scanning. This may be understood as follows. In either system, interlaced or sequential, the vertical scanning frequency must be the same and must be high enough to avoid the sensation of flicker. In the standard television system this frequency is 60 cycles per second. In a sequential system, *all* of the scanning lines must be traversed in the basic vertical scanning period. However, in the two-to-one interlaced system, only *half* of the scanning lines are traversed in the same period. Thus, obviously, the horizontal velocity of motion of the aperture in the interlaced system is only half of the velocity in the sequential system, and likewise the signal frequencies are reduced by the same factor.

Interlaced scanning has certain inherent faults among which are interline flicker, and horizontal break-up when objects in the scene move in the horizontal direction.

Interline flicker results from the fact that adjacent scanning lines are separated in time by 1/60 of a second, and that each line is repeated only at intervals of 1/30 of a second. It is apparent in any part of a scene where some detail of the scene is largely reproduced by a few adjacent scanning lines, and where the contrast in the detail is high. For example, the top edge of a wall which is oriented in the scene so as to be nearly parallel to the scanning lines might be reproduced by only two or three adjacent lines. The 30-cycle flickering of the line segments forming the edge of the wall would be quite noticeable. In the limiting condition, where the wall is exactly parallel to the scanning lines, the edge would be reproduced by only one line repeated at intervals of 1/30 of a second. This is probably the worst possible condition, but one which is encountered rather infrequently. The top and bottom edges of the raster nearly always produce objectionable interline flicker because they are nearly parallel to the scanning lines. Interline flicker, like any other type of flicker, is most objectionable in scenes where the highlights are very bright and the contrast is high. When the brightness and contrast are low, interline flicker becomes negligible.

Break-up exists when an object in the scene moves in the horizontal direction rapidly enough so that the total motion in 1/60 of a second is equal to one or more picture elements. Then vertical edges of the object become jagged lines instead of smooth lines and there is apparent loss in horizontal resolution. This is roughly illustrated in Figure 3-3 where two rectangles

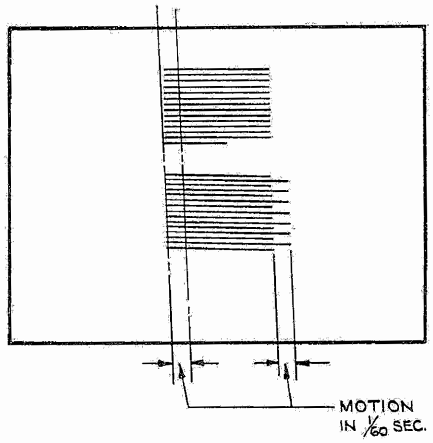


Figure 3-3 - Effect of Horizontal Motion on Resolution of Vertical Edges in 2-to-1 Interlaced System. Upper Object Stationary. Lower Object Moving to Right.

are shown, the upper one being stationary, and the lower one moving toward the right. The moving rectangle is shown as though it started moving from a position directly below the other. In the moving rectangle, signal is generated, *in both fields*, from the starting position of the left edge because of the storage of information in the pickup tube during the interval between fields. Thus the storage effect causes actual blurring of the trailing edge of a moving object. This is illustrated by the thin extensions of the scanning lines in the second field at the left side. The leading edge of the moving object may have a more definite jagged appearance because the storage effect in the pickup tube cannot fill in the spaces. In non-storage pickup devices, both edges will appear jagged.

The geometrical distortion, illustrated by the tendency for the moving rectangle to become rhombic, is characteristic of any scanning system, whether interlaced or sequential. It is similar to the effect produced by a focal-plane shutter in a photographic camera.

Further consideration makes it clear that higher ratios of interlacing would produce these troubles in aggravated form, which would be intolerable. Another objection to higher ratios of interlacing is an illusion of crawling of the scanning lines either up or down, depending on motion of the observer's eyes. The effect is extremely annoying and tends to distract the observer's attention from the scene.

The type of interlacing adopted for commercial television is known as *odd-line* interlacing. The total number of lines is an odd integer. Thus the number of lines in each of two equal fields is a whole number plus a half. In this system, the use of perfectly uniform vertical scanning periods (equal to half the product of the total number of lines and the period of one line) and constant vertical scanning amplitude, results in consecutive fields which are displaced in space with respect to each other by half a line, thus producing interlacing of the lines, as illustrated by the 13-line system in Figure 3-1. Specifically, as stated above, the total number of lines in the standard system is 525; the number per field is 262-1/2; the vertical scanning frequency is 60 cycles per second; the number of complete pictures (frames) per second is 30; and the horizontal scanning frequency is 60 x 262-1/2, or 15,750 cycles per second.

Interlacing may also be obtained when the total number of lines is an *even* number, but *even-line* interlacing requires that alternate fields be displaced vertically one-half line with respect to each other by the addition of a 30-cycle component to the amplitude of the vertical scanning sawtooth wave (see Figure 3-2). This frame frequency component must have a degree of accuracy that is impractical either to attain or maintain. Hence even-line interlacing is not used for commercial television.

One other factor has influenced the choice of the particular number of scanning lines. This is the need for an exact integral relationship between horizontal and vertical scanning frequencies. It has been the practice to attain this relationship by using a series of electronic counting circuits. To secure a high degree of stability, the characteristic count of each circuit was limited to a small integer less than ten. Thus the h/v frequency ratio was required to be related to the combined product of several small integers. In the RCA synchronizing generator equipment, for example, there are four such circuits counting the numbers 7, 5, 5, and 3 respectively. The combined product of these four numbers is 525, the number of lines per frame. The product of 525 and 60 is 31,500 which is the frequency of the master oscillator in the sync generator. To obtain the correct frequency for the horizontal scanning system, another counter circuit divides the master oscillator frequency by two to yield the required frequency of 15,750 cycles.

SYNTHESIS OF THE PICTURE SIGNAL

The basic part of the signal applied to the reproducer is the series of waves and pulses generated during the actual scanning lines in the pickup or camera tube. No matter what else is done in the equipment intervening between the two ends of the system, this basic part of the signal should be preserved in character with the greatest possible accuracy. However, during the retrace periods, the pickup tube may generate signals which are spurious or which at least do not contain valuable picture information. Furthermore, retrace lines in the reproducing tube itself, especially during vertical retrace, detract from the picture. It is therefore desirable to include in the picture signal, components which will eliminate spurious signals during retrace and the retrace lines themselves in the reproducer. These results may be obtained by adding synthetically some pulses known as blanking pulses.

Blanking pulses are applied to the scanning beams in both the camera tube and the kinescope in the receiver. *Camera blanking* pulses are used only in the pickup device and never appear directly in the final signal radiated to the receiver. They serve to close the scanning aperture in the camera tube during retrace periods. In orthicon tubes, the picture signal during retrace thus goes to reference black or to some level constantly related to reference black. This is a useful result to be discussed later. In iconoscopes, no such constant relationship to black exists during retrace, and the only function of camera blanking is to prevent spurious discharge of the mosaic during the retrace periods.

Kinescope blanking or *picture blanking* pulses are somewhat wider than corresponding camera blanking pulses. They become integral parts of the signal radiated to the receiver.

The function of the kinescope blanking pulses is to suppress the scanning beam in the kinescope (reproducing tube), or in other words, to close the aperture in the receiver during the retrace periods, both horizontal and vertical. They are simple rectangular pulses having duration slightly longer than the actual retrace periods in order to trim up the edges of the picture and eliminate any ragged appearance. They are produced in the sync generator from the same basic timing circuits that generate the scanning signals; hence they are accurately synchronized with the retrace periods. Typical waveshapes of a basic camera signal and blanking pulses are illustrated in Figure 3-4, A and B respectively. Only parts of two scanning line periods are shown, and the pulse in B is therefore a single *horizontal blanking pulse*. The result of adding the signals in A and B is shown in C, where it

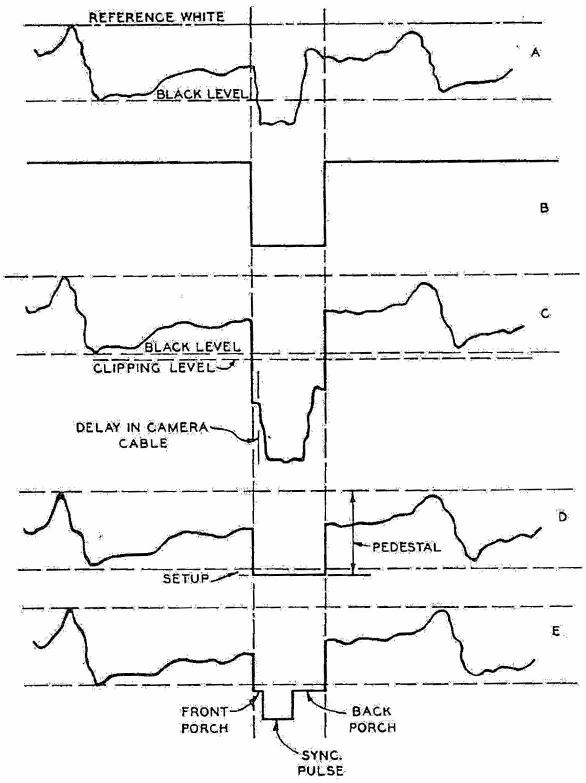


Figure 3-4 - Steps in Synthesis of Picture Signal.

may be seen that the unwanted spurious part of the camera signal has been pushed downward out of the territory of the basic picture signal. This unwanted part may now be clipped off and discarded, leaving the signal illustrated in D.

The blanking signal, shown only in part in Figure 3-4B, actually contains pulses for removing visible lines during both horizontal and vertical retrace periods. The horizontal pulses recur at intervals of 1/15,750 of a second and are only a small fraction of a line in duration; but at times corresponding to the bottom of the picture they are replaced by *vertical blanking pulses* which are just like the horizontal pulses, except that they are of much longer duration, approximately 15 scanning lines long, because the vertical retrace is much slower than the horizontal. The period of recurrence of the vertical blanking pulses is of course 1/60 of a second. Both horizontal and vertical blanking pulses, and their approximate relationship, are shown in diagrams 1 and 2 of Figure 3-5.

The picture signal shown in D of Figure 3-4 may be considered as partly natural and partly synthetic. It is important to point out here that the natural part, or basic camera signal, may contain certain noise components arising from the fact that the output of the pickup tube usually is not large compared to the noise threshold of the first picture amplifier stage or some other part of the system, such as the scanning beam in an image orthicon. On the other hand, the blanking pulses, or synthetic parts of the signal, are added at a relatively high-level part of the system and are therefore noise-free (at least in the transmitted signal). The importance of noise-free blanking pulses will become apparent in the discussions of other functions which they perform.

Details of horizontal blanking pulse shape are shown in diagram 5 of Figure 3-5. That part of the diagram below the point marked Blanking Level is a synchronizing (sync) pulse which will be considered later. The overall vertical dimension β is the maximum height of a blanking pulse. Thus the top horizontal line is Reference White Level, as indicated in diagram 3. The duration, or width, of the pulse must be sufficient to cover the horizontal retrace in the most inefficient receiver. Thus, the circuit limitations in such receivers set a minimum limit to the horizontal blanking width which was the basis for the RMA specification in Figure 3-5. This minimum is indicated by the width near the peak (lower end) of the pulse and is prescribed by the sum of two dimensions $x + y$, the value of which is 16.5% of the horizontal period, H. The impossibility of producing infinitely steep sides on the pulse is recognized in the greater maximum width (18% of H) allowed at the upper end of the pulse and in the obviously sloped sides.

Because of inevitable discrepancies at the extremes of the sides of the pulse, all measurements of pulse widths are made at levels slightly removed from the extremes of the sides. These levels are shown by dotted horizontal lines in diagram 5 of Figure 3-5, spaced 10% of β from top and bottom of the pulse.

Details of the vertical blanking pulses are shown in diagrams 1 and 3 of Figure 3-5. The width of the pulses is not limited by circuit considerations, as is the width of horizontal blanking. The limitation here is the requirement of television film projectors of the intermittent type, that the scene be projected on the pickup tube only during the vertical blanking period. The maximum period of 8% is ample for the operation of present-day film pickup systems, the criterion being that enough time must

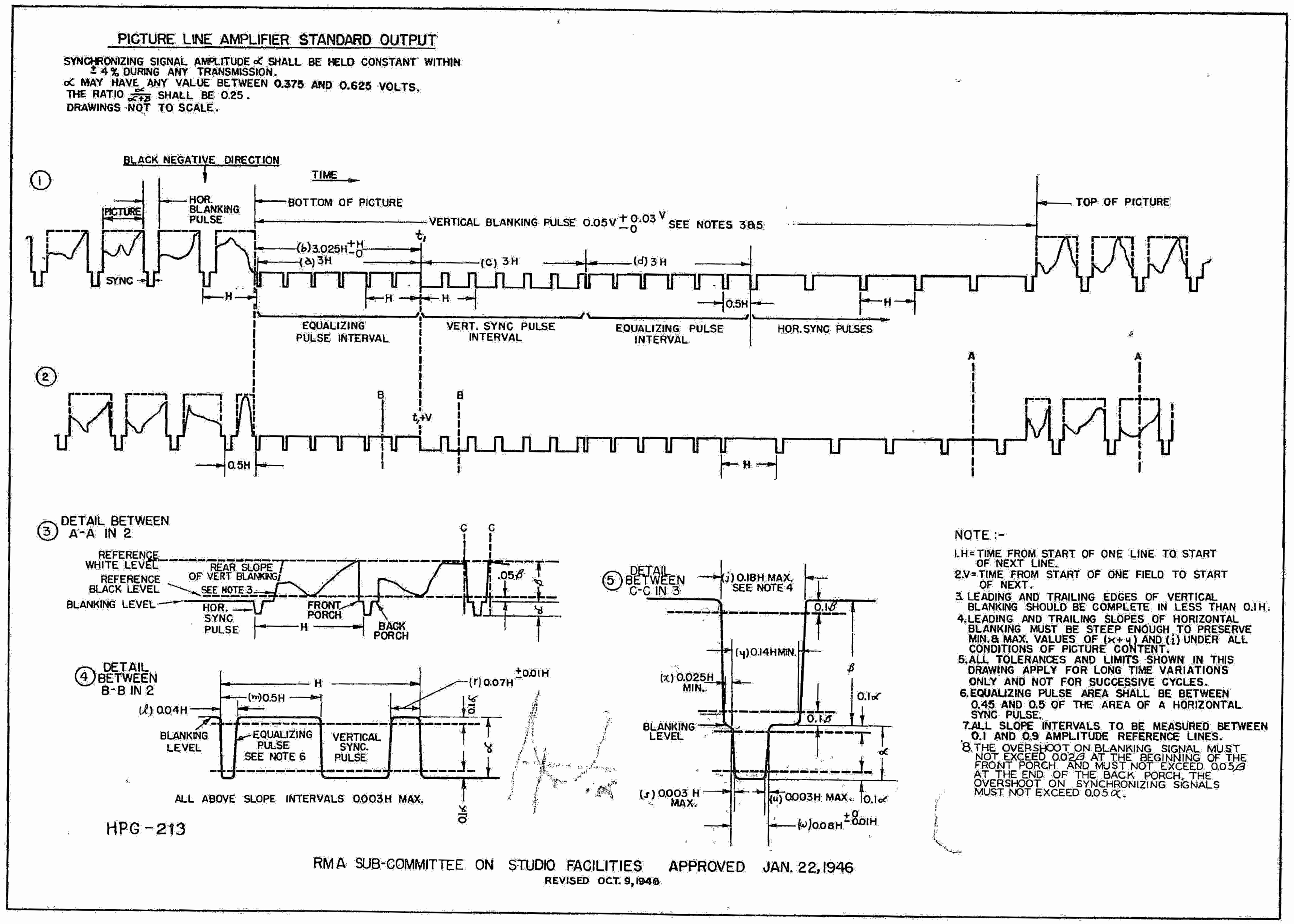


Figure 3-5 - Standard Television Signal

be allowed for projection so that there is adequate storage of photoelectric charges on the sensitive surface of the pickup tube. The minimum period of 5% is an indication of expected system improvements in the future, when it will be possible to reduce waste of picture transmission time in vertical blanking. The present usefulness of the 5% minimum is to require receiver manufacturers to maintain vertical retrace periods at less than 5% and thus avoid the need for modifying old receivers when improvements are made in the system. The problem of film projectors is discussed in a later section.

The final step in synthesizing the complete composite picture signal which goes to the modulator in the transmitter is to add the synchronizing pulses which are required for triggering the scanning circuits in the receiver. These pulses, like blanking pulses, are essentially rectangular in shape. The blanking pulses serve as bases or pedestals (inverted) for the sync pulses, as shown in E of Figure 3-4. Here is one of the most important reasons for having noise-free blanking. The synchronizing function in the receiver is a very critical one, and it is important that nothing be allowed to distort the sync pulses either in shape or timing, as noise during the blanking intervals would do. The nature of the vertical sync signal is rather complicated; it is not illustrated in Figure 3-4, but will be discussed later along with other details of synchronizing.

The sync signal is not added individually to the output of each camera, but is added at the studio output so that switching from one camera to another will not cause even momentary interruptions in the flow of synchronizing information to the receivers.

THE D-C COMPONENT OF THE PICTURE SIGNAL

The visual and aural senses differ in one important respect which places a requirement on the television transmission system which has no counterpart in the sound transmission system. The response of the ear to sound is actually a response to variations in air pressure. While the ear is very sensitive to rapid variations in pressure, it is completely unconscious of absolute values of air pressure, or of slow variations in pressure, as sound. In other words, there is a definite low limit to the frequency of pressure variations which the ear accepts as sound. Therefore there is no need, for a sound transmission system, to pass frequencies below the aural limit which is somewhere in the neighborhood of 15 cycles per second. The circuits may be a-c coupled without loss of essential information. Even the best of practical systems have a low-frequency cutoff at about 30 cycles, and most others cut off somewhere between 50 and 100 cycles.

The eye, on the other hand, is sensitive to absolute intensities of light and to slow variations of intensity. As the frequency of variation increases, the eye rapidly loses its ability to follow the changes and tends to produce a sensation which is an average of the variations. It is this averaging ability that enables the eye to interpret a rapid succession of still pictures as a portrayal of smooth motion. This phenomenon is the basis of both motion picture and television systems.

The important point, in the present discussion, is that the eye recognizes a slow change in light intensity. The period of the change may be a fraction of a second or it may be a minute, an hour, or a half-day in length. A television system must be capable of conveying these slow changes, no matter how long the period, to the receiver. The rapid scanning of the image of the scene in the camera produces a signal containing these slow changes as well as very rapid variations caused by the passage of the scanning beam over small light and dark areas of the image. The slow changes often have periods so long that they may be considered as d-c levels which simply change value occasionally. Hence, the signal is said to contain a d-c component. The television system must either pass the entire spectrum, including the d-c component, in each of its stages, or the signal must contain such information that it will be possible to restore the d-c component, which would be lost in an a-c-coupled system, when it finally arrives at the reproducer. Because of the well-known difficulties in constructing multistage d-c coupled amplifiers, it is desirable to use an a-c coupled system. It is fortunate that relatively simple means are known for d-c restoration, thus making possible the use of an a-c coupled system.

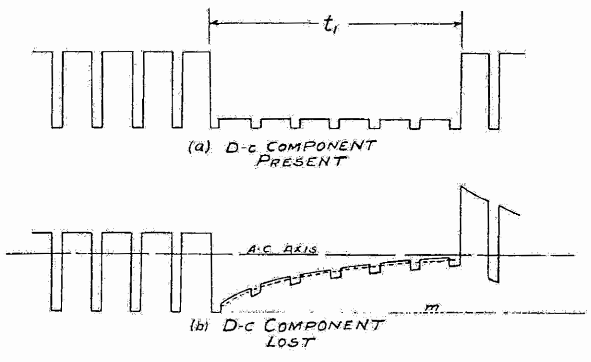


Figure 3-6 - D-C Component

Figure 3-6(a) illustrates a signal which contains a d-c component in the form of a temporary change in the amplitude of the pulses. The period t_1 embracing the low-amplitude pulses may be of any arbitrary length. The original signal is characterized by the constant level of the negative peaks of all the pulses, regardless of amplitude. After passing through an a-c coupled system (in which the time constants of the coupling networks are short compared to the period t_1), the signal becomes distorted approximately as shown in Figure 3-6(b). Here the negative pulse peaks no longer fall on a constant level, but the signal tends to adjust itself in a consistent manner about an axis called an a-c axis.

The a-c axis of a wave is a straight line through the wave, positioned so that the area enclosed by the wave above the axis is equal to the area enclosed by the wave below the axis. The broken line marked a-c axis in Figure 3-6(b) is actually the correct axis only for a wave composed of large pulses like the first four at the left. During the transient condition following the first short pulse, the line shown is not the true a-c axis, but represents the operating point of the amplifier in the a-c coupled system. The actual a-c axis of the short pulses (shown by the dotted line) gradually adjusts itself to coincide with the operating point of the amplifier. This adjustment is shown by the exponential rise of the signal during the interval t_1, but it is interrupted before completion by the resumption of the large pulses. Thence a second transient condition takes place, leading to a gradual restoration of the signal to its original form.

The departure of the pulse peaks from the original constant level indicated by the line m, is called *loss of the d-c component* or loss of "lows". It is interesting to note that this loss causes an increase in the peak-to-peak amplitude of the signal, a condition which is undesirable, especially in high-level amplifiers.

BLACK LEVEL

An absolute system of measurement must have a fixed standard reference unit or level. This rule applies to the problem of reproducing absolute light intensities. The simplest and most obvious reference for such a system is zero light, or *black level* as it is often called. This is a reference level which can be reproduced arbitrarily at any point in the system. Now if the television signal can be synthesized in such a way that frequent short intervals have some fixed relationship to actual black in the scene, then it becomes possible to restore the d-c component by forcibly drawing the signal to a fixed arbitrary level during these intervals.

D-C INSERTION AND D-C RESTORATION

Because the blanking or retrace periods are not useful for transmitting actual picture information, they offer convenient intervals for performing special control functions such as d-c restoration as mentioned in the previous paragraph. If the peaks of the blanking pulses are coincident with black level, or differ from black level by a constant amount, then d-c restoration can be accomplished simply by restoring these peaks to an arbitrary reference level. Thus, in Figure 3-6(b), if the peak of each pulse can be restored to the line m, then the signal will appear as in (a) and the d-c component will have been restored. Small errors will remain, corresponding to the displacements in level between pulses, but these are usually negligible and in any case do not become cumulative. Hence the restoration is essentially complete.

It now becomes apparent that an extremely important step in the synthesis of the television signal is that of making the peaks of the added blanking pulses bear some fixed relationship to actual black level in the scene. It was pointed out previously that the peaks of these pulses are produced by clipping off unwanted portions of the signal, as illustrated in Figure 3-4, C and D. A second, and most important, function is performed when the clipping is controlled in such a way that the resultant peaks have the required fixed relationship to black level. This process of relating the blanking peaks to actual black level is called *d-c insertion*, or insertion of the d-c component. A subsequent process, later in the system, of bringing these peaks back to an arbitrary reference level is called *d-c restoration*. D-C restoration must be accomplished at the input of the final reproducing device (the kinescope) in order to reproduce the scene faithfully.

It is desirable to restore the d-c component at other points in the system also, because the process reduces the peak-to-peak excursions of the signal to a minimum by removing increases in amplitude caused by loss of the d-c component. In a similar way, it is possible to remove switching surges, hum, and other spurious signal components which have been introduced by pure addition to the signal. Maintaining minimum excursion of the signal is important, especially at high-level points in the system, in order to avoid saturation in amplifiers and consequent distortion of the half-tones in the scene. For a specific example, d-c restoration helps to maintain constant sync amplitude in high-level amplifiers. In other words, it makes possible economies in the power capabilities of amplifiers such as the final stage in the picture transmitter.

Diagram 3 in Figure 3-5 illustrates part of a typical picture signal including two horizontal blanking pulses. It may be seen that there is a distinct difference between actual black level and blanking level which is prescribed as 5% of maximum blanking pulse amplitude. This difference is usually called *setup* and its magnitude was set as a reasonable compromise between loss of signal amplitude range and the need for a tolerance in operating adjustment. Setup is desirable as an operating tolerance in the initial manual adjustment of the clipper in that part of the system where the d-c is *inserted*. It simply insures that no black peaks in the actual picture signal are clipped off.

The accuracy with which setup is maintained depends on characteristics of the pickup or camera tube. Some types of pickup tubes produce signals during blanked retrace periods which are the same as, or are constantly related to, black level. In systems where such tubes are used, the magnitude of setup may be held constant automatically at whatever value is determined in the initial manual adjustment of the clipper circuit. In general, pickup tubes employing low-velocity scanning, such as the image orthicon, provide this kind of basic black level information. The iconoscope differs from orthicons in this respect, because the secondary emission resulting from the high-velocity scanning produces a potential distribution on the mosaic in which black level is far from the level existing during the retrace periods when the beam is cut off. In fact, the difference between black level and blanking level varies continuously as the scene brightness changes, because the potential distribution caused by resettling of the secondaries likewise changes. Automatic maintenance of setup, or pedestal height, cannot therefore be obtained by reference to the signal during blanked retrace periods in the iconoscope, but may be obtained by reference to actual black peaks in the picture signal. Where such reference is not practical, a manual control may be readjusted from time to time to keep the setup at the required value.

SYNCHRONIZING

The horizontal and vertical scanning circuits in a receiver are two entirely independent systems, both of which require extremely accurate information to keep them in step with the corresponding scanning systems in the camera, where the signal originates. Because the duration of sync pulses may be rather short, these pulses may be added to the picture signal in such a way as to increase the overall amplitude of the final signal without increasing the average transmitted power level very much. Thus, simple amplitude discrimination can be used to separate the synchronizing information from the incoming composite signal in the receiver. It is, however, desirable that a second increase in amplitude should not be used to distinguish between horizontal and vertical sync. The reason for this is that a further increase in signal amplitude would make necessary an increase in the peak power rating of the transmitter or else would unnecessarily restrict the power available for the picture and horizontal sync portions of the signal.

A synchronizing system has therefore been chosen in which both vertical and horizontal pulses have the same amplitude, but different waveshapes. Frequency discrimination may then be used to separate them in the receiver. The shapes of these pulses and their relation to the blanking pulses are illustrated in detail in Figure 3-5. Figure 3-7 is a functional block diagram showing the steps necessary to "utilize" the sync signals.

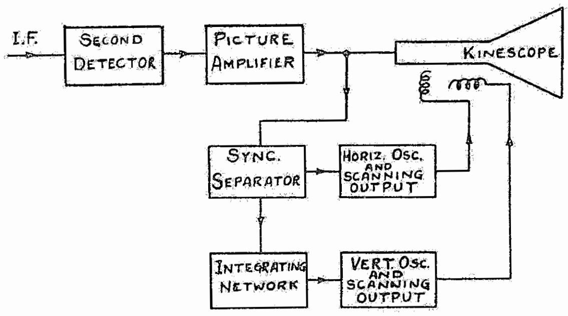


Figure 3-7 - Block Diagram of Picture-signal Amplifier and Scanning Circuits in Typical Receiver.

Diagrams 1 and 2 of Figure 3-5 illustrate a typical complete composite picture signal in the neighborhood of the vertical blanking pulse in each of two successive fields. Interlacing of the scanning lines is shown by the time displacement of the horizontal blanking pulses in one diagram with respect to those in the other diagram. This displacement is one-half of the interval of a scanning line ($H/2$).

All sync pulses appear below black level in an amplitude region which is sometimes called *blacker-than-black;* hence they can have no effect on the tonal gradation of the picture. Horizontal sync pulses are (except during the first portion of the vertical blanking interval) simple rectangular pulses, such as those appearing at the negative peaks or bases of the horizontal blanking pulses and during the last portion of the vertical blanking pulses. The duration of a horizontal sync pulse is considerably less than that of the blanking pulse, and the leading edge of the sync pulse is delayed with respect to the leading edge of blanking, forming a step in the composite pulse which is called the *front porch*. Correspondingly, the step formed by the difference between the trailing edges of sync and blanking is called the *back porch*. The purpose in forming the front porch is to insure that the horizontal retrace in the receiver (initiated by the sync pulse) does not start until after the blanking pulse has cut off the scanning beam. It also insures that any discrepancies which may exist in the leading edge of blanking do not effect either the timing or the amplitude of sync.

The choice of the nominal width of horizontal sync (0.08 *H*, see diagram 5 in Figure 3-5 was influenced by three factors. First, the width should be as great as possible so that the energy content of the pulses will be large compared to the worst type of noise pulses which may be encountered in the transmission process, thus providing maximum immunity to noise. Second, the width should not be greater than is necessary to meet the first condition, because average power requirements of the transmitter may thereby be minimized. Modulation of the picture transmitter is such that sync pulses represent maximum carrier power; hence it is desirable to keep the duty cycle as small as possible. Third, the horizontal sync pulses should be kept as narrow as possible so as to maintain a large difference between these pulses and the segments of the vertical sync pulses described in the following paragraph. Such a large difference makes it easier to separate the vertical sync from the composite sync signal. It has also been recognized that the back porch is useful for a special type of clamping for d-c restoration. Hence it should be as wide as possible.

Vertical sync pulses are also basically rectangular in shape, but are of much greater duration than the horizontal pulses, thus providing the necessary means for frequency discrimination to distinguish between them. However, each vertical sync pulse has six *slots* cut in it, which make it appear to be a series of six wide pulses at twice horizontal frequency, *i.e.*, wide compared to horizontal sync pulses. The slots contribute nothing to its value as a vertical sync pulse but do provide means for uninterrupted information to the horizontal scanning circuit.

Before and after each vertical pulse interval are groups of six narrow pulses called *equalizing pulses*. These also are for the purpose of maintaining continuous horizontal sync information throughout the vertical sync and blanking interval. The repetition frequency of the equalizing pulses and the slots in the vertical pulses is twice the frequency of the horizontal sync pulses. This doubling of the frequency does two things. First, it provides an arrangement in which the choice of the proper alternate pulses makes available some kind of a horizontal sync pulse at the end of each scanning line in either even or odd fields. Second, it makes the vertical sync interval and both equalizing pulse intervals exactly alike in both even and odd fields. The importance of this latter result will become evident in following paragraphs. It is important to point out that the leading edge (downward stroke) of each horizontal sync pulse and of each equalizing pulse, and the trailing edge (again the downward stroke) of each slot in the vertical pulses are responsible for triggering the horizontal scanning circuit in the receiver; hence the intervals of *H* or *H/2* apply to these edges.

Perhaps the most difficult problem in synchronizing, and the one in which there is the largest number of failures, is that of maintaining accurate interlacing. Discrepancies in either timing or amplitude of the vertical scanning of alternate fields will cause displacement, in space, of the interlaced fields. The result is non-uniform spacing of the scanning lines, which reduces the vertical resolution and makes the line structure of the picture visible at normal viewing distance. The effect is usually called *pairing*. The maximum allowable error in line spacing in the kinescope, to avoid the appearance of pairing, is probably 10% or less. This means that the allowable error in timing of the vertical scanning is less than one part in 5000. This small tolerance explains why so much emphasis is placed on the accuracy of vertical synchronizing.

The presence of a very minute 30-cycle component in the *vertical* scanning invariably causes pairing. The fact that the rasters produced in alternate fields are displaced with respect to each other by half a line means that the *horizontal* sync signal has an inherent 30-cycle component. It is this situation, and the need to prevent any transfer of the 30-cycle component into the vertical deflection, which account for the introduction of the double-frequency equalizing pulses before and after the vertical sync pulses. The vertical sync pulses are separated from the composite sync signal, before being applied to the vertical scanning oscillator, by suppressing the horizontal sync pulses in an integrating network similar to that illustrated in Figure 3-8.

Most receivers employ integrating networks of three stages instead of the two illustrated.

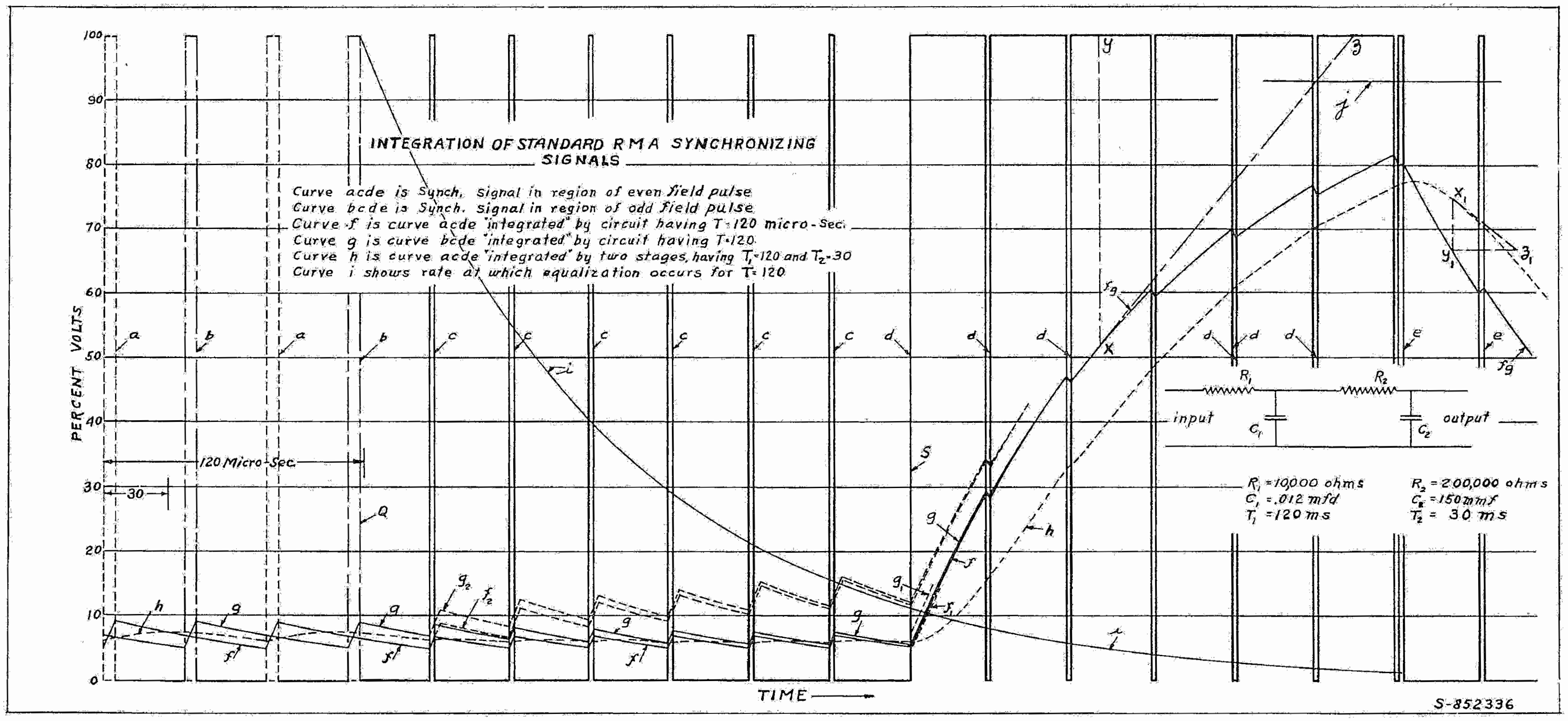


Figure 3-8 - Integration of RMA Synchronizing Signals however, the general character of the circuit action is clearly shown by the wave-form diagrams* in Figure 3-8. In simple terms, the equalizing pulses before the vertical sync pulses cause the integrating network to "forget" the difference between alternate fields by the time the vertical sync pulses begin. This is illustrated by the gradual convergence of curves f and g during the equalizing pulse interval, as the result of integration in the first stage alone. The effect of further integration in the second stage is shown by curve h, which is typical of the pulses applied to the vertical oscillator in a receiver. Thus, the 30-cycle component is effectively eliminated, from the standpoint of accurate timing of the start of vertical retrace, by the addition of the first set of equalizing pulses and the slots in the vertical pulse itself. The second set of equalizing pulses which follow the vertical pulse affect to some extent the impedance of the circuit to which the vertical scanning oscillator is coupled, and thus affect the amplitude of its output; hence these pulses help to provide more nearly constant output of the oscillator. Both sets of equalizing pulses contribute materially to the necessary accuracy of vertical synchronizing.

The width of an equalizing pulse is half the width of a horizontal sync pulse (see diagram 4 of Figure 3-5, and Figure 3-8). This width is chosen so that the a-c axis of the sync signal does not change at the transition from the line-frequency horizontal sync pulses to the double-frequency equalizing pulses. The curves f_2 and g_2 in Figure 3-8 illustrate the undesirable effect of making the equalizing pulses the same width as the horizontal sync pulses. There is a slight rise in the integrated wave during the equalizing pulse interval, which could cause premature triggering of the vertical oscillator in the receiver if the hold control were adjusted near one end of its range. This rise in the integrated wave results from the change in the a-c axis.

The width of the slots in the vertical sync pulses is approximately equal to the width of the horizontal sync pulses. The slots are made as wide as possible so that noise pulses or other discrepancies occurring just prior to the leading edge of a slot (*i.e.*, near the end of the preceding segment of a vertical pulse) do not trigger the horizontal oscillator. Premature triggering can happen if the noise pulse is high enough and if it occurs very close in time to the normal triggering time. Increased time-separation (a wider slot) reduces likelihood of such premature action. Here again, the requirements of special clamping also are met more easily if the slots are made as wide as possible.

A further important advantage of the RMA system of separating the vertical sync by frequency discrimination is that the integrating network is a potent factor in reducing the effect of noise on vertical synchronizing. Noise signals contain mostly high-frequency components; hence they are almost completely suppressed by the integrating circuit.

Differentiation, or suppression of the low-frequency components, of the sync signal before it is applied to the horizontal scanning oscillator is done sometimes, but it is not necessary, and has not been indicated in Figure 3-7.

The methods just described for synchronizing the scanning circuits in a television receiver are complicated by the need for transmitting the complete information over a single channel. In the case of the scanning circuits in the cameras, however, the situation is very different. The cameras and the synchronizing generator are so close to each other that there is no problem in providing as many wire circuits as may be desired. Therefore it is customary to use what are called *driven* scanning circuits in cameras and sometimes in picture monitors used with the cameras. Separate pulse signals, called *driving signals*, are produced in the synchronizing generator for exclusive use in the terminal equipment. Horizontal and vertical driving signals are completely independent of each other in the RCA system and are carried on separate transmission lines to the points of application. The driving signal pulses trigger directly the sawtooth generators which produce the scanning wave forms. This method reduces interlacing errors in the terminal equipment to the errors inherent in the driving signals.

Figure 3-9 illustrates a portion of the scanning lines appearing on a kinescope as a result of the application of a television signal composed of RMA sync and blanking pulses. The group of lines shown are those occurring in the neighborhood of the vertical retrace period including a few before and a few after the vertical blanking pulse. As noted on the diagram, the triggering of the lines has been displaced both vertically and horizontally so that the shadows produced by the sync and blanking pulses appear near the center of the raster rather than

* *Diagram prepared by A. V. Bedford, RCA Laboratories, for presentation to the N.T.S.C.*

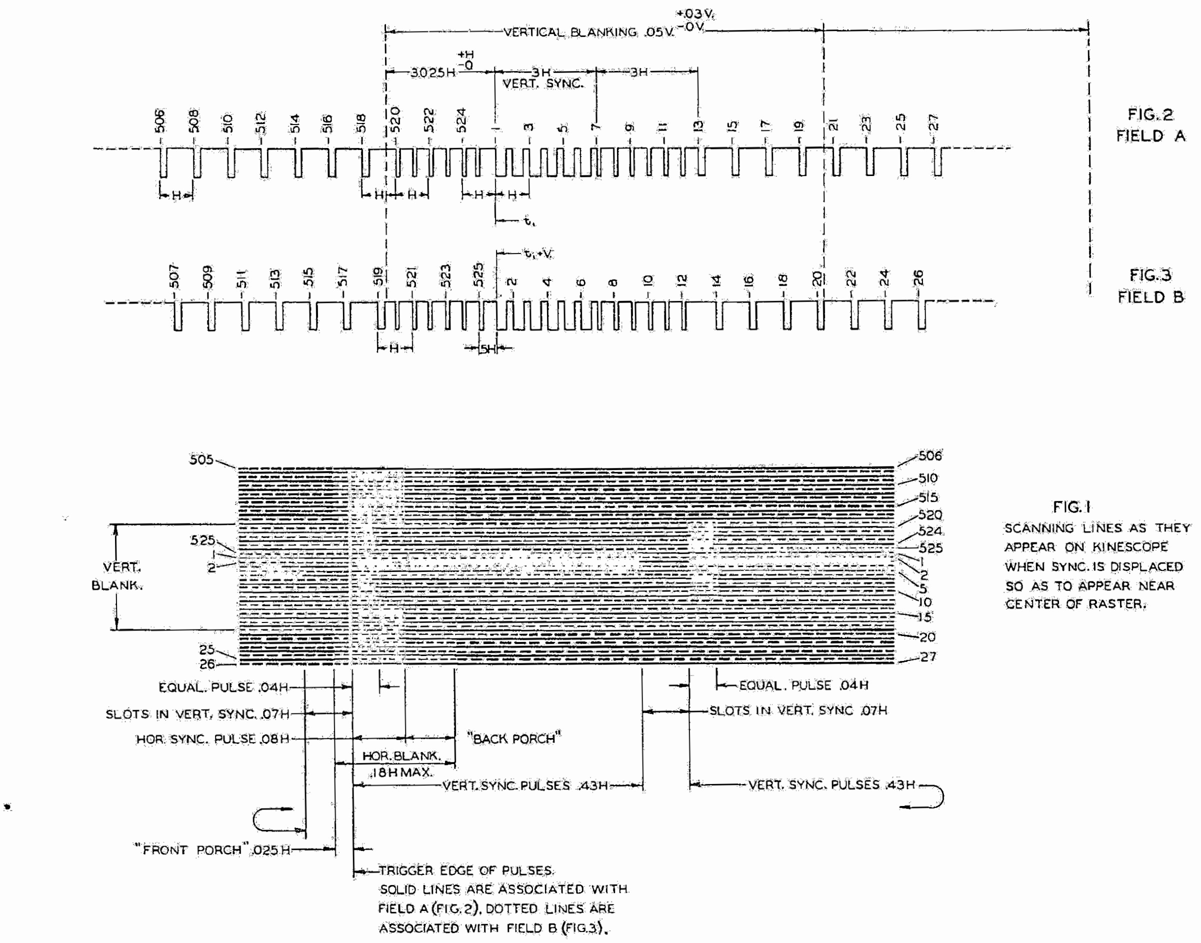


Figure 3-9 - Television Synchronizing Signal and Pulse Cross

in the normal positions at the edges of the raster. This displacement is brought about simply to clarify the illustration of the effect of the pulses on the raster.

The shadows produced thus are called a *pulse cross*. When expanded vertically so that individual scanning lines become easily apparent, the pulse cross becomes a ready means of checking the performance of the sync generator. The shadows produced by each different kind of pulses are indicated clearly on the diagram. With linear scanning, the horizontal dimensions of the shadows are measures of time or pulse width, and, because of the expanded scale, they provide a relatively accurate means of measuring pulse width. Furthermore, by counting appropriate lines, the numbers of equalizing pulses, slots, vertical sync pulses, etc. can be checked easily.

A useful piece of station test equipment can be made by modifying the deflection circuits in a picture monitor to provide the displacement of the lines and the extra large vertical expansion described.

AUTOMATIC FREQUENCY CONTROL OF SCANNING

The constant search for means of immunization against the effects of noise has brought about the development of automatic frequency control (afc) of the scanning circuits in television receivers. In triggered circuits, each scanning line (and each field) is initiated individually by a pulse in the incoming signal. In contrast to this, in an afc system, scanning generators are governed by stable oscillators which, in turn, are controlled by voltages obtained from phase comparison of the incoming sync pulses with the scanning signals themselves. The time-constant of the comparison circuit is usually made long, compared to the period of the scanning, so that random noise pulses have very little effect on the resulting control voltage, and correspondingly little effect on the scanning frequency. The fact that such afc circuits are keyed provides a further immunization factor by eliminating the possible effect of all noise pulses except those which coincide with the short keying intervals. The use of afc scanning circuits makes possible accurate synchronizing of a receiver under such bad conditions of noise that the masking of the picture by the noise renders it completely unusuable. Thus, failure to synchronize may be largely eliminated as a limiting factor in picture reception.

AFC may be used with both vertical and horizontal scanning circuits, but so far is being used commercially for horizontal circuits only. One reason for not using afc with the vertical circuits is that the time-constant must be very long to provide a stable control voltage. As a result, the circuit will not recover from an extended interruption of the incoming signal until an intolerably long time has elapsed. The frequency of the oscillator drifts during an interruption, and may not recover for a large number of seconds after the signal returns. During the period of recovery, the raster rolls over continuously at a decreasing rate until control is restored. The time-constant of the horizontal circuit, on the other hand, may be short enough so that recovery takes place in less than one field. Triggered scanning circuits, of course, recover from signal interruptions very rapidly, but they do not have the same high immunity to noise that the afc circuits have.

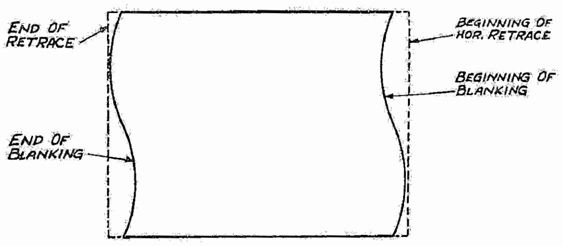


Figure 3-10 - Effect of Frequency Modulation of Horizontal Sync and Blanking on Shape of Raster in Receiver with AFC of Horizontal Scanning.

As a result of the use of afc circuits in receivers, a high degree of frequency stability is required in the horizontal sync and blanking signals. Frequency modulation of the horizontal pulses is intolerable because it causes the right- and left-hand edges of the blanked raster in the receiver, as well as vertical lines in the scene, to assume the same shape as the modulating wave. As shown in Figure 3-10, the border of the complete raster in the receiver is rectangular, but frequency modulation of the horizontal sync and blanking will distort the shape of the border produced by blanking. Frequency modulation by a 60-cycle sine wave is illustrated.

Horizontal retrace begins along a straight vertical line regardless of timing; and since this retrace is controlled by a stable oscillator in the receiver which is not responsible to short-time changes in sync timing, the presence of variations in sync timing and of corresponding changes in blanking pulse timing, will show as a displacement of the edges of the blanked raster. The frequency stability of the sync generator must therefore be at least equal to the stability of the oscillators used in afc receivers. The maximum rate of change of frequency allowable in a sync generator has been specified by RMA as 0.15% per second. This is a rather strict tolerance, as indicated by the fact that it allows a total displacement of only 1/32 of an inch (approx.) in a period of one field in a picture 10 inches wide.

FILM PROJECTION

The use of standard sound motion picture film for television program material offers a special problem which arises from the difference in the picture repetition rates use. For reasons explained previously, the rate used for television is 30 frames and 60 fields per second. The standard speed for sound film, both 16mm and 35mm, is 24 frames per second, and since each frame is projected twice, the pictune rate is 48 per second. The basic problem of reconciling the frequency difference has been met by using special projectors for television, in which alternate frames of the film are projected twice and the remainder are projected three times. In this way, 60 pictures are obtained in place of the usual 48, but the average speed of the film through the projector is unchanged; hence the sound take-off is entirely normal.

Another problem also presents itself in the use of intermittent film projectors for television. The vertical scanning period occupies from 92% to 95% of the total period. If the projected image is to be thrown on the pickup tube during the scanning period at all, it must be for the entire time so that all parts of the area will be subject to the same lighting conditions. Such an arrangement would leave only the vertical retrace period (5% to 8% of the total, or approximately one thousandth of a second) in which to pull down the film to the next frame. 35mm film will not stand up under accelerations produced by sprocket-hole pull-down in such a short period; hence some other scheme must be used. The method which has been adopted for use with intermittent projectors makes use of the storage property of certain kinds of pickup tubes, such as the iconoscope. The frame of film is projected with very intense illumination during the vertical blanking period only, while neither the pickup tube nor the receiver is being scanned. Then the light is cut off and the pickup tube is scanned in the absence of any optical image from the film. The signal generated during this scan results from charges stored on the sensitive surface during the preceding flash of light. While the light is cut off during the scan there is ample time to pull the film down before the next flash of light, without exerting destructive forces. The pulses of light may be obtained by chopping the output of a continuous source with a rotating disk, or (with a special type of arc lamp) by pulsing the source itself by electronic means. The storage properties of pickup tubes for this purpose must be sufficiently good so that dissipation of the stored charges is negligible between light pulses. Appreciable dissipation causes loss of contrast at the bottom of the picture.

Another solution to the problem of film projection in television is the use of a continuous projector, a type which produces a stationary image from continuously moving film by means of moving mirrors or lenses. This solution has not been accepted commercially so far because of practical difficulty in making the optical system sufficiently accurate to stop motion of the image completely.

The film problem in England, Europe, and other areas where 50-cycle power systems are standard, and where the television field frequency is also 50 cycles per second, is simpler in one respect, namely that it is not necessary to use the two-three ratio for projection of alternate frames of film. Instead, the film is projected as it is in theaters where each frame is projected twice. No attempt is made to compensate for the difference between the 24 frame taking speed and the 25 frame projection speed. The results are an approximate 4% increase in the apparent speed of motion of objects in the scene (which is probably negligible) and a slight rise in the pitch of all sounds. This latter effect is the more objectionable of the two, though generally it is not noticeable in speech and many other ordinary sounds. The change in pitch is undoubtedly noticeable to the trained musician in the case of musical sounds and must produce an unpleasant mental reaction to the music. However, no easy solution to the problem is known, and the situation is accepted without serious complaint. The other aspects of the film problem are not affected by the use of 50 fields instead of 60.

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The preceding discussion is necessarily brief and cannot serve as much more than an outline for further reading. There are many papers dealing more comprehensively with the details and problems associated with the various parts of the television system. References to some of these are included in the following bibliography. Most of the papers referred to also include references to others which, *in toto*, comprise a comprehensive list.

One book deserves special mention as a reference covering much of the engineering background of our television system. It is entitled, "Television Standards and Practice," (McGraw-Hill Book Co., 1943), and is essentially an abridged version of the proceedings of the National Television System Committee, as edited by Donald G. Fink. It includes a statement of the standards recommended by the Committee to the Federal Communications Commission, discussion of the investigations on which the recommendations were based, and references to pertinent papers.

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PART II

FUNDAMENTAL CIRCUITS

INTRODUCTION

The foregoing discussion of basic concepts shows that television circuits use vacuum tubes and components in ways that differ significantly from audio- or radio-frequency circuit applications. Sinusoidal wave forms are the exception rather than the rule. Usually the complex wave forms observed in television circuits are rectangular pulses, sawtooth shapes, or combinations of both. Vacuum-tube grids may be driven from a point well below cut-off potential into the positive region where grid current flows. The vacuum tube may operate as a switch in which total voltage and current values are used rather than small incremental quantities. Also time becomes an important factor since certain circuits must function in a particular manner with respect to time. The following notes are concerned with some fundamental television circuits employing concepts outlined above.

OVERDRIVEN AMPLIFIER

An overdriven amplifier is one in which the grid voltage is varied from a point below the tube cut-off voltage to some value in the positive region where grid current flows. This type of amplifier may be used as a limiting or clipping device or as a pulse amplifier. A circuit diagram is shown in Figure 3-11.

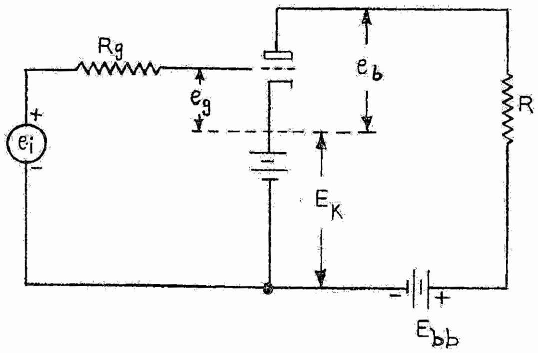

Figure 3-11 - Overdriven Amplifier

In the overdriven amplifier the following symbols apply:

$\bar{r}_p$ = d-c plate resistance = e_b/i_p. For $e_g = 0$ assume $\bar{r}_p$ constant

= 10,000 ohms approximately for 6SN7.

$\bar{r}_g$ = d-c grid resistance = e_g/i_g. Assume $\bar{r}_g$ constant for given tube

= 1,000 ohms approximately for 6SN7.

E_{co} = grid voltage for plate current cut-off

= E_{bb}/μ for triodes only.

R_g = grid limiting resistor which limits grid voltage to a value slightly positive with respect to the cathode.

Figure 3-12 shows an equivalent circuit and the resultant wave forms when a sinusoidal voltage e_i is applied to the grid. In this equivalent circuit, switches S_1 and S_2 are open when the grid voltage is below cut-off. They are closed when the grid voltage is positive with respect to the cathode.

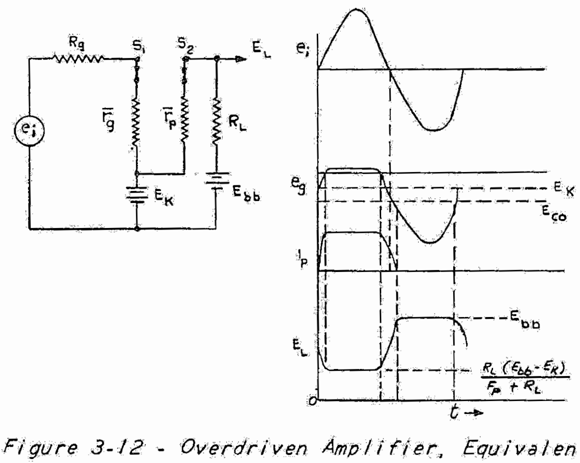

Figure 3-12 - Overdriven Amplifier, Equivalent Circuit and Wave Form.

When $e_i = 0$, the grid voltage e_g is equal to E_k. When $e_i - E_k = 0$, grid current flows and limits e_g to a slightly positive value. The grid voltage remains positive and constant, because of the drop across R_g (note that $R_g >> \bar{r}_g$), until e_i approaches the 180-degree point of the cycle. During the first half-cycle i_p rises rapidly to a value determined by $\overline{r}_p$, R_L, E_k, and E_{bb}, and then remains constant until e_g becomes negative. When $e_i + E_k$ is equal to the cut-off voltage, i_p falls to zero and no plate current flows for the remainder of the cycle. During the time that e_g is zero or slightly positive,

$$E_L = \frac{R_L \,(E_{bb} - E_k)}{\overline{r}_p + R_L}$$

When plate current is cut off (S_2 open),

$$E_L = E_{bb}.$$

In this circuit the sinusoidal input voltage has been clipped at both top and bottom to give a rough square-wave output voltage.

CATHODE FOLLOWER

A linear cathode-follower stage differs from the ordinary amplifier circuit in five ways: (1) the signal polarity is not inverted, (2) the gain is less than 1, (3) the output impedance is low, (4) the input impedance is high, and (5) the input capacitance is lowered. It may be used (1) after a pulse-shaping circuit to prevent loading of the circuit, (2) to drive tubes requiring grid power without altering the waveshape, or (3) as a device to match high to low impedances. It can deliver high currents to a low-impedance load without altering the waveshape. The basic circuit and equivalent circuit are shown in Figure 3-13 for incremental quantities.

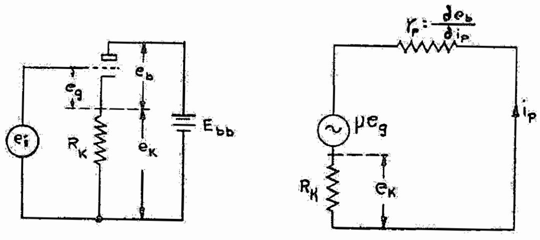

Figure 3-13 - Cathode Follower Circuit

From Figure 3-13 it may be seen that $e_g = e_i - e_k$; hence μe_g becomes $\mu(e_i - e_k)$. Then in the equivalent circuit

$$i_p = \frac{\mu(e_i - e_k)}{r_p + R_k}$$

and the output voltage e_o is

$$e_o = e_k = i_p R_k$$

$$= \frac{\mu(e_i - e_k)\, R_k}{r_p + R_k}$$

from which the gain A is expressed

$$A = \frac{e_k}{e_i} = \frac{\mu\, R_k}{r_p + (\mu + 1)\, R_k}$$

If the amplification factor μ is large, compared to 1, the gain A becomes

$$A = \frac{\mu\, R_k}{r_p + \mu\, R_k}$$

In another form, the gain may be expressed

$$A = \frac{\frac{\mu}{r_p} R_k}{1 + \frac{\mu}{r_p} R_k}$$

$$= \frac{g_m\, R_k}{1 + g_m\, R_k}$$

The last equation resembles the equation for the gain of an ordinary amplifier stage reduced by the factor $\frac{1}{1 + g_m R_k}$ The grid-to-cathode input capacity is reduced by the same factor, and the total input capacity becomes

$$C_i = C_{gp} + \frac{C_{gk}}{1 + g_m\, R_k}.$$

Similarly, the output impedance is reduced to

$$Z_o = \frac{R_k}{1 + g_m R_k}$$

A multivibrator is a circuit arrangement in which two tubes operate as switching elements to control the duration of current flow in the two load resistances. It may be compared to an oscillator, in that its action can be self-sustained. Such a multivibrator is called a "free-running" multivibrator. It may be synchronized to a desired frequency by either a sine wave of the given frequency or by a pulse whose repetition rate is equal to the desired frequency. There is also a type of multivibrator known as a "flip-flop", "one-kick", or "one-shot" multivibrator. This type of multivibrator performs one cycle of operation only when triggered by an external synchronizing signal.

Figure 3-14 is a circuit diagram of an unbiased free-running multivibrator. Capacitor

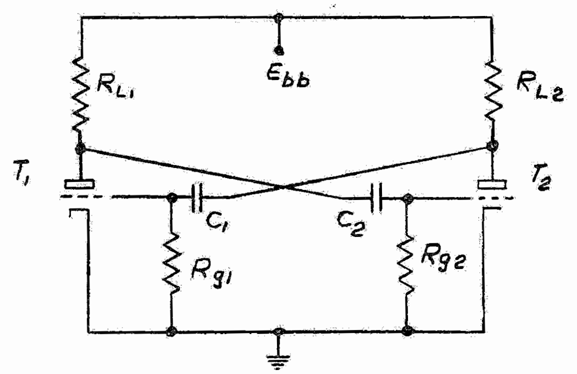

Figure 3-14 - Unbiased Free-running Multivibrator

C_1 couples the grid of T_1 to the plate of T_2. Similarly, C_2 couples the grid of T_2 to the plate of T_1. The circuit operates as follows: Suppose E_{bb} is applied when both tubes tend to conduct. Any small difference in circuit values or tube characteristics will result in one tube carrying more current than the other. Suppose more current flows in T_1. The greater voltage drop in R_{L1} is impressed on the grid of T_2, making that grid more negative and decreasing the current flow in T_2. The plate potential of T_2 rises, and this drives the grid of T_1 toward positive potential, causing T_1 to draw a still greater current. The effect is cumulative and results in T_1 carrying maximum current while T_2 is cut off.

The cycle of operation following the cut-off of T_2 is shown in Figure 3-15. The plate vol-

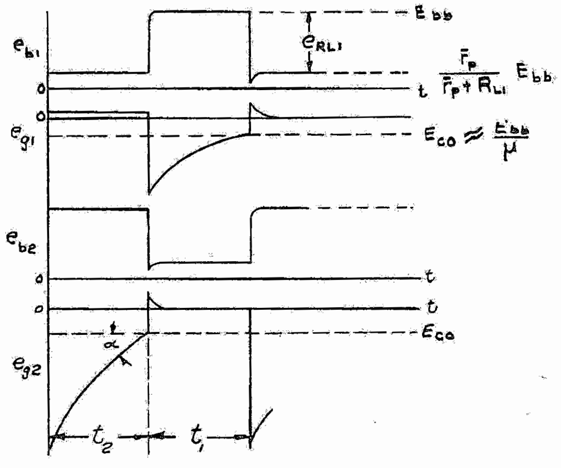

Figure 3-15 - Operating Cycle of Free-running Multivibrator

tage of T_1 drops to a value equal to $\frac{\bar{r}_p}{\bar{r}_p + R_{L1}} E_{bb}$. Since the grid of T_2 is coupled to the plate of T_1 through C_2, the grid voltage of T_2 also drops below zero by an amount equal to e_{RL1}. Grid voltage e_g tends to go positive because e_{b2} rises to E_{bb} when T_2 is cut off; however, C_1 charges quickly through R_{L2} and r_{g1} and leaves e_{g1} at approximately zero potential. Capacitor C_2 begins to discharge exponentially through R_{g2} and R_{L1} and $\bar{r}_{p1}$ in parallel. The equivalent circuit, for C_2 discharging, is shown in Figure 3-16.

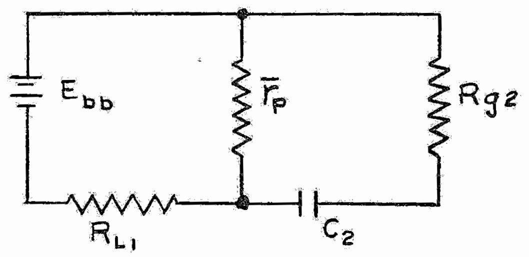

Figure 3-16 - Capacitor Discharge Circuit

The equation for the discharge of a capacitor is

$$e_c = E_o \epsilon^{-t/RC}$$

where e_c = voltage on capacitor at time t
E_o = total discharge voltage
RC = time constant of discharge circuit.

Since we are interested primarily in e_{g2}, we shall consider the voltage across R_{g2}. At the beginning of the discharge $e_{g2} = e_{RL1}$. The steady-state condition toward which e_{g2} is tending is zero volt; however, when e_{g2} reaches cut-off voltage, tube T_2 will begin to conduct, and tube T_1 will be cut off.

From Figure 3-10 it can be seen that the total resistance in the discharge path is

$$R_{g2} + \frac{\overline{r}_{p1}\, R_{L1}}{\overline{r}_{p1} + R_{L1}}$$

Stray capacity is neglected in the calculations, but tends to round the corners of the plate-voltage wave form as shown in Figure 3-15. From the foregoing we may write the equation for the voltage on the grid of T_2:

$$e_{g2} = e_{R_{L1}} \epsilon^{-t/RC}$$

$$E_{CO} = e_{R_{L1}} \epsilon^{-t_2/\left(R_{g2} + \frac{\overline{r}_{p1}\, R_{L1}}{\overline{r}_{p1} + R_{L1}}\right) C_2}$$

(specific equation to point of cut-off).

In the usual design problem all the constants are known or can be determined, with the exception of R_{g2} and C_2. The value of R_{L1} is determined by the amplitude of plate-voltage change desired. E_{CO} and $\overline{r}_{p1}$ depend upon the tube type. The time interval t_2 is known for a particular application and is the time that T_2 is not conducting. The product $R_{g2}C_2$ may be calculated from the equation for voltage across R_{g2}.

The operation of T_1 follows an identical cycle when it is cut off. The sum $t_1 + t_2$ of the cut-off periods determines the total period of the cycle. The frequency of the multivibrator may be varied by varying either or both grid resistors.

It will be noted in Figure 3-15 that e_g approaches E_{CO} at the angle α. Since this angle is small, any variation in tube characteristics or components causing a shift in E_{CO} will alter the cut-off period t because the point of intersection of e_g with E_{CO} will change. When it is essential that t remain nearly constant over a long period of operation, the grid may be returned to E_{bb} as shown in Figure 3-17.

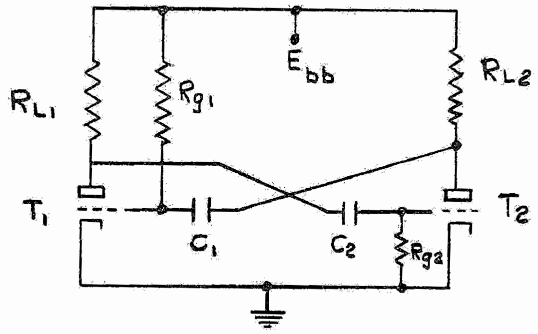

Figure 3-17 - Grid Return Circuit

The discharge wave form for e_{g1} is shown in Figure 3-13. In this example, e_{g1} is heading for E_{bb} instead of zero potential as in the previous case, and the angle α is large. Small

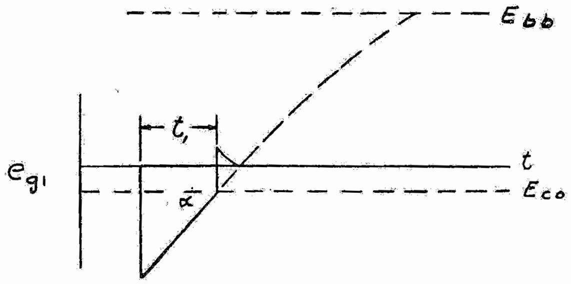

Figure 3-18 - Grid Voltage Wave Form

variations in E_{CO} will not greatly alter the intersection of e_{g1} and E_{CO}, and thus t_1 will remain very nearly constant. The equation for the discharge of C_1 becomes

$$E_{bb} + E_{CO} = (e_{R_{L2}} + E_{bb})\, \epsilon^{-t_1/RC}$$

An example of the so-called "flip-flop" multivibrator is shown in Figure 3-19 with the associated wave forms. Tube T_2 is normally conducting, and plate-current flow through R_k keeps T_1 cut off. When the grid of T_1 receives the trigger pulse, e_{p1} decreases and drives the grid of T_2 below cut off. T_2 remains cut off until C_2 discharges to the cut-off potential, at which

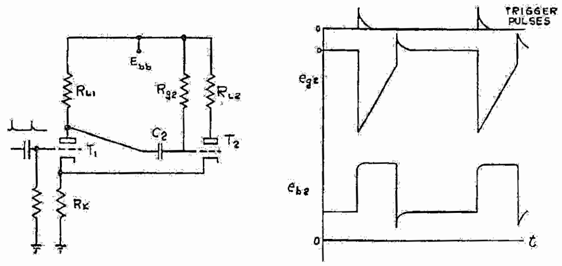

Figure 3-19 - "Flip-flop" Multivibrator

point T_2 resumes conduction until another trigger pulse is received.

CATHODE-COUPLED MULTIVIBRATORS

The multivibrators discussed thus far operate well up to pulse repetition rates of several thousand pulses per second. At higher repetition rates, stray capacity tends to cause unstable operation. To minimize stray-capacity effects and extend the stable range of operation, one can resort to cathode coupling between stages.

A cathode-coupled multivibrator is shown schematically in Figure 3-20.

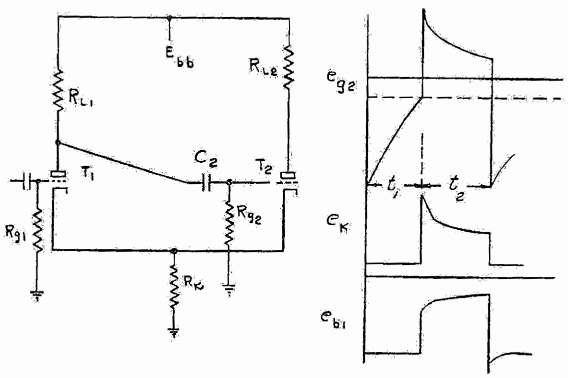

Figure 3-20 - Cathode-coupled Multivibrator

When $+B$ voltage is applied to the circuit, plate-current flow establishes across R_k a bias voltage common to both tubes. At the same time the voltage drop across R_{L1} is impressed on the grid of T_2, reducing the plate current in T_2 and lowering the bias voltage across R_k. With lower bias voltage, T_1 carries a larger plate current, and the resulting plate-voltage drop drives the grid of T_2 more negative. The process is cumulative and rapid, so that T_2 is cut off quickly.

Capacitor C_2 discharges in normal manner until the grid reaches cut-off potential. This cycle is shown, during t_1, in Figure 3-20. Then T_2 begins to conduct. The flow of T_2 plate current through R_k reduces the plate current in T_1 because of increased bias. The plate voltage of T_1 rises as the plate current decreases, and this voltage rise is coupled to the grid of T_2 through C_2. The grid of T_2 is driven positive by this cumulative process. Heavy current in T_2 cuts off T_1.

Now C_2 begins to charge through R_{L1} and the parallel combination of r_g and R_{g2}. It charges quickly until the grid voltage is reduced to cathode potential. Then T_2 grid current ceases, and C_2 continues to charge through R_{L1} and R_{g2}. At the end of the time interval t_2 plate current in T_2 has been reduced sufficiently to allow T_1 conduction. From this point on, the cycle is repeated.

This type of multivibrator is inherently unstable because neither tube can keep the other in cut-off condition. Also, the effect of stray capacity is reduced by interstage coupling to one grid only, and the input capacity of that grid is reduced by cathode-follower action. The controlling signal is coupled between stages by a low-impedance circuit in the cathodes, in which the effect of stray capacity is lessened. A multivibrator of this type may be operated in reliable manner at a pulse repetition rate of a million pulses per second.

CLIPPING CIRCUITS

Clipping circuits are used to eliminate undesired portions of complex wave forms by limiting the amplitude excursion in either the positive or negative direction, or in both directions. Clippers or limiters are usually applied in circuits in which pulses are formed and shaped to desired specifications.

Figure 3-21 illustrates a simple peak-clipping circuit which may be used to form a square

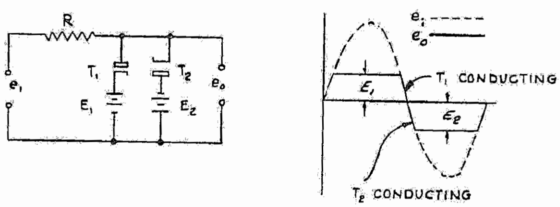

Figure 3-21 - Peak Clipping Circuit

wave from a sinusoid. Two diode elements, T_1 and T_2, are connected as shown. Bias battery E_1 keeps tube T_1 cut off until input voltage e_i increases in the positive direction to a value equal to E_1. Further increase in e_i causes conduction of T_1 and results in a voltage drop across R. As e_i increases, both the current in T_1 and the voltage drop across R increase so that the output voltage e_0 is fairly constant after e_i becomes slightly greater than E_1. A similar condition holds for the negative half-cycle of the input voltage. In this, the output voltage increases in the negative direction until bias E_2 is overcome. Then conduction in T_2 limits the output voltage to a fairly constant value. Wave forms for the cycle of operation are shown in Figure 3-21.

To produce, from a sine wave, a square wave with a short rise time by using a clipper of this type, it is necessary to connect several stages in cascade, inserting amplifiers between stages.

The overdriven amplifier discussed on previous pages may also be used for performing a clipping operation. Figure 3-22 is a practical

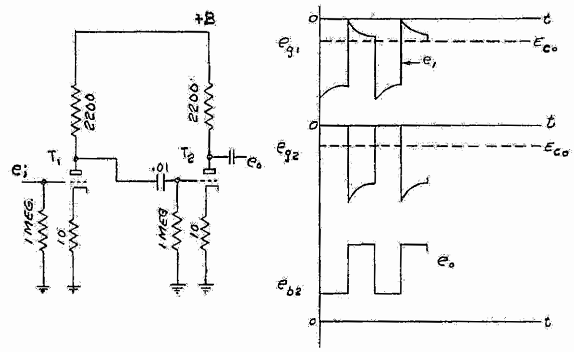

Figure 3-22 - Triode Clipper Circuit

triode clipper circuit utilizing a type 6SL7 tube. It may be desirable, for example, to remove overshoots or distortion in the tops of a square wave such as e_i in Figure 3-22. If e_i is symmetrical about the a-c axis, the grid of T_1 will assume a bias voltage, due to grid current, which will permit just the tip of the pulse to reach zero potential with respect to the cathode. If the pulse amplitude is sufficiently large, the negative excursions will drive the tube beyond cut off and eliminate the overshoot on the negative half-cycle.

The signal on the plate of T_1 consists of a square wave with overshoots eliminated in the positive half-cycle. By passing this signal through T_2, the overshoot is eliminated in the negative half-cycle, and a clean square wave is obtained in the output.

In the foregoing examples of clipping circuits, the action was symmetrical about the a-c axis. In some cases it may be desirable to clip only the tips of positive pulses, retain the tips, and eliminate the remainder of the wave form. Such a circuit and the appropriate wave forms are shown in Figure 3-23. Suppose the input voltage consists of alternate positive and

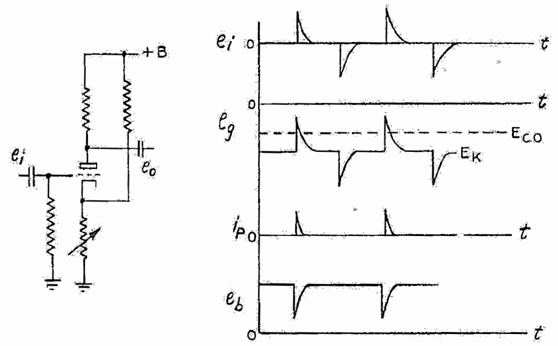

Figure 3-23 - Clipper Circuit

negative pulses obtained by differentiating a square wave. It is desired to clip the positive pulses midway between the axis and the tips. Voltage relationships are shown in Figure 3-23.

The clipping level is set by adjusting the bias in the cathode circuit so that, without signal, the tube is biased beyond cut-off. Only the positive tips of the input pulses cause plate current to flow.

In the clipping circuits described thus far, no attempt has been made to compensate for the inherent curvature near cut-off in the plate-current-cut-off type of clipper. Where the clipper is used to clip the blanking pulse and establish black level, as in Figure 3-4C, it is imperative that the slope of the grid characteristic curve remain constant to the clipping point; for this will prevent squashing of the video signal near black level and avoid change in the transfer characteristic. The linear clipper shown in Figure 3-24 accomplishes the desired result.

The linear clipper circuit includes a pentode V_1, a load resistor R_2 in series with a diode section V_2, and an additional load resistor R_1

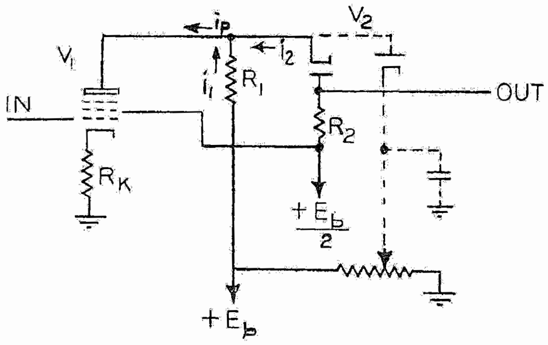

Figure 3-24 - Linear Clipper Circuit

in parallel with R_2 and V_2. The value of R_1 is approximately 20 to 30 times that of R_2. Both plate and screen supplies are regulated.

Figure 3-25 shows the characteristic curve for the linear clipper.

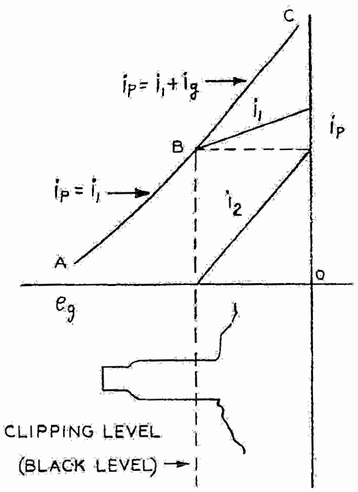

Figure 3-25 - Linear Clipper Characteristic

When V_1 is operating on the linear portion B-C, the plate current i_b is $i_1 + i_2$, and the plate voltage is less than $E_b/2$; hence V_2 conducts and causes signal current i_2 to flow in R_2. At point B the plate voltage of V_1 is equal to $E_b/2$. Between B and A the cathode voltage of V_2 is greater than $E_b/2$; therefore i_2 is zero, and there is no change in signal output voltage. Thus the blanking pulse is clipped at black level. Only the linear portion of the curve between B and C is used for the picture signal.

Cut-off in this clipper is abrupt. The use of a pentode tube permits an abrupt change in external load resistance without affecting plate current. As i_2 approaches zero, the ratio i_1/i_2 changes rapidly, resulting in a rapid change of V_2 cathode voltage in the cut-off region.

By adjusting the equivalent bias on V_1, point B can be made to coincide with black level.

A serious drawback of the linear clipper is capacity feed-through of transients. This trouble can be cured by connecting a second diode element, as shown by the dotted lines in Figure 3-24. By proper bias adjustment the second diode can be made to conduct at a potential just above cut-off of the limiter so that unwanted signals are shunted to ground when the limiter is inoperative.

BLOCKING OSCILLATORS

A blocking oscillator is a form of self-pulsed oscillator that is used as a simple means for obtaining a short pulse at some desired repetition rate. Figure 3-26 is a schematic diagram of a simplified blocking oscillator circuit. The coupling coefficient of the iron-core transformer T is very nearly unity. The connection polarities of the transformer must be as

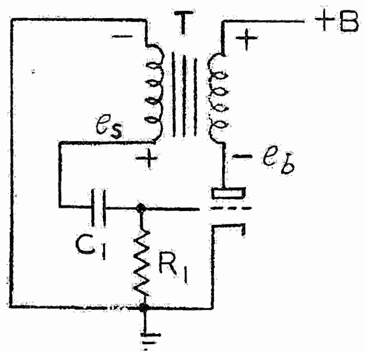

Figure 3-26 - Blocking Oscillator Circuit

shown. Wave forms for the operating cycle are given in Figure 3-27.

When B+ is applied and plate current starts to flow, a voltage develops in the primary winding, due to the inductance drop $L\,di/dt$. This voltage is coupled to the secondary so as to cause the grid voltage to rise in the positive direction. Thus, the plate current is further increased. The effect is cumulative and causes the grid to go positive quickly. As the grid

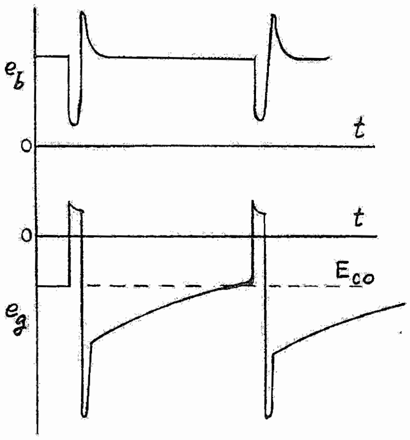

Figure 3-27 - Blocking Oscillator Operating Cycle

is driven positive, two actions occur: Grid current flows and charges C_1; and plate voltage is reduced to such a low value that further increase in grid voltage will not increase the plate current. The secondary voltage then ceases to increase, and C_1 begins to discharge. The discharge of C_1 lowers the grid voltage, causing a decrease in plate current. The induced voltage in the secondary is in the negative direction, due to the change in di_p/dt, and the grid is driven quickly below cut-off. Then C_1 is discharged through R_1 and the transformer secondary until the grid voltage is less than cut-off. When plate current starts to flow, the cycle is repeated. The unbiased blocking time is roughly 2 or 3 times C_1R_1, depending upon the transformer turns ratio, inductance values, and self-resonant frequency. The blocking oscillator frequency may be controlled by varying the bias on the grid or on the cathode, or by varying R_1.

The blocking oscillator may be synchronized by applying either a sine-wave or a pulse voltage across a resistor in the ground lead of the transformer secondary.

STEP-CHARGING CIRCUITS

A step-charging circuit is one in which the potential across a capacitor is built up in a series of steps. Its fundamental use is in a frequency-dividing system in which a blocking oscillator is triggered after a number of steps have been completed. Figure 3-28 shows a simple step-charging circuit and the resultant wave forms.

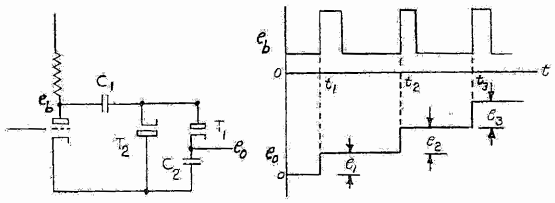

Figure 3-28 - Step-charging Circuit

Assume that the plate voltage e_b rises quickly at time t_1 and that $e_o = 0$. Diode section T_1 will conduct and charge C_1 and C_2. The voltage across C_2 is given by the equation.

$$e_1 = \frac{e_b C_1}{C_1 + C_2}$$

When e_b returns to its minimum value, diode section T_2 will conduct and discharge C_1. At time t_2 diode T_1 conducts again, charging C_1 and C_2. This time, however, the total change in e_b is not divided between C_1 and C_2 since C_2 has on it the voltage developed at time t_1. Let e_1 denote voltage to which C_2 was charged during the initial pulse. Then

$$e_2 = \frac{(e_b - e_1)\, C_1}{C_1 + C_2}$$

Each succeeding step may be calculated in the manner shown above. The voltage on C_2 during the preceding step must be subtracted from the peak-to-peak plate voltage in determining the amplitude of the next step.

At the end of a given number of steps, a blocking oscillator is triggered, C_2 is discharged, and the cycle is repeated.

NON-LINEAR MIXERS

In some television applications, specifically in the synchronizing generator, it is necessary to mix two pulses in such a manner that the resultant signal is not the algebraic sum of the two pulses. In effect, a third pulse is created which differs in character from the original pulses. Consider the circuit in Figure 3-29. The 6L7 tube is biased at such a high value that both grids must receive positive pulses before plate current can flow. During the time that a positive pulse exists on both grids, plate current flows and gives an output voltage as shown.

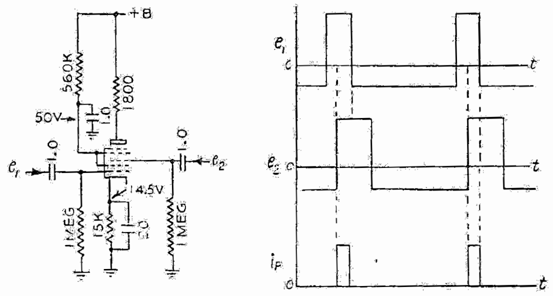

Figure 3-29 - Non-linear Mixer

Figure 3-30 is a curve taken on a type 6L7 tube for the electrode voltages shown. If the bias voltage is set at -14.5 volts as indicated, either grid potential may be reduced to zero without plate current flow.

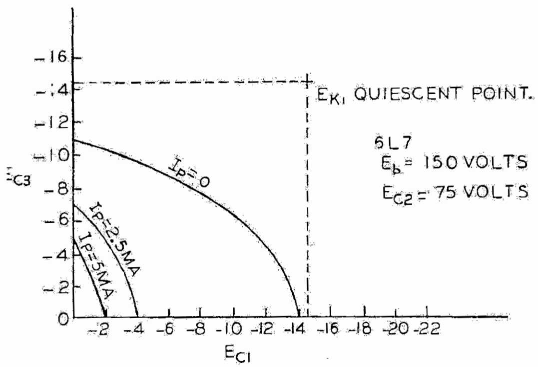

Figure 3-30 - Characteristics of Type 6L7 Tube

A-F-C DISCRIMINATOR CIRCUIT

To improve receiver performance, certain limitations have been recommended by the RMA Committee on Standards for the maximum acceleration of the synchronizing-signal frequency. Also, it is desirable to lock the frequency of the sync generator to a local 60-cycle power supply so as to simplify studio and remote operation. Since the local power-supply frequency may change by an amount exceeding RMA standards during sudden load changes, a means of delayed frequency control must be devised, in which the acceleration of frequency, in cycles per second per second, does not exceed the recommended standard. Such a circuit is the lock-in circuit shown in Figure 3-31.

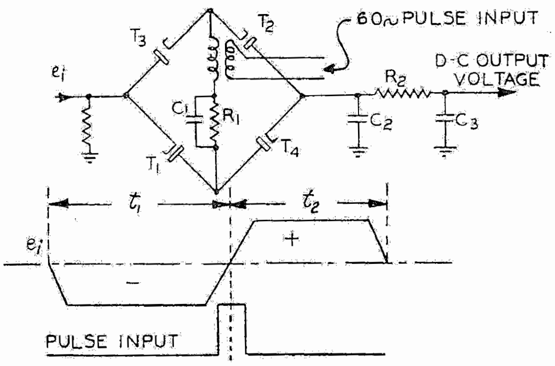

Figure 3-31 - Frequency Control Circuit

The circuit consists of four diode elements in a balanced bridge network. The sine wave of the local 60-cycle supply is clipped and applied across the bridge. A 60-cycle pulse voltage, derived from the synchronizing signal oscillator, is applied to the center leg of the bridge through a transformer. The phase of the local power supply voltage is adjusted so that the pulses occur at the zero-voltage point.

With reference to Figure 3-31 it may be seen that, normally, T_1 and T_2 would conduct during the negative half-cycle; however, a bias voltage, built up across R_1C_1, prevents conduction. Similarly, T_3 and T_4 would normally conduct during the positive half-cycle, except for the bias voltage. The pulse voltage overcomes the bias voltage when the clipped sine wave is passing through zero, and T_1 and T_2 conduct briefly during a small portion of the negative half-cycle, while T_3 and T_4 conduct momentarily during a small portion of the positive half-cycle. If the pulse voltage is in phase with the power line voltage and of the same frequency, the net charge on C_2 will remain the same.

If, however, the frequency relationship between C_1 and the pulse voltage should change, the charge on C_2 will change because the diodes will conduct more during one of the half-cycles than during the other. The time constant of R_2C_3 may be adjusted to provide for slow changes in the d-c output voltage, thus preventing erratic changes in power-line frequency from appearing in the control voltage.

The d-c output voltage is used to control a reactance tube for changing the frequency of the pulse voltage.

REACTANCE TUBE CIRCUIT

A reactance tube is used for controlling the frequency of an oscillator by varying the effective tank circuit in the plate of the oscillator. Figure 3-32 shows a typical reactance tube circuit.

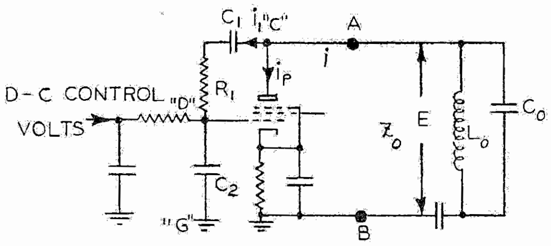

Figure 3-32 - Reactance Tube Circuit

In this circuit, $R_1 >> 1/j\omega C_1$. Let the impedance between points *"C"* and *"D"* be $Z_1 = R_1$. The impedance from *"D"* to *"G"* is $1/j\omega C_2$, which we shall call Z_2. Then we may draw the equivalent circuit shown in Figure 3-33.

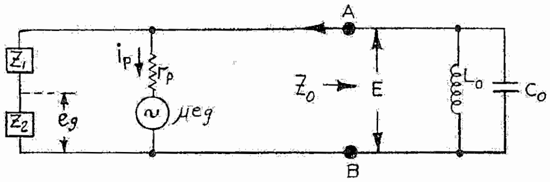

Figure 3-33 - Reactance Tube Equivalent Circuit

If $\quad Z_1 >> Z_2, \; e_g = \dfrac{E\,Z_2}{Z_1 + Z_2} \approx \dfrac{E\,Z_2}{Z_1}$

If $\quad Z_1 + Z_2 >> r_p, \; i_p \approx i$

but $\quad i_p = g_m e_g = \dfrac{g_m Z_2 E}{Z_1}$

The admittance, looking into the reactance tube plate is

$$Y_{AB} = \frac{i}{E} \approx \frac{g_m Z_2}{Z_1}$$

or

$$Y_{AB} = \frac{g_m}{j\omega C_2 R_1}$$

Thus, the admittance of the reactance is equivalent to an inductance which would vary with g_m. The vector diagram of current and voltage relationships is shown in Figure 3-34.

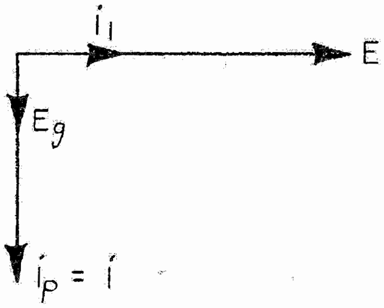

Figure 3-34 - Reactance Tube Circuit Vector Diagram

Mutual conductance of the reactance tube is varied by changing the d-c grid bias. This circuit, in conjunction with the lock-in circuit previously described, may be used to keep the frequency of an oscillator synchronized to a local power source.

SAWTOOTH GENERATORS

A sawtooth generator is a device whose output voltage has a repeating triangular wave shape of which the positive slope is constant. Thus

$$\frac{de}{dt} = Constant$$

This type of voltage is used as a time base for the scanning of cathode-ray or kinescope tubes. In view of the present television standards, we shall be concerned with sawtooth wave forms whose frequencies are 60 cycles and 15,750 cycles per second.

Figure 3-35 is a circuit diagram of a sawtooth generator commonly used in television equipment.

Assume that C is charged at the beginning of a cycle. Pulse e_i is applied to the grid

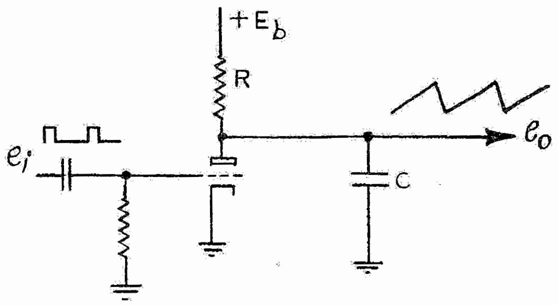

Figure 3-35 - Sawtooth Generator Circuit

with sufficient amplitude to drive the grid positive. The triode conducts heavily, discharging C. During the positive pulse interval, the flow of grid current produces a bias voltage across the grid resistor, of sufficient amplitude to cut the tube off when the pulse goes negative. Capacitor C charges exponentially through resistor R while the tube is cut off between pulses.

In the analysis of the sawtooth generator, certain assumptions will be made. First, we shall assume complete discharge of C during the pulse. Usually $r_p << R$ and t_1 is long enough to permit the voltage across C to discharge to $E_b r_p/(r_p + R)$. With $r_p << R$, we have $e_1 \approx 0$.

Now we shall define a linearity factor λ. Consider Figure 3-36 in which we have a linearly increasing voltage of constant slope de/dt. Such a voltage may be obtained by making the charging current constant, or

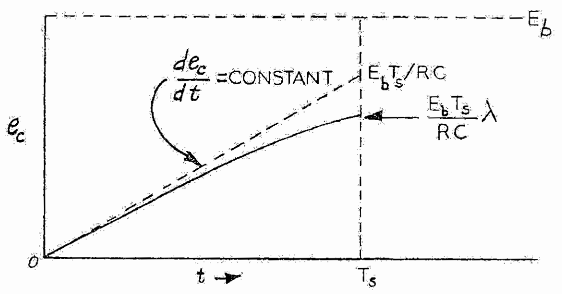

Figure 3-36 - Analysis of Sawtooth Voltage

$$e_c = \frac{1}{c}\int_0^t i\,dt$$

$$= \frac{i}{c}\,t$$

$$= \frac{E_b\,t}{RC}$$

In the ordinary sawtooth generator circuit, the charging current is not constant, but varies exponentially so that the voltage on C at the time T_s is less than the voltage for the ideal case by a factor λ, or

$$e_c = \frac{E_b\,T_s}{RC}\,\lambda$$

The linearity factor λ is thus expressed as the percentage of ideal voltage to which C charges in a simple circuit.

From the simple circuit we know that

$$e_c = E_b(1 - \epsilon^{-T_s/RC});$$

hence

$$\lambda = (1 - \epsilon^{-T_s/RC})\,\frac{RC}{T_s}$$

If we expand the exponential term about zero by means of a McLaurin's series we obtain

$$\lambda = [1 - (1 - \frac{T_s}{RC}) + (\frac{T_s}{RC})^2\,\frac{1}{2} - (\frac{T_s}{RC})^3\,\frac{1}{6} + (\frac{T_s}{RC})^4\,\frac{1}{24} + \ldots]\,\frac{RC}{T_s}$$

$$= 1 - (\frac{T_s}{RC})\,\frac{1}{2} + (\frac{T_s}{RC})^2\,\frac{1}{6} - (\frac{T_s}{RC})^3\,\frac{1}{24} + \ldots]$$

Let us take the first two terms and rearrange

$$\frac{2}{\lambda} - 2 = \frac{T_s}{RC}\,\frac{1}{\lambda}$$

If we restrict the value of λ to the limits 0.75 to 1, we may write

$$\frac{T_s}{RC} \approx \frac{2}{\lambda} - 2$$

Substituting, we obtain

$$e_c \approx E_b \left(\frac{2}{\lambda} - 2\right) \lambda$$

$$\approx 2 E_b (1 - \lambda)$$

The preceding equations are useful for determining the charging time constant and output voltage for a given supply voltage and linearity requirement. The linearity factor usually varies from 0.90 to 0.95.

BLOCKING OSCILLATOR SAWTOOTH GENERATOR

The sawtooth generator described above requires a pulse driving signal of fairly good rectangular wave shape. If the driving pulse fails, no sawtooth output is obtained. A blocking oscillator can be used as a sawtooth generator to provide output voltage even though the synchronizing source may fail. Figure 3-37 is a circuit diagram of such a generator

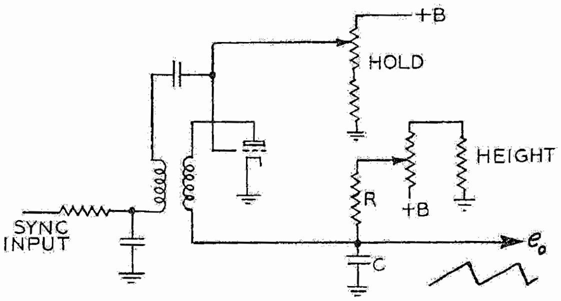

Figure 3-37 - Blocking Oscillator Sawtooth Generator

In this circuit, *R* and *C* form the sawtooth through the charging and discharging action of the tube. Assume that *C* is charging through *R*. Then the blocking oscillator conducts heavily, discharging *C*. When the grid is driven below cut-off, the tube ceases conduction, and *C* charges through *R*. The blocking oscillator is synchronized by a pulse signal whose wave form need not be rectangular. Frequency is adjusted to the synchronizing signal by the "hold" control. Amplitude of the sawtooth is adjusted by the "height" control.

LINEARITY

The linearity of the output voltage from the conventional sawtooth generators described above varies with the time constants used; and the voltage always is an exponential, instead of a linear, function of time. Special methods may be applied to improve the linearity of the sawtooth. One means by which linearity may be corrected is shown in Figure 3-38.

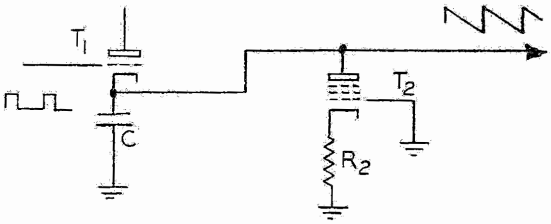

Figure 3-38 - Sawtooth Linearity Circuit

In this circuit *C* is charged during the input pulse and discharged through a constant-current pentode T_2. Since the discharge current is very nearly constant, the voltage on the capacitor becomes.

$$e_c = -\frac{1}{C}\int_0^t i\,dt$$

$$= \frac{Kt}{C}$$

The plate resistance of the pentode can be increased by using a large cathode resistor R_2, thus increasing the effective plate resistance by $1/(1 - g_m R_2)$

It will be noted that the output is inverted from the conventional sawtooth generator.

By the use of feedback to the pentode of Figure 3-38, a perfect sawtooth may be obtained. In Figure 3-39 a portion of the sawtooth output

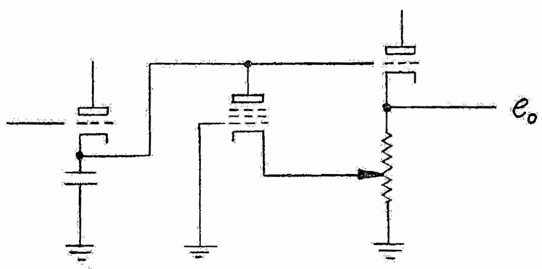

Figure 3-39 - Linearity Feedback Circuit is fed back to the cathode of the constant-current pentode. The effective plate resistance is high as C begins to discharge, and decreases as the discharge proceeds. Not only may a linear sawtooth be obtained, but a strong overcorrection may be attained.

These, and other methods of linearity correction, are described in the December 1946 issue of "Electronics" in the paper "Linear Sweep Circuits" by Robert P. Owen.

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Puckle, O.S., *Time Bases*, John Wiley and Sons, Inc., New York, 1943

Owen, Robert P., "Linear Sweep Circuits," *Electronics*, December, 1946

Brainerd, J.G., *Ultra-high Frequency Techniques*, D. Van Nostrand Company, Inc., New York, 1942.

MAGNETIC DEFLECTION

Deflection of the electron beam in kinescope and camera tubes is accomplished by a uniform magnetic field at right angles to the tube axis. When it travels through the magnetic field, the electron is subjected to a transverse force which causes it to move along an arc of a circle. On leaving the magnetic field, the electron continues along a straight line which is tangent to the arc at the field boundary, as shown in Figure 3-40. The electron emerges from

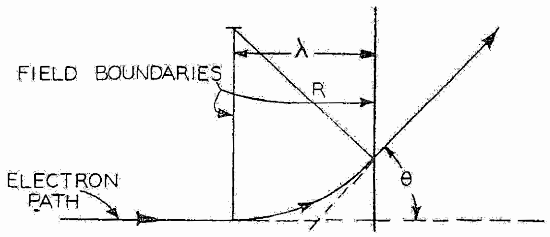

Figure 3-40 - Electron Path at Field Boundary

the field at an angle θ with respect to the original direction of motion. The total angle of deflection is 2θ. In present-day kinescope tubes, the maximum angle of deflection is $50°$ and is limited by inside neck diameter and length of field λ.

The magnetic field required for deflecting the electron beam in a television kinescope or pick-up tube is produced by passing a sawtooth current through a pair of series-connected coils on opposite sides of the tube neck. Formerly, the coils which make up the yoke were wound on a flat rectangular template, and then formed around a cylinder of a diameter equal to, or greater than, the tube neck. Present coils are machine-wound, and the cylindrical forming occurs during the winding process. Figure 3-41 is a rough sketch of a modern yoke winding.

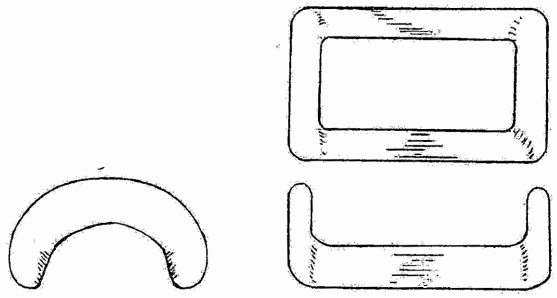

Figure 3-41 - Deflection Yoke Winding

The number of ampere-turns required to produce a given angle of deflection is calculated from

$$NI = \frac{2.68\, l_o \sin\theta \sqrt{E_a}}{\lambda}$$

where
NI = ampere-turns of winding
l_o = length of air gap, inches
λ = length of magnetic field, inches
θ = 1/2 total deflection angle
E_a = accelerating potential, volts.

Note that the above value NI is for half the total deflection angle. To obtain NI for the total deflection angle, multiply by 2.

For a standard 4:3 aspect-ratio television raster, the value of the horizontal-winding ampere-turns is

$$(NI)_H = 0.8\, NI$$

while for the vertical winding it is

$$(NI)_V = 0.6\, NI.$$

VERTICAL DEFLECTION CIRCUIT

Figure 3-42 shows a vertical deflection circuit in its simplest form. The vertical yoke

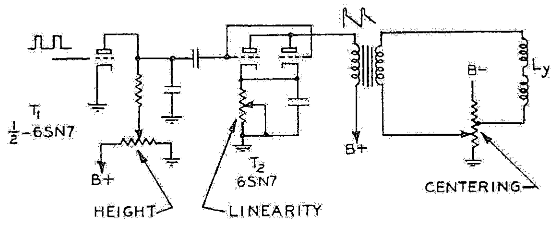

Figure 3-42 - Vertical Deflection Circuit

winding is transformer-coupled to a 6SN7 triode, T_2, with both sections parallel-connected. The driving sawtooth is generated in a conventional sawtooth generator, T_1.

Practical values for the vertical winding of the yoke are $L = 43$ millihenrys and $R = 70$ ohms. At the vertical scanning frequency the load impedance becomes

$$Z_y = 70 + j18.1$$

In the design of the transformer and driving circuit, the inductive component of the load is neglected. The problem then becomes one of designing a transformer which will match the yoke resistance to the driver tube and present sufficient primary inductance for good low-frequency response. A type 6SN7 triode provides sufficient output to deflect a 9-kv beam. The plate resistance of the 6SN7, parallel-connected, is approximately 3500 ohms. For maximum power output, the load should be $2r_p$; therefore the reflected load of the yoke should appear as 7000 ohms on the primary side. The transformer turns-ratio becomes

$$N_p/N_s = \sqrt{Z_p/Z_s} = \sqrt{7000/70} = 10/1.$$

Good low-frequency response is obtained by making the primary inductance large. In the RADIOTRON DESIGNER'S HANDBOOK the ratio of low-frequency gain to mid-band gain is given as

$$A_r = \frac{1}{\sqrt{1 + (r_p/L_p)^2}}$$

If the response at 60 cycles is to be 1 db down, A_r becomes 0.89, from which

$$L_p/r_p = 1.94.$$

For the circuit of Figure 3-42 the primary inductance should be 18 henrys. Actually, for standard vertical-deflection transformers, L_p varies from 40 to 60 henrys.

Some control of linearity may be obtained by varying the bias voltage of T_2. Usually, the sawtooth amplitude and bias are adjusted together to place the operating point in the most linear portion of the tube curves.

The picture is centered by adjusting the centering potentiometer so that a steady d-c current flows in the yoke. Current may be caused to flow in either direction to move the picture in either direction.

No external damping of the yoke winding is required, in the majority of cases, since the plate resistance of the tube is reflected to the transformer secondary. If external damping is required, a resistor of proper value may be placed across the yoke winding.

AUTOMATIC LINEARITY CONTROL

Picture linearity may be corrected by the linearity correction devices previously discussed. Additional tubes and circuit components are required, however, and if the expense is justified, an automatic control may be used.

Figure 3-43 shows an automatic linearity control circuit recently developed by the Advanced Development Section, Home Instruments Department.

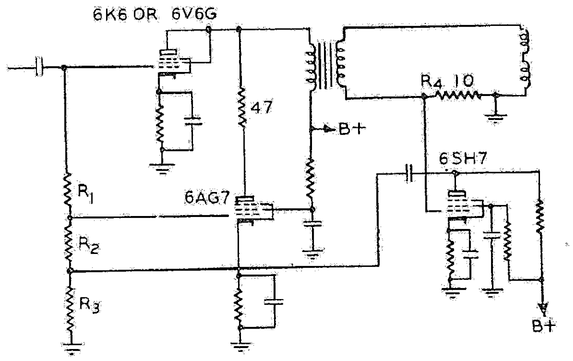

Figure 3-43 - Automatic Linearity Control Circuit

The circuit operates as follows. A pilot voltage is developed across R_4 in the yoke circuit, which is proportional to the current in the yoke. This voltage contains the distortion of the current sawtooth and is shown in Figure 3-44(a).

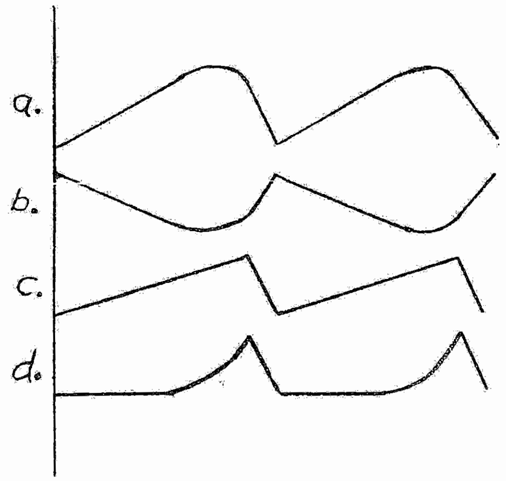

Figure 3-44 - Sawtooth Linearity Control
a. *Current Sawtooth Containing Distortion*
b. *Plate 6SH7 Sawtooth Plus Distortion*
c. *Input Sawtooth*
d. *Grid 6AG7 Distortion Only.*

The distorted sawtooth is amplified in a high-gain pentode and fed to the top of R_3. A sawtooth of good linearity is fed into the driver tube, 6K6 or 6V6, and also to R_1. In the resistance network R_1R_2 the linear sawtooth is compared to the distorted sawtooth, and the existing distortion is placed on the grid of the 6AG7. The distortion signal causes the plate of the 6AG7 to draw a current which cancels the original distortion.

In this system, picture size may be changed over wide limits with negligible vertical distortion. The values of R_1 and R_2 should be less than one-half megohm to prevent integration of the linear sawtooth.

HORIZONTAL DEFLECTION CIRCUITS

Circuit design for magnetic deflection of the electron beam at horizontal-line frequencies requires a different approach than for vertical deflection. At 15,750 cycles per second, the yoke presents a load which is almost entirely reactive. Unless means are devised to recover a portion of the power fed into the yoke during trace time, a relatively high amount of power must be expended in deflecting the beam. An ideal cyclic system such as that discussed by Otto Schade in RCA REVIEW for September, 1947, requires wattless power. This system will form the basis for study of the horizontal-deflection problem.

Consider the simple circuit of Figure 3-45. The yoke is represented by L_y, C_y, R_y. Suppose that switches S_1 and S_2 are open at the time $t = 0$. At the beginning of the deflection cycle, S_1 is closed, applying voltage E across L_y, R_y, and C_y. If it were not for R_y, the current through L_y would increase linearly with time, as expressed by the relation

$$\frac{di}{dt} = \frac{E}{L}.$$

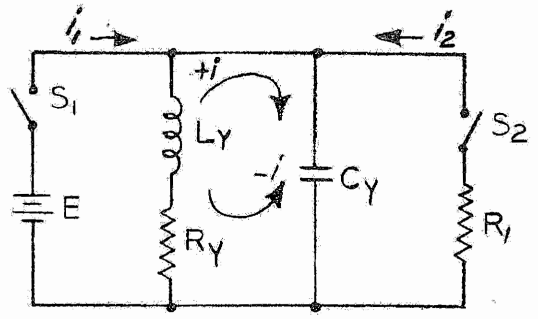

Figure 3-45 - Equivalent Horizontal Deflection Circuit

Since R_y is present, the current rises exponentially until switch S_1 is opened. At this point, the beam has been deflected to the right-hand side of the picture. The magnetic field must be reversed quickly in order to return the beam to the left-hand side of the picture, to begin another trace.

Since L_y, C_y, and R_y form a resonant circuit, the fastest means for reversing the field is to permit the winding to oscillate for approximately one-half cycle at its natural resonant frequency.

When S_1 is opened, the magnetic energy stored in the field of L_y is converted into potential energy by the flow of $+i$ into C_y, and back into magnetic energy by the flow of $-i$, resulting in an almost complete reversal of the field. Losses in the resonant circuit limit the completeness of reversal to

$$\frac{i_1}{i_2} \approx \epsilon^{-\pi/2Q}$$

Figure 3-46 indicates the current and voltage waveshapes in the yoke for a complete deflection cycle.

When the current in the yoke has reached the value $\hat{i}_2$ in the negative direction, switch S_2 is closed, which places damping resistor R_1 across

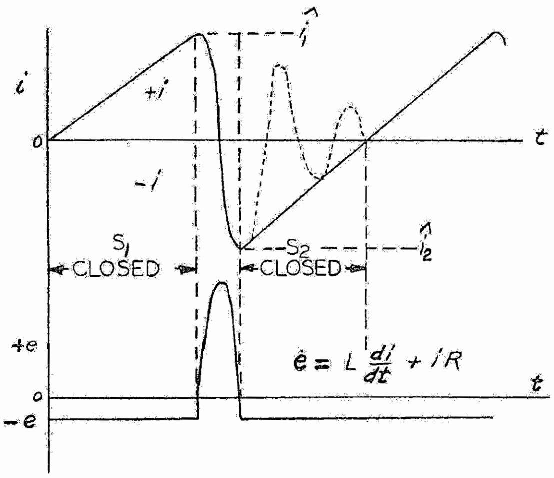

Figure 3-46 - Deflection-cycle Waveshapes

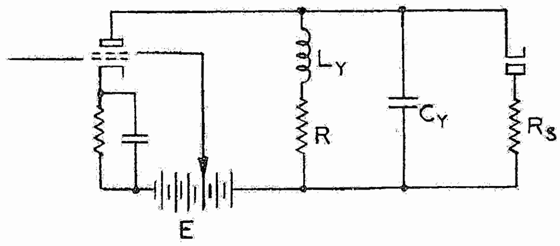

Figure 3-47 - Electronic Switch and Generator Circuit

the oscillating circuit. If S_2 were not closed, the yoke would continue to oscillate, as shown by the dotted lines in Figure 3-46. Closing of S_2 causes the oscillatory circuit to be slightly overdamped, so that $-i$ decays exponentially. When $-i$ reaches zero, S_1 is closed again to begin another cycle.

Because of the presence of the iR drop in the inductance, the resultant current wave form in the yoke is exponential instead of being linear with time, as desired. If the iR drop can be canceled, the total voltage E may be applied to L_y, resulting in a linear current in the yoke. Suppose we insert a generator in series with E, whose characteristic is

$$\frac{\Delta e}{\Delta i} = -R .$$

Then a linear rise of current in L_y may be obtained. Reference to the plate-family of curves for a vacuum tube reveals that a tube may serve as such a generator and as an electronic switch to replace S_1. Also, we may use a vacuum tube to replace S_2 and add $-R$ for the oscillatory phase. Such a circuit is shown in Figure 3-47.

The operation of the tube may be plotted from its family of curves. Refer to Figure 3-43. The load line $-R$ is so drawn that it intersects the plate-voltage, or zero-current, axis at the point $E - L\ di/dt$. A plot of current-versus-time is obtained from the intersection of the $-R$ line with the grid-voltage lines. The grid-voltage waveshape e_g is obtained for the tube by plotting E_{c1} against time for corresponding values of current i.

The diode characteristic is plotted in a similar manner. The voltage causing diode conduction, however, becomes $L\ di/dt$ and is equal to the drop across the inductance during trace time. The load line for the diode must be drawn for $r_d + R_s$, where r_d is the diode resistance. Schade states that linearity in the diode circuit occurs when

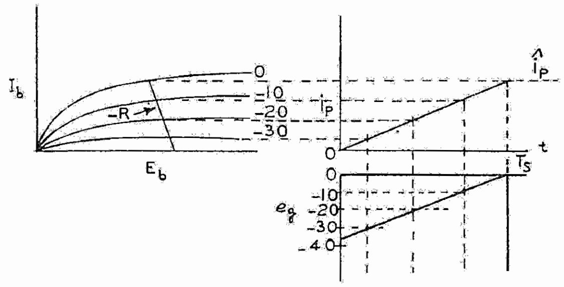

Figure 3-48 - Tube Operating Curves

$$R_s = [E - (\hat{i}_1 R + E_d)]/\hat{i}_2 .$$

The circuit operation may be improved by replacing the diode with a controlled triode. For simplification, a transformer is added, and the circuit becomes the one shown in Figure 3-49.

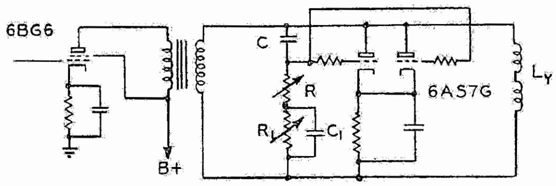

Figure 3-49 - Improved Switching Circuit

The combined characteristics of the beam tetrode and the triode are shown in Figure 3-50. Note that the characteristics resemble those of the ordinary push-pull arrangement.

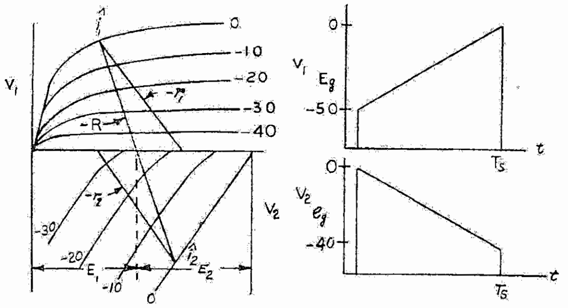

Figure 3-50 - Combined Characteristics of Beam Tetrode and Triode

In the ideal case, in which there are no losses, the 6BG6 driver tube supplies half the deflection current, and the 6AS7 triode damper supplies the remainder from the stored energy. Such utilization of current is shown in Figure 3-51. Because of losses in the actual circuit, the driver tube must supply about 60% of the total deflection current.

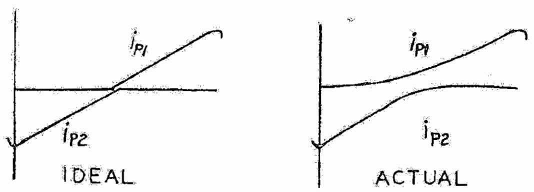

Figure 3-51 - Deflection Circuit Operation

The control-grid voltage for the triode damper is generated by differentiation of the pulse voltage across the yoke. The values of RC are determined by the equation

$$\frac{T_S}{R_C} \approx \frac{2}{\lambda} - 2$$

where λ = linearity of voltage rise = 0.3 to 0.8.

Usually R is made variable for adjusting linearity.

The combination R_1C_1 has a long time-constant and is placed in the grid circuit for establishing grid bias for the triode by the flow of grid current on the peaks of the grid voltage.

REFERENCES

Maloff, I. G., and Epstein, D. W., *Electron Optics in Television,* McGraw-Hill Book Co., Inc., New York, 1938.

Schade, O. H., *"Magnetic Deflection Circuits for Cathode-Ray Tubes,"* RCA Review, September, 1947.

VIDEO AMPLIFIERS

REQUIREMENTS - The nature of the picture signal imposes certain requirements upon the video amplifier, which must be met if fine picture detail is to be resolved. First, the bandwidth must satisfy the relation

$$f_n = A_r n \frac{10^6}{H_a \times 2}$$

where f_n = fundamental frequency for n lines
A_r = aspect ratio = 4/3
n = number of lines to be resolved
H_a = active trace time, microseconds.

Since horizontal blanking occupies 16% of the horizontal period, the active trace time, H_a, is 0.84 × 63.5 = 53.3 microseconds. To resolve 400 lines, the bandwidth must be

$$f_n = \frac{4}{3} \times \frac{400 \times 10^6}{53.3 \times 2} = 5 \text{ megacycles}$$

In practice, the output of the video transmitter is specified by standards to include all frequencies between 30 cps and 4 megacycles per second. Hence, the video amplifier must amplify, without discrimination, at least those frequencies between 30 cps and 4 megacycles per second. Usually, the video amplifier is designed with a bandwidth exceeding these limits.

Also, the video amplifier must have a minimum time-delay discrimination. This requirement is fulfilled when the phase angle between input and output voltages is proportional to frequency.

Finally, there are requirements for the video amplifier which are set by standards or practice, some of which are output-voltage levels, terminal impedances, permissible signal-to-noise ratio, etc.

FREQUENCY RESPONSE - Low-frequency response of an RC-coupled amplifier is determined by the time-constant of the coupling capacitor and grid-leak resistor. In practice, good low-frequency response is obtained by making the time-constant large or by using clamp circuits.

High-frequency response is limited by shunt capacity across the plate-load resistor. This shunt capacity includes the tube input and output capacity, wiring capacity, and stray capacity of circuit components. Good high-frequency response is obtained by utilizing the various shunt capacities as elements of a low-pass coupling filter.

Figure 3-52 is a diagram of a constant-K low-pass filter consisting of one full section and one half-section terminated in its character-

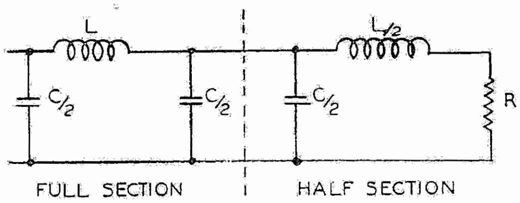

Figure 3-52 - Constant-K Low-pass Filter

istic impedance. The terminating half-section is added for impedance matching purposes. When the low-pass filter is properly designed and terminated, the characteristic impedance is constant to almost the cut-off frequency. Connection of the coupling filter to the amplifier tubes is shown in Figure 3-53. In Figure 3-53

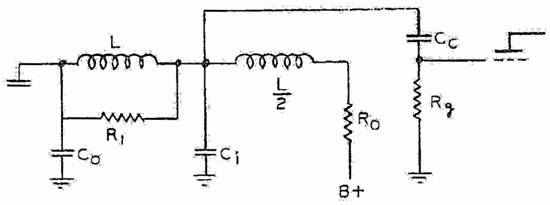

Figure 3-53 - Coupling Filter-connection

C_o and C_i are output and input capacities of the tubes. For this particular type of low-pass filter, the following equations apply:

$$f_c = \text{cut-off frequency}$$

$$R_o = \frac{1}{\pi f_c C_i}$$

$$L = R_0^2 \; C_i$$

$$C_o = \frac{C_i}{2}$$

$$R_1 = (5 \text{ to } 10)\; R_o$$

$$A = g_m R_o$$

The resistor R_1 is added to lower the Q-factor of the series inductance. A peak in the response curve will result prior to cut-off if the Q-factor is not optimum.

The characteristics of the constant-K low-pass filter depend upon the components being pure inductances and pure capacitances. It has been shown that distributed coil capacity converts the constant-K type into an M-derived filter, as shown in Figure 3-54.

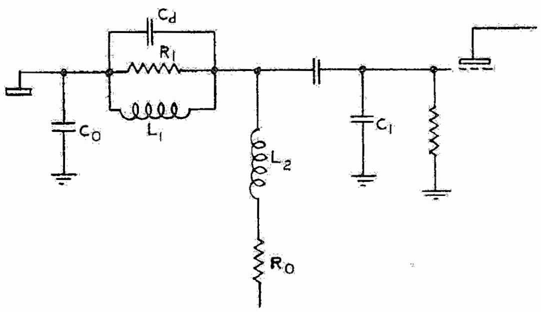

Figure 3-54 - Effect of Coil Capacity on Constant-K Filter

Equations for the M-derived filter are as follows:

$$M = \sqrt{1 + \left(\frac{C_d}{C_o}\right)^2} - \frac{C_d}{C_o}$$

$$R_o = \frac{M}{\pi f_c C_i}$$

$$L_1 = R_0^2 \; C_i$$

$$L_2 = 0.8\; L_1 \text{ approx.}$$

$$R_1 = (5 - 10)\ R_o$$

$$A = g_m R_o$$

Practical video amplifiers use the M-derived low-pass filter as a means of coupling amplifier stages.

Diagnosis curves are given in Figures 3-55 and 3-56 to aid in the alignment of video amplifiers using low-pass filter coupling.

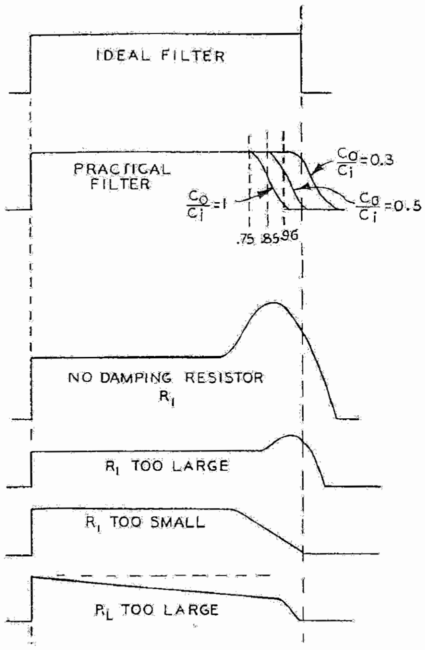

Figure 3-55 - Diagnosis Sheet #1

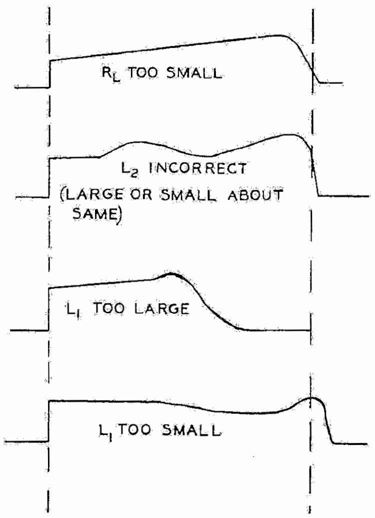

Figure 3-56 - Diagnosis Sheet #2

CLAMP CIRCUITS

The clamp circuit is often used as a "D-C Restorer"; however, it can also be used to restore low frequencies in a video amplifier. Its operation in the latter application will be described first.

Consider the video signal for a half-black, half-white picture applied to the input of a video amplifier whose frequency response is very poor below the horizontal-scanning frequency. Assume that horizontal blanking pulses are also introduced at the amplifier input, and that they are of greater amplitude than any other part of the video signal. A sketch of the picture and the corresponding video signal with horizontal blanking is shown in Figure 3-57.

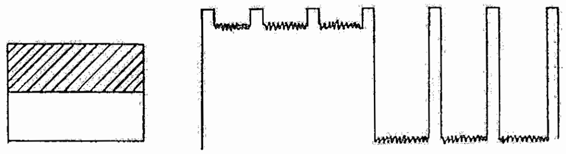

Figure 3-57 - Half-black, Half-white Picture and Video Signal

After this signal has passed through the amplifier, the low-frequency components will be missing, and the signal will distribute itself about an a-c axis as shown in Figure 3-58. Low-frequency components are, in this case, considered to be any components of less than the horizontal scanning frequency. The transitory periods immediately following the change from black to white are not shown in Figure 3-58.

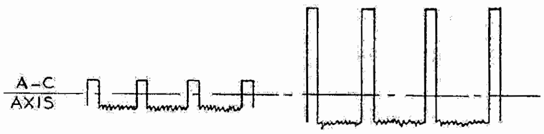

Figure 3-58 - Video Amplifier Output Signal

Note that, if we could bring the peaks of the horizontal blanking pulses that occur during the "black" portion of the picture to the same level as the peaks of those that occur during the "white" portion, the signal would again be identical to that in Figure 3-60. In other words, the low-frequency components would then be restored because the output signal would be similar in shape to the input signal.

Figure 3-59 shows a hypothetical circuit for bringing all of the blanking pulses to the same level. The time-constant of R and C must be sufficiently small so that C is discharged before the blanking pulse is over. The switch is so controlled that it closes at the start of the blanking pulse and opens *before* the end of the pulse. Because of these conditions (grid of the tube floating during the time the switch is open, grid-side of C being always brought to ground potential during the pulse, and switch opening before the pulse is over), *the remaining portion of the pulse always falls at the same point on the tube's operating characteristic.* As explained in the preceding paragraph, this is equivalent to restoring the low-frequency components.

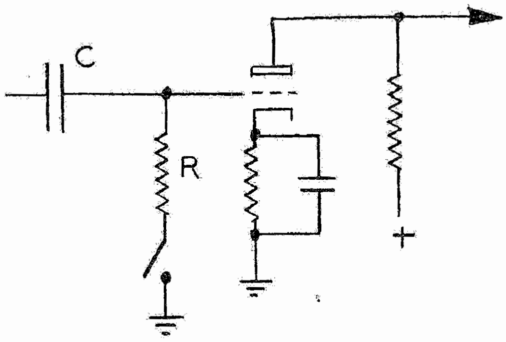

Figure 3-59 - Simplified Clamp Circuit

A clamp circuit, shown in Figure 3-60, is electrically equivalent to the arrangement of Figure 3-59. The diodes replace the switch of Figure 3-59, and the switch control is supplied by the diode keying pulses. The circuit "clamps" on the periodic pulses in the video signal, which in the given example are horizontal blanking pulses; hence its name.

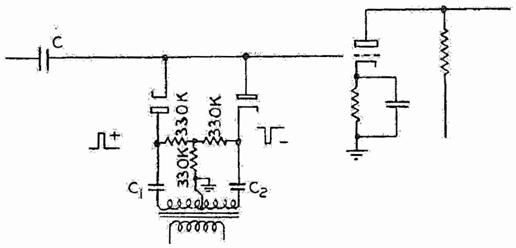

Figure 3-60 - Clamp Circuit

The keying pulses (which should not be confused with the clamp pulses) are 180 degrees out of phase, so that both diodes become conducting simultaneously. By using this balanced arrangement, the keying pulses cancel out at the grid of the amplifier tube, and are therefore not added to the desired signal. It is essential that the keying pulses end before the clamp pulses, as explained above. Horizontal synchronizing signal makes ideal keying pulses.

The amplitude of the keying pulses must also be greater than the clamp pulses, so that the diodes can be made conducting during keying time. Practice has shown that the keying pulses should be one-and-a-half to two times as large as the clamp pulses.

It is important to note that pulses other than horizontal blanking can be used to clamp on. The only requirements are that they be greater in amplitude than any other part of the video signal, that their peaks represent constant amplitude in the input signal, and that their frequency be sufficiently high for the amplifier to pass them without frequency or phase distortion. Of course, they must not interfere with the desired signal. Hence, for television work, they must occur at horizontal scanning frequency, and during horizontal blanking time. Their polarity, with respect to the video signal, is unimportant.

The coupling capacitor C and the resistance between grid and ground during keying time must have a sufficiently small time-constant for C to discharge during keying time. For 15-kc clamp pulses, the value of C can be between 100 and 500 $\mu\mu f$. The coupling capacitors, C_1, and C_2, are also somewhat critical because the diodes are self-biased by them and their associated "leak" resistors. The bias developed is proportional to the amplitude of the keying pulses, in a manner similar to that of a conventional diode detector. Values for C_1 and C_2 are best determined by experiment. Values which have been used in the past lie between 0.003 μf and 0.1 μf for the leak resistances shown.

In some cases it may be desirable to return the grid of the amplifier tube to a fixed-bias source instead of to ground. The bias source is then introduced in series with the ground lead shown in Figure 3-60.

The source of the keying pulses is of importance. A center-tapped transformer is desirable because it provides balanced pulses easily, and the source-impedance is low. A tube with load resistors in both plate and cathode circuits (cathode-follower type of phase inverter) can be used to provide keying pulses. This tube should preferably operate with a negative-polarity pulse on its grid, so that the tube is cut off during keying time. Otherwise, the source-impedance will be different for the positive and negative output pulses, due to cathode-follower action. An unbalance in the source-impedance may adversely affect the operation of the clamp circuit.

The clamp circuit can be modified to advantage when only single-polarity keying pulses are available. This is shown in Figure 3-61. In this circuit, a single keying pulse makes both diodes conducting because they are in series, as far as the keying pulse is concerned. The disadvantage of this circuit is that a small amount of the keying pulse is super-imposed on the video signal because of the unbalance. When horizontal sync pulses are used as keying pulses, this circuit will add a small amount of sync to the video signal; this will some times be an advantage rather than a disadvantage. If vertical sync is unavoidably present along with the horizontal, the coupling capacitors C_1 and C_2 should be increased to 0.5 μf.

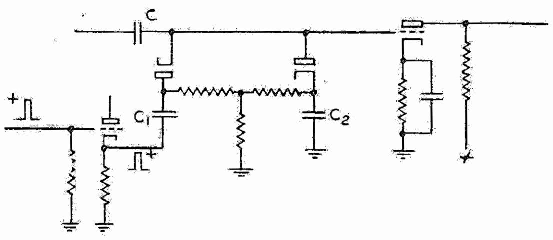

Figure 3-61 - Clamp Circuit for Single-polarity Keying Pulses

Another version of the clamp circuit is shown in Figure 3-62. Only single-polarity keying pulses are required. The source-impedance of the keying pulses can be high, but the circuit provides a low-impedance path between grid and ground during keying time.

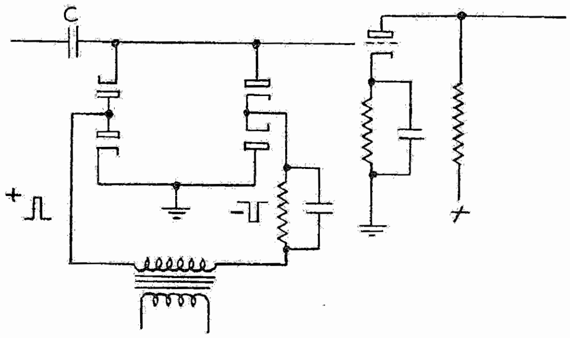

Figure 3-62 - Clamp Circuit for Single-polarity Keying Pulses

THE CLAMP CIRCUIT AS A D-C RESTORER

Since, as was just shown, the clamp circuit effectively restores low frequencies, the same reasoning can be extended to say that the clamp circuit will also restore the d-c component of the video signal. Without d-c restoration, the a-c axis of any signal will pass through the operating point of the tube characteristic to which that signal is applied. However, when the clamp circuit is used on the grid of an amplifier tube (or kinescope), the clamp pulses in the signal are always referred to the same point on the characteristic, regardless of signal amplitude or wave form. In other words, the a-c axis is shifted as the signal amplitude and wave form vary, and a shift in the a-c axis of a wave is equivalent to adding a d-c component. The clamp circuit has the advantage over the simple, single-diode type of d-c restorer in that it responds very quickly to signal changes, whether they be increasing or decreasing; whereas the simpler type has appreciable time lag when the signal suddenly decreases.

PICK-UP TUBES

TYPES - Two types of pick-up tubes are in general use today, namely, the iconoscope and the image orthicon. The iconoscope dates back to about 1923, when it was developed by Dr. V. K. Zworykin. It is still being used for motion-picture pick-up. The image orthicon has replaced the iconoscope for live-talent pick-up. Image-orthicon development was hastened by war-time requirements, and progress on the stabilization and improvement of this tube has been rapid.

ICONOSCOPE - The iconoscope pick-up tube may be used where the scene is illuminated by incident light of approximately 1500 foot-candles. Under ideal lighting conditions, the pictures obtained have excellent resolution and low noise-level. The intensity of illumination, however, limits the use of the iconoscope for outdoor events. When incandescent lighting is used in studios, the problem of removing the heat arises. At present, the iconoscope is used in film cameras only where the motion-picture projector provides ample illumination on the iconoscope mosaic.

The iconoscope contains a photo-sensitive mosaic, a collector ring, and an electron gun. A sketch of the tube is shown in Figure 3-63. The electron gun is set at an angle with the mosaic in order to clear the front of the tube, so that an optical image may be focused on the mosaic.

A uniform mica plate, 0.001 inch thick, is the basic structure upon which the mosaic is constructed. A fine coating of silver oxide is sifted upon the mica. Then, the structure is baked in an oven. The heat produces pure silver from the silver oxide. The pure silver congeals into thousands of small droplets. Then the mica plate is placed in the presence of cesium vapor and oxygen, and a glow discharge is passed through the tube. Silver oxide, cesium oxide, and pure cesium are formed. By this process, small photo-sensitive islands are formed on the mica. The mosaic is completed by coating the back side of the mica with colloidal graphite to form the signal plate which is capacity-coupled to the photo-sensitive surface. Better color response is obtained by the process of silver sensitizing, in which a small particle of pure silver is heated in a filament while the tube is on the pumps. Silver vapor settles on the photo-sensitive islands and gives the mosaic better response toward the blue end of the visible spectrum.

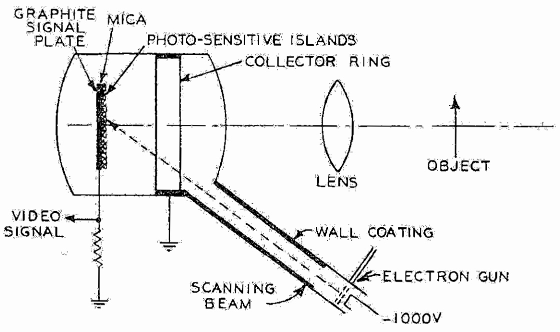

Figure 3-63 - Iconoscope Pick-up Tube

The iconoscope is a storage-type device in which the varying illumination of an optical image on the mosaic causes emission from the photo-sensitive islands. The charge on each picture element represented by the photo-sensitive element remains constant until released by the scanning beam. The operation of the iconoscope is best understood by considering first the action resulting from scanning the mosaic in darkness, *i.e.*, with no optical image or light on the mosaic.

With the collector ring grounded and the cathode potential fixed at -1000 volts, the beam acquires a kinetic energy of 1000 electron volts by the time it reaches the mosaic. On striking the mosaic, the beam causes secondary emission of electrons from the photosensitive islands, *i.e.*, each beam electron knocks several secondary electrons off the photosensitive island. The ratio of secondaries to beam electrons is 6:1 under dark conditions. The secondary electrons, for the most part, rain back on the mosaic. Enough secondaries travel to the collector ring for the collector-ring current to be equal to the beam current (since the mosaic is completely insulated, the current leaving it must equal the current arriving in the scanning beam)

Figuratively speaking, the scanning beam plows along the mosaic, causing an eruption of secondary electrons from the photo-sensitive surface. The element under the scanning beam charges up to about 2 volts, due to loss of electrons. This value represents the maximum charge which the element can attain by secondary emission and is known as the white level. As the scanning beam moves across the mosaic, part of the electron shower can fall back on the scanned area and reduce the positive charge on picture elements just scanned. At the right-hand edge of the mosaic, however, the scanning beam is turned off for retrace, and no more secondaries are generated to discharge the last part of the trace. Similarly, the beam is cut off at the bottom of the mosaic for vertical retrace; therefore, the last few scanning lines do not receive a proportionate share of the electron rain and remain partially charged. Remember that this action is occurring in complete darkness.

As the electron beam starts scanning the second frame, it encounters elements of the mosaic on the right-hand side and on the bottom that are partially charged, due to the loss of electrons. These elements appear as though they had been exposed to white light. When the beam scans them, fewer secondaries are emitted, and a signal voltage is impressed on the signal plate. This voltage has the waveshape shown in Figure 3-64 vertical and horizontal scans. It is an unwanted signal, that is due to uneven redistribution of secondary electrons, and it is called a shading signal. For eliminating shading signals, equal-amplitude opposite-phase signals are fed

Figure 3-64 - Shading Signal Waveshape

into an amplifier stage following the pick-up tube. The unwanted signal may, fortunately, be effected by a combination of parabolic and saw-tooth signals which can be generated quite easily.

Now we may consider the action of the scanning beam when the mosaic is illuminated by a scene. Bright areas in the scene cause the islands to emit electrons. These electrons travel to the collector ring or redistribute themselves over the mosaic. Suppose a gray tone causes a photo-sensitive island to charge up to +1.5 volts. Then, when the scanning beam comes along, this particular element can only be charged by a differential of 0.5 volt to the white level. On the other hand, a black area leaves the element discharged until it is scanned, at which time the element can charge to the full 2-volt white level. The video signal current in the load resistor is shown in Figure 3-65.

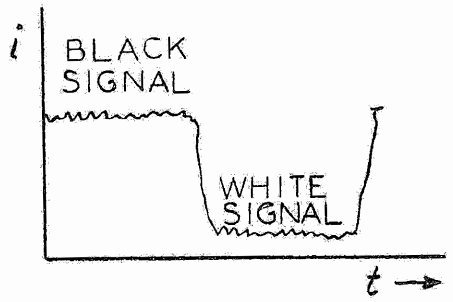

Figure 3-65 - Video Signal for Illuminated Scene

In Figure 3-63 the electron gun for the iconoscope is shown at an angle with the mosaic. This geometrical arrangement produces an effect known as keystoning. For a given angle of deflection of the scanning beam, more of the mosaic top is scanned than the bottom. If no correction were applied to the horizontal scanning generator, the resultant pattern on a monitor would appear as shown in Figure 3-66.

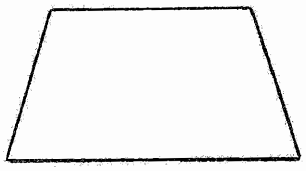

Figure 3-66 - Keystone Pattern

To correct for keystoning, the horizontal scanning generator is modulated by a 60-cycle sawtooth that increases the horizontal scanning current peak-to-peak value linearly at a 60-cycle rate, so that the angle of deflection becomes larger as the beam is deflected vertically.

IMAGE ORTHICON PICK-UP TUBE - The image orthicon is at least 100 times more sensitive than the iconoscope. Also, it is free from the annoying shading and edge-flare effects of the iconoscope. It will deliver a satisfactory picture, without readjustment, when the scene brightness changes by a factor of 100 to 1. A satisfactory picture may be obtained when the incident light on the scene is only 10 foot-candles. The sensitivity of the image orthicon makes it an ideal tube for pick-up of outdoor events.

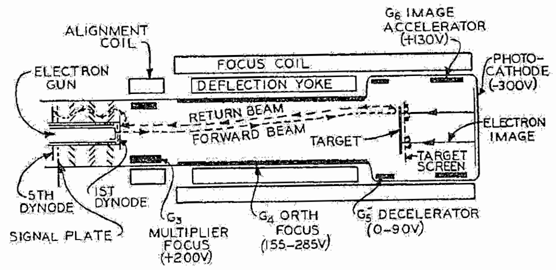

Figure 3-67 - Image Orthicon Pick-up Tube

A sketch of the image orthicon tube construction is shown in Figure 3-67. The tube contains an electron gun with a grid for controlling the current in the scanning beam. The #3 grid, sometimes called the "persuader", causes electrons from the first dynode to go to the second dynode. The #4 grid, which is the coating on the tube wall, together with the magnetic focusing field, focuses the electron beam on the target. The decelerating ring, grid #5, produces an electric field which improves corner focus.

The target is a special glass membrane stretched in a metal ring. The thickness of the glass is approximately 0.0001 inch. On the image side of the target, and at a distance of 0.001 inch, is a mesh screen having 250,000 holes per square inch.

Grid #6 is a ring placed between the target and photocathode. It aids the focusing of electrons from the photocathode on the target.

The photocathode is a transparent layer of cesium in type 2P23 tubes, antimony in type 5769 tubes, and bismuth in type C73150 tubes. The cesium tubes have high infra-red response, while the antimony and bismuth tubes have a more uniform color response in the blue regions.

IMAGE ORTHICON OPERATION- When an optical image is focused on the photocathode, electrons are emitted in proportion to the light and dark areas of the scene. Since the photocathode is at a potential of about -300 volts with respect to the ground and the target screen, the electrons are accelerated toward the target. The action of the focusing coil and the #6 grid focuses the electrons on the target. Thus the optical image is converted into an electron image which bombards the target.

Bombardment of the target causes an emission of the electrons from the glass. Secondary electrons released by the target are collected by the screen. Secondary emission leaves a positive charge pattern on the front of the target, corresponding to the electron image.

Because of the thinness of the glass target, it does not matter, for the electron beam, on which side of the glass the positive charge lies. Upon its arrival near the target rear surface, the beam deposits enough electrons to neutralize the charge. The remainder of the beam turns around and heads toward the rear of the tube. During frame time, the deposited electrons flow through the glass and unite with the positive charge.

The returning electron beam is equal to the electrons emitted by the cathode (nearly a constant number) minus those electrons deposited on the target. The returning beam, therefore, is the original beam modulated by the video signal.

An electron multiplier is located at the rear of the tube. The construction of this multiplier is shown in Figure 3-68.

It is such as to offer an almost opaque surface to the electrons entering from the front. Electrons leaving each dynode, however, find negligible resistance to their travel.

The return beam containing the video information strikes the first dynode, causing secondary emission. The secondary electrons are persuaded to the second dynode by the action of the "persuader," or multiplier focus electrode. As the beam travels from dynode to dynode, the original

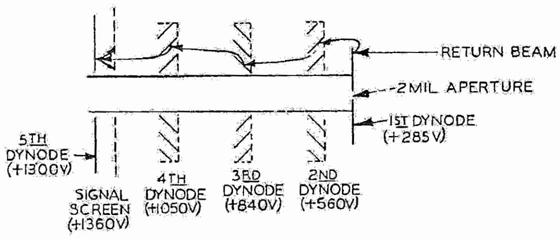

Figure 3-68 - Electron Multiplier for Image Orthicon

return beam is multiplied by secondary emission. The final signal is removed from the signal plate. The overall gain in the electron multiplier can be as high as 2000.

In tube manufacture, the electron gun may become tilted with respect to the tube axis. Electrons emitted from such a gun would enter the focus field with a transverse component of velocity. A force would be developed, which would cause the beam to travel in a radius about the tube axis. The net effect is a spiraling of the beam down the tube. To correct for misalignment of the electron gun, an alignment coil is placed just in front of the gun. It produces a transverse field which cancels the deflection of the beam due to gun tilt.